ANALYSIS OF TISSUE VARIABILITY AND ADAPTIVE TRANSCUTANEOUS POWER

by

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Implantable medical devices comprise a large and growing market, providing both short- and long-term treatment for a variety of medical conditions. This includes continuous therapy devices such as pacemakers and ventricular assist devices, and prostheses such as cochlear implants and retinal prosthesis. Wireless powering enables fully implantable devices that are less limited in size and lifetime by non-rechargeable battery power. Wireless energy transfer also enables communication with the implanted device, allowing monitoring and providing programmability. However, wireless transcutaneous energy transfer is complicated by unpredictability and variability of biological tissue. The overall goal of the proposed work is to model and study tissue variability and its effects on power transfer and absorption in tissue. First, we will examine power transfer mechanisms in tissue with varying dielectric properties, and the associated behavior of different antenna topologies transmitting through tissue. Then, we will examine the effects of tissue variability on optimal frequency for power transfer through tissue, balancing power delivery and tissue absorption. The next focus will be on developing and evaluating tissue phantoms to model variable properties. Finally, we will propose strategies for adapting a passive implantable device in response to variations in the tissue environment.

**Keywords:** dielectric properties, electromagnetics, tissue variability, transcutaneous power transfer, wireless implantable medical devices.
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1.0 INTRODUCTION

The majority of Section 1.1 has been previously published in and reprinted in accordance with MDPI Open Access Policy from [1]. K. N. Bocan and E. Sejdić. Adaptive transcutaneous power transfer to implantable devices: A state of the art review. Sensors, 16(3), 2016. DOI: http://dx.doi.org/10.3390/s16030393

1.1 MOTIVATION

Implantable medical devices have become a huge market, with over 20 million individuals estimated to have an implanted medical device and over $300 billion in associated costs in the U.S. in 2000. Over 1 million patients in the U.S. have cardiac pacemakers and 250,000 new pacemakers are implanted each year, 100,000 implantable cardioverter defibrillators (ICDs) are implanted each year, and 120,000 patients in the U.S. have cochlear implants [2].

The term “implantable medical device” has been used to encompass many devices, from pacemakers to orthopedic implants to heart valves. For the purposes of this work, implantable medical device will refer to any implanted device that requires electrical power: for example, to stimulate the atrial node in the case of a pacemaker, or to power a small microprocessor in the case of an implanted sensing device.

Currently, most implantable medical devices are powered by implanted batteries or percutaneous wires. Powering implantable devices wirelessly has the potential to eliminate the need for percutaneous wires, reduce the size of the implant by replacing or supplementing battery power, and reduce or eliminate the need for battery replacement surgeries. Eliminating percutaneous wires reduces infection risk and improves patient mobility, while reducing
the size of the implanted device can also reduce infection risk at the subcutaneous pocket, as well as reducing the obtrusiveness of the implant to the patient. Even non-electronic implantable devices (such as orthopedic implants or heart valves) could potentially benefit from miniature wireless electronics to provide sensing capabilities.

1.2 DIRECTIONS AND GOALS

Wireless energy transfer is a broad research area that has recently become applicable to implantable medical devices. Wireless powering of and communication with implanted devices is possible through wireless transcutaneous energy transfer. However, designing wireless transcutaneous systems is complicated due to the variability of the environment.

The goal of adaptive wireless research is to design systems that adjust to variations in their environment. Adaptive systems provide the ability to maintain performance in the face of both unpredictability (variation from expected parameters) and variability (changes over time). Current strategies in adaptive (or tunable) systems include sensing relevant metrics to evaluate the function of the system in its environment, and adjusting control parameters according to sensed values through the use of tunable components. Some challenges of applying adaptive designs to implantable devices are challenges common to all implantable devices, including size and power reduction on the implant, efficiency of power transfer, and safety related to energy absorption in tissue. Challenges specifically associated with adaptation include choosing relevant and accessible parameters to sense and adjust, minimizing the tuning time and complexity of control, utilizing feedback from the implanted device, and coordinating adaptation at the transmitter and receiver.

Adaptive methods have been studied for transcutaneous energy transfer, primarily focusing on compensating changes in antenna alignment and positioning or impedance matching of transmitter or receiver circuitry. However, there is a lack of research focusing specifically on the effects of tissue variations. There exists a largely unquantifiable tissue variability among patients, making individual parameters unpredictable. Even in a single patient, parameters can vary over time, meaning that a device that initially works properly could diminish in
functionality, thereby decreasing the lifetime of the device. Therefore, there is a need to investi- 
gate the effects of tissue variations on transcutaneous systems and to develop methods to compensate for these variations.

1.3 DISSERTATION SCOPE

Wireless implantable medical devices necessitate electromagnetic energy transfer through biological tissue, which presents unique challenges as a wireless transmission medium in its structural complexity and dynamic properties. Transcutaneous energy transfer to implanted devices is highly dependent on the properties of the dynamic tissue medium. Even with measurements and empirical estimations of tissue dielectric properties, there exists uncertainty in how these tissue properties can be expressed as functions of physiological tissue parameters. Additionally, reported values suggest variation in tissue parameters among patients and within one patient across locations on the body and over time. It is therefore necessary to determine how these variations affect assumptions about electromagnetic transmission mechanisms including optimal frequency. Finally, to fully account for variations in application of a transcutaneous system, adaptive methods are needed to adjust parameters of the system in response to changes in the tissue environment.

There is a wealth of literature on wireless power transfer, energy harvesting, antenna design, and tissue electromagnetics. Design concepts and challenges from each of these fields are combined in the design of wireless implantable medical devices. As such, there is much that cannot be covered in the focused background for this work. Although there are significant biomaterials and biocompatibility challenges associated with implantable medical devices, the focus of this work is on wireless transcutaneous powering of implantable devices. Material selection and hermetic sealing of implanted components is an important area of literature, however these topics are outside of the scope of the current work. Additionally, the literature on wireless power transfer covers many components of system design, including power amplifier design, rectifier design, antenna design and voltage regulation. These components are essential and have been studied in terms of improving efficiency, and background
information is provided in this work where relevant. Similarly, this work will address but not focus on aspects of wireless communication in transcutaneous operation, instead maintaining the focus on transcutaneous wireless power transfer.

The first focus of this work is on the tissue medium itself: how tissue can be acceptably modeled in studying wireless power transfer, how its structure and properties vary over time, and how this affects power transfer and absorption. The second focus is on adaptive methods to compensate tissue variability, including changes in optimal frequency and adaptive power capture at the receiver through tunable effective area of an antenna embedded in the tissue. Optimal frequency maximizes power to the implant versus absorption in tissue, while optimal effective area maximizes power to the load based on power at the implant antenna, representing two stages of transcutaneous powering of an implanted device. The goal of this work is to investigate the effects of tissue variability at each stage and propose adaptation methods to accommodate tissue variability.

1.4 MAIN CONTRIBUTIONS

The proposed research will provide a comparison of planar antenna topologies for use in variable transcutaneous systems, expressions of tissue properties and their variations in terms of physiological parameters, relation of tissue variability to optimal frequencies in tissue, and development of an adaptive passive implanted receiver to accommodate tissue variations.

- Effects of variable tissue on achievable power gain will be examined at several frequencies and related to transmission mechanisms in tissue. The power gain will be determined through electromagnetic simulation.
- The effects of tissue variability on power gain and specific absorption rate (SAR) will be compared for dipole and loop antenna topologies. Power gain and SAR will be determined through electromagnetic simulation, and tissue dielectric properties will be varied according to reported values.
- Variability in tissue properties will be related to optimal frequency, defined as the frequency that maximizes power delivered to an implant with minimal absorption in the
surrounding tissue. A method of tuning at the transmitter will be proposed according to the expected variations in optimal frequency.

- Experimental tissue phantom formulations will be developed to represent variability in tissue structure, hydration, and ionic content. The dielectric properties will be measured for each phantom formulation, and the phantoms will be used to verify simulations of power gain through variable tissue.
- Expressions of tissue dielectric properties and their variability will be developed in terms of parameters such as hydration and ionic concentration. Expressions will be evaluated using tissue phantoms.
- A definition of effective area of an implanted receiver will be developed for a system operating in the midfield. A control method will be proposed to adapt a passive receiver in response to tissue variations. Controllable parameters, tunable components, and feedback will be evaluated.

1.5 DISSERTATION ORGANIZATION

Chapter 2 provides background on wireless implantable devices, electromagnetic design concepts, a review of the current literature on adaptive transcutaneous systems, and a review of tissue models for electromagnetic powering. Chapter 3 discusses the investigation areas of the dissertation work and the motivation for each point.

Chapters 4 and 5 cover transmission mechanisms, transcutaneous antennas, and power gain in relation to variable tissue. Chapter 6 expands the investigation of antenna topologies and transmission mechanisms in variable tissue in terms of optimal frequency, efficiency and specific absorption rate (SAR). Chapter 7 investigates simulation and experimental tissue models, and Chapter 8 presents a method of tuning a passive receiver in a variable tissue environment. Finally, Chapter 9 summarizes the impact of the dissertation work and suggests directions for future efforts regarding tissue variability and adaptive methods.
2.0 BACKGROUND

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2.1 GENERAL CONCEPTS

2.1.1 Wireless Implantable Devices

Fully implantable medical devices have already effected vast improvements in patient monitoring and treatment by eliminating percutaneous cables that are prone to infection and limit patient mobility. These include continuous therapy devices such as the implantable cardioverter/defibrillator (ICD), electronic pacemaker, implantable neurostimulators, and fully implantable drug delivery pumps [4–8].

The benefits of implantable devices also include prostheses, such as the cochlear implant and retinal implant [9–11]. Cortical implant research promises neural control of movement prosthetics and the valuable inclusion of sensory feedback [2]. Implantable electrodes have been developed for nerve and muscle stimulation [2].
Implantable sensing devices are particularly important in situations where the biological signals to be accessed are inside the body (neural, muscular, body fluids) and cannot be reliably sensed non-invasively [12]. This includes wireless implantable biosensors such as those used with insulin pumps, although commercial sensors are tethered to facilitate removal [13].

Functional challenges associated with fully implantable devices include powering the implanted device and monitoring or changing settings on the device. Commercial pacemakers and defibrillators rely on non-rechargeable batteries, where the battery is the determining factor in the size of the implant, and periodic surgeries are necessary to replace the device due to the limited battery lifetime (every 5 - 8 years for pacemakers, every 3 - 5 years for ICDs) [2].

Depending on the application, implantable devices demand varying levels of power. For example, a pacemaker requires on the order of 10 $\mu$W–1 mW, while a retinal prosthesis requires approximately 45 mW, while a ventricular assist device (VAD) requires 5–25 W, as illustrated in Figure 2.1 [14–17]. Commercially available VADs still require percutaneous drive lines and an external battery pack due to their high power requirements [2, 18]. There is also a considerable range of device sizes, with the VAD being much larger than a retinal prosthesis due to its functional requirements. A third consideration in addition to power level and size is whether a device needs continuous power. Interruption of power will impede the function of any device, but the consequences vary in severity. While interruption of power to a prosthesis will cause a decrease in quality of life and potentially secondary safety issues due to loss of sensory information, the case of a VAD power interruption is immediately life threatening. Meanwhile, an implanted biosensor may only need intermittent power to perform a sensor reading. These device considerations determine which powering methods are feasible for an implanted device.

Reviews and discussions of implantable device powering methods are provided in [16, 19, 20]. A focused comparison of inductive and ultrasonic energy transfer is provided in [21], and a review of acoustic energy transfer is given in [22]. Energy harvesting methods utilizing temperature gradients or piezoelectric materials have been developed, but as of yet cannot provide sufficient power for functioning implantable devices [17, 19]. While the optimal pow-
Figure 2.1: Range of power requirements of example implantable medical devices.

The powering method ultimately depends on the application, electromagnetic (EM) energy transfer has proven to be a promising wireless powering method with a broad range of system designs demonstrated in the literature, capable of delivering varying levels of power depending on size constraints of the device and the required function. Therefore, the focus of this review is on electromagnetic transcutaneous energy transfer.

Transcutaneous EM energy transfer has already enabled improvements to implantable devices by providing a method of powering an implantable device while reducing the dependence on implanted batteries and enabling remote communication with the implant. The cochlear implant was the first commercial wirelessly powered implantable device [9, 10, 23, 24]. Wireless power is provided to the cochlear implant through several millimeters of tissue, along with communication of processed auditory information from an external microphone. Retinal prostheses are a recent development and have been successfully implanted in humans to restore sight [11]. There have also been research efforts towards a fully implantable wirelessly powered VAD, even with the challenging power requirements [18].

Wireless transcutaneous systems generally include an external antenna and an implanted antenna, as illustrated in Figure 2.2. EM energy is transmitted through the tissue from the external antenna, captured by the implanted antenna, and rectified to power implanted circuitry. The energy captured at the implant can be utilized to recharge a battery or to directly power a battery-less implanted device. Eliminating the implanted battery typically enables miniaturization of the implant, improving patient safety and comfort by lowering infection risk and lessening the obtrusiveness of the implant [2]. Depending on the power requirements and the need for continuous power, the power source of an implantable device can
be primarily wireless, with a battery to provide backup power for critical functions, thereby prolonging the lifetime of the implant while providing a fail-safe for power interruptions [25].

Modulation of transcutaneous energy can be performed to achieve communication between the external and implanted sides. Back communication (implant to external) can be accomplished using a transmitter on the implant (active), or by modulating the energy from an external transmitter (passive), with active communication generally requiring more power at the implant. While the focus of this review is on electromagnetic energy transfer, similar communication has been achieved with other wireless powering methods such as ultrasound [26–28]. Wireless transcutaneous communication allows control of the implanted device behavior as well as access to information from the implant, such as readings from implanted sensors. This enables programmability of implanted devices and opportunities for post-surgical or remote long-term monitoring by reading information from the implant.

The main challenge associated with wireless transcutaneous energy transfer is achieving sufficient and reliable energy transfer to power the implant within safety limits on absorption in tissue. Electromagnetically, tissue is a lossy dielectric material with properties described by permittivity and conductivity. EM energy absorbed in tissue causes heating, hence there are fundamental safety concerns involved in transcutaneous energy transfer.

The goal of maximizing energy transfer to the implant while minimizing energy absorption in tissue has logically led to definitions of efficiency in terms of the power delivered to the implanted load, absorbed power, and input power [29–31]. The goal in the design of a system then becomes to maximize this efficiency. Conceptually, efficiency represents a distribution of input power in the source, antennas, system components, and load [32]. A more specific definition of efficiency for a particular system depends on multiple parameters, including: Antenna topologies, antenna dimensions and separation, the properties of the media surrounding the antennas, and operating frequency. These parameter determine the characteristics of the electromagnetic fields in the system. Once the system has been characterized, there are numerous strategies for optimizing the efficiency, whether through frequency tuning, impedance matching, or load tuning.
2.1.2 Environmental Variations

While a transcutaneous system can be optimized for a particular configuration, wireless transcutaneous energy transfer in practice is complicated by unpredictability and variability of the physiological environment. Variations occur due to antenna misalignment, movement of the antennas or the patient, implant migration, and changes in tissue structure and properties. Changes in the environment directly affect the performance and safety of a transcutaneous system, through effects on impedance and field characteristics. Impedance changes can reduce power transfer at the transmitter and receiver, as well as degrading efficiency if the impedance deviates from an optimal load or if there is increased absorption in tissue [33–35]. The degradation in power transfer and efficiency can lead to other problems in the system: Power reflections at the transmitter can damage the system, while reduced power delivery to the receiver can cause interruptions in function, and excess power delivery to the receiver could damage receiver circuitry or lead to tissue heating and safety concerns [36–40].

To avoid or lessen the degradation in system performance, sources of environmental variations must be examined and accommodated. One source of variation is changes in positioning or alignment of the antennas, whether due to movement of the patient or implant migration. Inductively coupled systems are particularly sensitive to variations in distance or alignment [41]. Some systems employ flexible antennas on the tissue surface, where movement can change the antenna geometry as well as the alignment [42]. In retinal prostheses, the implant antenna is in near constant movement due to its location in the eye [43]. Migration of an implant can occur after implantation, resulting in uncertainty of the implant location and potential misalignment of the antennas [44]. Distance or alignment variations can cause reduced power delivery and interruption of the implant function [45].

Although less attention is devoted to potential changes due to differences in tissue characteristics, there is evidence that tissue varies among patients and over time [46–53]. These variations are in addition to the known frequency dependence of tissue dielectric properties. Changes in body chemistry (such as hydration or fat content), inflammation, fibrous encapsulation, and changes in cellular structure are well-documented physiological processes that
affect tissue properties. Changes in tissue thickness will result in variable separation between an external and implanted antenna, causing either increased coupling due to thinner tissue or greater attenuation of the fields through thicker tissue. Changes in tissue structure or tissue chemistry, and associated changes in tissue properties, will affect the antenna electrical size and impedance as well as the optimal frequency of operation.

Attempts to minimize the effects of variations have been realized through design of insensitive systems and adaptive systems. Insensitive systems are designed for consistent operation over a range of environmental parameters, without the use of tuning components [37, 54]. Adaptive systems include tunable components to adjust the characteristics of the system in response to changes in the environment. The focus of this review is adaptive systems specifically designed for transcutaneous operation in the electromagnetic near or mid-field.

2.1.3 Wireless System Architecture

A wireless system typically consists of the components shown in Figure 2.2. At the primary or transmitter (external), the ac transmit signal is generated and fed to the transmit antenna. The EM field generated by the transmit antenna is captured by the receive antenna and converted to a dc voltage to power the receiver circuitry.

Oscillators and power amplifiers are common at the transmitter to generate the ac transmit signal. The power amplifier class and amplifier efficiency are design concerns at the transmitter. A matching network is often included to match the transmit antenna to the optimal load of the power amplifier to improve efficiency. At the receiver, rectifier efficiency is a design concern, as well as voltage regulation to supply a stable voltage to the receiver circuitry. The choice of transmit and antenna topology and dimensions depends on the application.

2.1.4 Field Regions

Wireless systems can operate in various field regions, depending on the operating frequency, antenna size, antenna separation and transmission medium. Field regions are characterized
by the types of fields present. The reactive near field is closest to the radiating antenna and contains primarily “stored” energy; the radiating near field contains both radiating and reactive fields with radiating fields dominating; the far field contains radiating electric and magnetic fields in planes transverse to the direction of propagation and the angular field distribution can be considered independent of distance from the radiating antenna [55, 56].

Definitions of the boundaries between the fields vary and the transition between regions is gradual, but the region boundaries can be approximated in terms of the electrical size of the antennas and the electrical distance between them (relative to the wavelength at the operating frequency) [55, 56]. The electromagnetic field region determines the interaction (if any) of the transmitting and receiving antennas, which guides the associated design and analysis of the system. Theoretically, any antenna system can be made to operate in any field region by varying the size or separation of the antennas. However, the efficiency of a system is expected to degrade when operating outside of the desired field region to the point where the system may cease to function.

Systems designed to operate in the reactive near field include inductively coupled or magnetically coupled systems, where the magnetic field of a transmitter coil induces a current in a receiver coil [42]. The term “antenna” is not typically used to refer to the coils in inductively coupled systems. Rather, the term “coil” or “resonator” is used to designate

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**Figure 2.2:** Simplified general wireless transcutaneous system architecture.
that the coils are designed as coupled (reactive) and not as radiating antennas. Capacitive coupling via the electric field is also possible in the reactive near field, and has enabled power transfer to implantable devices in proximity to metallic implants [57]. Efficiency in an inductively coupled system decreases significantly with greater coil separation or disparate transmit and receive coil dimensions due to reduced coupling [14, 31, 44]. This effect tends to limit the miniaturization of implantable inductively coupled devices. Magnetic resonance systems have been shown to have high efficiency up to antenna separations of 1–2 coil diameters by maximizing the coil quality factor, but efficiency is still dependent on similar size of the transmit and receive resonators [14, 31, 58].

Weakly coupled systems can be designed to have higher efficiency in the radiating near field region by deliberately utilizing differently sized antennas [59]. Such a system can be designed with the receive antenna dramatically miniaturized relative to the transmit antenna. This “midfield” region of operation has been recently investigated for transcutaneous powering of miniature implantable devices.

Far field systems operate at a distance and frequency such that there is no detectable coupling between the transmit and receive antennas [56]. This region is less used for implantable devices to avoid problems with attenuation of propagating fields in tissue. The focus of this review is on transcutaneous systems with an external antenna on or near the tissue surface and an implanted antenna, limiting the discussion to reactive near field and midfield systems.

For transcutaneous applications, inductively or magnetically coupled systems are well suited when the coil separation can be on the order of the coil diameter to maintain high efficiency, and where the external and implanted coils can be of similar size. Such systems have been shown as capable of delivering tens of watts of power to an implanted device at small coil separations [14, 45]. In cases where miniaturization of the implanted device is a priority, the separation between the implanted and external antennas is likely to be larger than the antenna dimensions. Therefore, the efficiency will be maximized by designing for midfield operation with asymmetrically sized antennas. Such systems have been shown to achieve power delivery of up to several milliwatts to a millimeter-sized implant [17].
2.1.5 Tissue Properties

Two main current paths can be identified in tissue, corresponding to permittivity and conductivity and the current paths described in Ampere’s Law. One path is conduction current, corresponding to the movement of ions in tissue and therefore the conductivity (\(\sigma\)). The other path is displacement current, a function of the electric flux and permittivity (\(\epsilon\)).

Conductivity is a function of ionic mobility in the tissue medium, where ions such as sodium and potassium act as charge carriers. Permittivity is a function of charge buildup at cell membranes (cell membrane capacitance) and alignment of molecular dipoles with an applied field [60]. Dielectric relaxation is also included in the complex permittivity, describing the effect of molecular dipole rotation delay in response to an applied field. Tissue conductivity and permittivity are frequency dependent, with conductivity increasing with frequency, and permittivity decreasing with frequency.

In general, conduction current can be utilized at low frequencies and small antenna (electrode) separations, where the movement of ions creates capacitive charge transfer. At higher frequencies or greater antenna separations, conductivity contributes to loss due to field vector directions and the inability of the ions to build up along interfaces. At higher frequencies, displacement current becomes significant, and the induced current due to greater flux competes against conductive losses. At GHz frequencies, dielectric relaxation losses also become significant. The specifics of energy transfer differ based on the tissue structure, thickness, and chemistry. Tissue properties have been measured for multiple tissue types across frequency, and parametric models have been developed to represent tissue properties over a range of frequencies [49, 60].

Tissue properties are particularly relevant to wireless power transfer due to their relationship to power dissipation in tissue, which can lead to tissue heating and therefore presents safety issues. Energy absorption in tissue is quantified as specific absorption rate (SAR), measured in watts per kilogram. SAR limits defined by several organizations are used as safety guidelines for transcutaneous energy transfer. These limits have been defined based on studies of physical and behavioral effects of electromagnetic field exposure in animals and humans, but are not specific to medical use.
The IEEE Standard for Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields 3 kHz–300 GHz (IEEE Std C95.1-2005) specifies limits of 0.4 W/kg and 10 W/kg for whole body and local SAR, respectively [120]. The International Commission on Non-ionizing Radiation Protection (ICNIRP) guidelines specify that general public exposure is limited to 0.08 W/kg whole body, 2 W/kg local for head and trunk, and 4 W/kg local for limbs, with local SAR averaged over 10 g of tissue for any 6 min period [61]. For uncontrolled exposures of the general population, the U.S. Federal Communications Commission (FCC) limits SAR to 1.6 W/kg averaged over any 1 g cube of tissue [62]. In addition to dielectric losses, heating can also occur due to heat dissipation of the receiver circuitry, sometimes exceeding the one-degree Celsius temperature increase that is a basis for SAR limits [43, 63, 64].

2.1.6 Operating Frequency

Due to the complex frequency dependent dielectric properties of tissue, the existence of an optimal frequency has been proposed that balances power transfer and losses in tissue [29, 30]. The optimal frequency was determined by defining a measure of efficiency in terms of power delivered to the load and power absorbed in tissue [29]. Ho et al. [59] showed that the highest efficiency is achievable in the midfield for weakly-coupled disparate-sized antennas, resulting in an optimal frequency in the sub-GHz to GHz range for a cm-size transmitter and mm-size receiver at cm-separation. Higher frequency has also been recently investigated as it enables miniaturization of antennas, greater antenna impedance, and greater open circuit voltage at the implant [30, 34, 41, 59, 65].

2.1.7 Transcutaneous Antennas

Particular antenna topologies tend to be favored for transcutaneous systems due to size constraints and the effects of tissue properties. Miniature antennas are desirable for implants to reduce patient discomfort and infection risk, but this results in electrically small antennas where impedance mismatch can lower efficiency [66]. Miniature antennas have been shown to be most efficient in weakly coupled systems operating in the midfield [29].
Tissue’s high permittivity (due to its composition of mostly water) reduces the net electric field, an effect of molecular alignment with an applied field. Loop or coil antenna topologies are widely used in the implantable device literature to take advantage of tissue’s non-magnetic permeability and utilize the magnetic near field. Fewer turns have been shown to be better in proximity to conductive high permittivity tissue, with quality factor decreasing with increasing turns [30]. The induced current at a loop receiver depends on the flux through the loop, hence the advantage of using higher frequencies. At frequencies and antenna dimensions such that there is midfield operation, it has been shown that both the electric and magnetic fields contribute to power transfer [67].

Inductive antennas (loops) act as better receivers due to their low impedance, while capacitive antennas act as better transmitters due to their larger impedance which limits current flow at the transmitter [16]. Low impedance can lead to power loss and heat generation due to high currents when used as a transmitter [41].

Printed antennas are favored due to the potential for miniaturization and manufacturability [30]. More complex antenna designs have been investigated to focus fields at the site of the implanted receiver, and to minimize excess power dissipated in tissue [17, 44, 68].

2.1.8 Power Gain and Efficiency

Because the function of any receiving device depends on its power supply, it is desirable to maximize power transfer to an implant. However, transcutaneous operation necessitates consideration of safety limitations relating to absorption of electromagnetic energy in tissue. The need to maximize power to the load while minimizing absorption in tissue then evokes definitions of efficiency.

Definitions of efficiency vary in terms of what the load power represents and what it is expressed relative to: power available to the load relative to the power absorbed in tissue [17, 29, 44]; power to the load relative to power available from the source, representing the transducer gain [69]; power to the load relative to power delivered by the source, with maximum efficiency achieved with an optimal load [34]; power to the load relative to power input to the network, maximized with a conjugately matched load in a weakly coupled system
Optimal load impedances have been derived to maximize efficiency according to the various definitions [31, 34, 59]. The definitions of efficiency overlap with definitions of power gain, but for passive systems the term efficiency is used to indicate that the power gain from source to load is less than one.

Defining efficiency relative to the power from the source, there exists an upper bound of 50% power efficiency with simultaneous conjugate matching [34]. While the load resistance is equal to the source resistance for maximum power transfer, the load resistance is larger than the source for higher efficiency (relative to the source power) [31]. Designing for maximum efficiency in this case may reduce the range of operation due to reduced power transfer to the load [31]. However, maximizing voltage at the implant is sometimes preferable even at the expense of power transfer, to ensure adequate turn-on voltage while minimizing losses in the system. Additionally, in the case of delivering power from a power amplifier, efficiency (defined as output power from the power amplifier relative to input power) is prioritized to minimize heat dissipation in the circuit. In this case, the power amplifier is designed to operate at maximum efficiency, rather than using conjugate matching, and an optimum load can be defined [34, 70].

Defining efficiency as load power relative to power absorbed in tissue, maximum efficiency is achieved through conjugate matching [29, 59]. For delivering power to a load through tissue, this efficiency is valuable when power dissipation in tissue is of more significant interest than power dissipation elsewhere in the system.

The system efficiency or gain can be decomposed into component efficiencies, including coupling efficiency that represents the efficiency between the transmit and receive antennas (coils) [71], power amplifier efficiency [34], and rectifier efficiency [72–74]. In midfield cases where losses in tissue are significant and lumped element models are no longer appropriate, network analysis allows representation of a linear system in terms of scattering parameters (S-parameters) to calculate efficiency or power gain (see Appendix C) [29, 70, 72, 75].
2.1.9 Impedance Matching

Impedance matching networks can be added to transform the real and reactive impedances looking into portions of a system, to minimize voltage reflections at interfaces, to conjugately match for maximum power transfer, or to achieve optimum loading.

Common network topologies include L, T, and pi networks consisting of inductors and capacitors in configurations capable of increasing or decreasing input impedance. The choice of topology depends on design parameters including the desired bandwidth, impedance transformation range, complexity, and available area [76].

Pi-match networks can be used to both increase and decrease impedance, while L-match networks can only be designed to transform the impedance in one direction (either increase or decrease) [70]. Pi networks can provide wider band matching, but L-match is appropriate when efficiency is of primary concern [34].

2.1.10 Impedance and Material Properties

Impedance is closely related to material properties of conductivity, permittivity, and permeability, and properties of inductance, capacitance, and resistance. Impedance is typically represented as a function of resistance ($R$) and reactance ($X$) as given in Equation 2.1. Resistance is then a function of geometry and conductivity, while reactance is a function of geometry and permittivity or permeability. Reactance can also be expressed in terms of inductance ($L$) or capacitance ($C$), also indicated in Equation 2.1.

$$Z = R + jX = R - \frac{j}{\omega C} = R + j\omega L \quad (2.1)$$

2.1.11 Resonance

Resonance in the context of antenna or circuit design typically refers to a system with completely real input impedance. A natural resonance associated with the system geometry has also been defined as where the determinant of the scattering matrix approaches zero [66]. For the purposes of this review, the focus will be on the first definition of resonance
due to its relevance to impedance matching. Resonance tuning is equivalent to impedance matching to achieve an all-real input impedance, implemented to maximize power delivered to or from a real impedance. Designing antennas with inherent resonance can avoid losses associated with an added matching network. Operating transmit and receive antennas at the same resonant frequency increases the voltage gain between the antennas [15, 43].

Near field systems tend to utilize resonant tuning to design transmitting and receiving antennas (coils) to function at the desired operating frequency. Resonance tuning can include added capacitors or inductors to compensate for an antenna’s reactive impedance, or antennas designed such that the inherent capacitance and inductance are tuned to resonance [66]. Magnetic resonance systems have been designed with three or four coils, where the additional coils perform impedance transformation at the transmitter and/or receiver [58, 77]. This is assuming that the coils are being designed with a specific frequency in mind. In some cases, the frequency is instead tuned to achieve resonance with the given impedance characteristics of the system. It has been noted that the optimum load differs from resonant tuning if the media between the transmit and receive coils is conductive, as in the case of transcutaneous operation [34].

2.1.12 Quality Factor

Quality factor ($Q$) is a measure of the energy stored in a system relative to the energy dissipated or lost per time. For a passive parallel RLC network at resonance, $Q$ is defined as in Equation 2.2 [70].

$$Q = \frac{\omega_0 E_{\text{stored}}}{P_{\text{avg}}} = \frac{R}{\sqrt{LC}}$$

(2.2)

The stored energy ($E_{\text{stored}}$) is the peak energy stored in the capacitance ($C$) or inductance ($L$), as the oscillatory energy is transferred between the two. The power dissipated ($P_{\text{avg}}$) is the power through the resistance ($R$).

Quality factor represents the ratio of the current flowing in inductors/capacitors to the net current through the network. Therefore, operating at high $Q$ increases the voltage swing across the LC part of the network. Higher $Q$ also corresponds to narrower fractional
bandwidth, as shown in Equation 2.3 [70]. High $Q$ for an antenna equates to low radiation efficiency, and there is a fundamental limit on the minimum $Q$ of electrically small antennas [56].

$$\frac{BW}{\omega_0} = \frac{1}{Q}$$

(2.3)

One strategy for achieving better radiation properties is space-filling or meandering, where an antenna occupies an equivalent area but is made to have an increased electrical length [56]. In the case of near field operation, however, designing electrically small antennas to operate at high $Q$ is desirable to increase the “stored” near field energy. The basis for magnetic resonance systems is designing coil antennas to operate at high $Q$ [14]. Because high $Q$ equates to narrower bandwidth, coupled systems can be more efficiently designed using separate coils (and separate frequencies) for power and data [74]. Quality factor of a material is inversely related to the loss tangent ($\tan \delta$) of the material, defined in Equation 2.4 in terms of conductivity and permittivity.

$$\tan \delta = \frac{1}{Q} = \frac{\sigma}{\omega \varepsilon}$$

(2.4)

### 2.2 ADAPTIVE TRANSCUTANEOUS SYSTEMS

The goal of adaptation in any system is to maintain desired performance despite environmental variations. How “performance” is defined and the strategies to maintain performance depend on the application and the operating field region of the antennas.

The general steps in the design of an adaptive system are illustrated in Figure 2.3: (a) Defining a parameter to optimize (a performance metric) based on the functional goal of adaptation; (b) developing a tuning method based on the available controls affecting the performance metric and the expected range of parameter variations; and (c) implementing sensing to determine or evaluate the tuning state. This section will provide a conceptual overview of the strategies employed in each stage of the design, and the following section will cover specific implementations in the literature.
2.2.1 Performance Metric

The parameter used as an indication of system performance depends on the desired function of the system and the system characteristics. Strategies include minimizing reflections at interfaces, maximizing power transfer, maximizing efficiency, and maintaining constant load voltage.

Maximum power transfer can be maintained in variable environments through tunable impedance matching. Minimizing reflection achieves maximum power transfer when the goal is matching to an all-real impedance. When matching to complex impedances, maximum power transfer is achieved through complex conjugate matching.
There are various definitions of efficiency, but power efficiency is typically defined as load power relative to input power or power absorbed in tissue [29, 34]. Design strategies to maintain efficiency in the presence of environmental changes are many and varied, but most can be generally classified into frequency tuning and impedance tuning.

### 2.2.2 Tunable Components

Tunable components are essential to adapting a system with the goal of optimizing a chosen parameter. In frequency tuning applications, a voltage controlled oscillator can be used to adjust the transmitter frequency [42, 78, 79]. The switching frequency of the power amplifier at the transmitter can also be controlled [74].

In impedance tuning applications, variable inductances and capacitances can be used in matching networks to transform impedances. Transductors provide current-controlled inductance, and can withstand large voltage and current, but they tend to be too bulky for wearable or implantable devices [42, 80]. Microelectromechanical systems (MEMS) inductors have been shown to provide high efficiency and tuning range, but variable inductors remain difficult to implement on-chip for miniature implants [81, 82].

Variable capacitances tend to be favored over inductances due to the widespread use of inductive coil antennas and the higher quality factor of capacitors on-chip [34, 76, 79]. Voltage-controlled capacitors (varactors) offer compact area, but they can exhibit non-linearities at radio frequencies and require an analog control voltage [42, 65, 69, 76, 83]. Switched capacitor banks have also been used to vary parallel matching capacitance, but their disadvantages include the greater space occupied and the necessary discretization of the capacitance values [77, 84–86].

Another approach to impedance tuning is duty cycling. Ahn et al. [77] implemented buck-boost converter load connection duty cycling to modulate effective load in an inductively coupled system; Si et al. [84] duty-cycled switched capacitors to achieve variable effective capacitance.
Variable output dc-dc converters and control of power amplifier supply voltage have been used to regulate supply voltage and input power at the transmitter [39, 42]. In some cases the antenna itself can become a tunable component, either through adjusting multiple feed points to achieve beamforming, or through changes in the antenna geometry [17, 87, 88].

2.2.3 Sensing and Feedback

Feedback can be implemented to evaluate the state of the performance metric and to determine the appropriate control of tuning components. The various approaches in the literature fall into three categories, illustrated in Figure 2.4: (1) Sensing and processing at the external transmitter to tune components at the transmitter; (2) sensing and processing at the implanted receiver to tune components at the receiver; and (3) sensing at the implanted receiver communicated to the external transmitter for processing and tuning components at the transmitter.

Figure 2.4: Existing approaches to adaptive transcutaneous system design.
Parameters used for feedback include the rectified voltage indicating received power [42, 89], reflected voltage indicating impedance mismatch [78], and phase differences to detect reactive impedance [42]. Information on parameters sensed at the receiver can be communicated to the transmitter passively by modulation of the power carrier, or actively through the use of an implanted transmitter.

The tuned components, sensed parameter, and the control method are closely related. The control method is designed based on the parameters available to sense, the controllable parameters that affect the optimized parameter and the subsequent choice of tunable components, and any design constraints on tuning time, size, and power consumption.

The tuning time must be controlled such that it does not interfere with primary functions of the system such as communication. An iterative algorithm such as gradient search is relatively simple to implement, but has a disadvantage in tuning time compared to single iteration methods based on direct calculation [81]. However, the power and memory requirements of complex vector calculations can limit the application of such methods on miniature implantable devices. A balance must be achieved among tuning time, complexity, and accuracy.

Adaptation control can be implemented at either the transmitter or receiver, or both the transmitter and receiver. Strategies at the transmitter include adjusting transmitter power and frequency, or tunable matching at the transmit antenna [14, 36, 38, 39, 42, 45, 83, 84, 90, 91]. Adaptation strategies at the receiver mainly include tuning matching networks to achieve optimal load [73, 91–93]. Adaptation of both sides of the network is necessary for maximum power transfer efficiency, and may require communication of the adaptation state from one side of the network to the other [65, 71]. For any adaptation strategy to extend to an implantable device, it must be miniature and low power to avoid adding significantly to the existing power and size requirements of the implant. There are fewer restrictions on the transmitter in terms of size and power consumption, but safety is still a primary concern when transmitting power through tissue [16].
2.2.4 Implementations of Adaptive Transcutaneous Systems

The combination of strategies discussed above depends on which environmental variation(s) are addressed and where in the system the adaptation is performed. A review of implementations in the literature is given in Table 2.1, indicating the goal of adaptation (including the performance metric), tuning performed and feedback at the external transmitter (Tx) and/or the implanted receiver (Rx), and the environmental variation for which the adaptation is intended to compensate. Designs of interest are categorized and discussed in the following subsections.

Table 2.1: Summary of literature on adaptive transcutaneous systems.

<table>
<thead>
<tr>
<th>Goal</th>
<th>Tuning</th>
<th>Feedback</th>
<th>Variation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Tx Rx</td>
<td>Tx Rx</td>
<td></td>
</tr>
<tr>
<td>[78] Min reflections</td>
<td>Frequency (VCO)</td>
<td>Reflected voltage</td>
<td>Resonant frequency</td>
</tr>
<tr>
<td>[79] Max efficiency</td>
<td>Frequency (VCO)</td>
<td>Antenna voltage</td>
<td>Resonant frequency</td>
</tr>
<tr>
<td>[42] Max received voltage</td>
<td>Frequency (VCO), Power (supply voltage)</td>
<td>Rectified voltage</td>
<td>Resonant frequency, coupling</td>
</tr>
<tr>
<td></td>
<td>Max driver efficiency</td>
<td>Amplifier $Z_L$ (transducer)</td>
<td>Phase between LC and coil driver voltage</td>
</tr>
<tr>
<td>[83] Resonance</td>
<td>Impedance (varactor)</td>
<td>Reflected voltage phase change</td>
<td>Distance and impedance</td>
</tr>
<tr>
<td>[84] Resonance</td>
<td>Impedance (duty-cycled $C$)</td>
<td>Frequency</td>
<td>Impedance</td>
</tr>
</tbody>
</table>
Table 2.1 (continued).

<table>
<thead>
<tr>
<th>Goal</th>
<th>Tuning</th>
<th>Feedback</th>
<th>Variation</th>
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<tbody>
<tr>
<td></td>
<td>Tx</td>
<td>Rx</td>
<td>Tx</td>
</tr>
<tr>
<td>[41]</td>
<td>Stable received power</td>
<td>Power (supply voltage)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>[15]</td>
<td>Stable received power</td>
<td>Impedance (duty-cycled C)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>[38]</td>
<td>Stable received power</td>
<td>Power (supply voltage)</td>
<td>Storage capacitor voltage</td>
</tr>
<tr>
<td>[74]</td>
<td>Max efficiency</td>
<td>Frequency</td>
<td>Resonator voltage</td>
</tr>
<tr>
<td>[81]</td>
<td>PA optimal load</td>
<td>Matching impedance (simulated)</td>
<td>Antenna impedance</td>
</tr>
<tr>
<td>[33, 90, 94]</td>
<td>Match to PA or LNA</td>
<td>Matching impedance (varactor, switched C bank)</td>
<td>Antenna impedance</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Matching impedance (varactor)</td>
<td>Antenna impedance</td>
</tr>
<tr>
<td>[16, 43]</td>
<td>Max power transfer efficiency</td>
<td>Power (supply voltage)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Impedance</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>[65, 72, 82, 91]</td>
<td>Max power transfer efficiency</td>
<td>Matching impedance (switched C bank)</td>
<td>Rectified voltage</td>
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<tr>
<td></td>
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<td>Matching impedance (switched C bank)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>Goal</td>
<td>Tuning</td>
<td>Feedback</td>
<td>Variation</td>
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<tr>
<td></td>
<td>Tx</td>
<td>Rx</td>
<td>Tx</td>
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<tr>
<td>[39]</td>
<td>Stable received power</td>
<td>Power (supply voltage)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>[80]</td>
<td>Max received voltage</td>
<td>Impedance (switched C bank)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td></td>
<td>Stable received power</td>
<td>Power (supply voltage)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>[73]</td>
<td>Max efficiency (optimum load)</td>
<td>Impedance (switched C bank)</td>
<td>Rectified voltage gradient</td>
</tr>
<tr>
<td>[36]</td>
<td>Stable received power</td>
<td>Frequency (ZVS)</td>
<td>Switch transistor drain voltage</td>
</tr>
<tr>
<td>[17]</td>
<td>Max efficiency (relative to absorption)</td>
<td>Field pattern (antenna feeds)</td>
<td>Rectified voltage</td>
</tr>
<tr>
<td>[93]</td>
<td>Max power transfer</td>
<td>Impedance (switched CLC pi bank)</td>
<td>Antenna port voltage</td>
</tr>
<tr>
<td>[45]</td>
<td>Max power transfer</td>
<td>Frequency (ZVS)</td>
<td>Resonant tank voltage</td>
</tr>
<tr>
<td>[85]</td>
<td>Max efficiency</td>
<td>Impedance (variable C)</td>
<td>S-parameters</td>
</tr>
<tr>
<td>[25]</td>
<td>Max power transfer</td>
<td>Matching impedance (variable C)</td>
<td>S-parameters</td>
</tr>
<tr>
<td></td>
<td>Max efficiency</td>
<td>Power</td>
<td>Reflected voltage</td>
</tr>
<tr>
<td>[14]</td>
<td>Stable power, max efficiency</td>
<td>Power, resonant frequency</td>
<td>Received power</td>
</tr>
</tbody>
</table>
2.2.4.1 Input Power Adjustment  Changes in the separation distance and alignment of antennas can cause variations in coupling and power transfer [96]. One strategy to compensate involves regulating input power at the transmitter. The motivation for this tuning is to avoid or reduce the power regulation burden at the implant, which contributes to heat dissipation [41].

Van Schuylenbergh and Puers [42] adjusted the transmit power using a voltage-controlled boost regulator, based on feedback of the dc input voltage at the internal regulator.

Wang et al. [38] compensated for coil movement or load changes, regulating transmitted power by adjusting the supply voltage to a power amplifier. The supply voltage was adjusted according to the detected voltage on a storage capacitor at the receiver, communicated to the transmitter over a second frequency band. Power level adjustment and back telemetry were staggered to avoid interference.

Si et al. [41] demonstrated power regulation at the transmitter, designed to limit the size and heat dissipation associated with power regulation at the receiver. Power was regulated by adjusting the supply voltage at the primary resonant converter based on feedback on the dc voltage at the receiver, communicated wirelessly to the primary.
Ng et al. [16, 43] proposed a system to compensate for eye movements and fibrous growth in retinal implants, regulating the supply voltage at the transmitter based on feedback on the voltage at the secondary, communicated through back telemetry. Adjusting the transmitter resonant frequency for maximum efficiency by tuning the transmitter capacitance was also proposed, based on voltage at the receiver. A primarily theoretical discussion was presented for tuning and feedback.

Kiani and Ghovanloo [39] designed a system to compensate for distance and angular alignment variations in an inductively coupled system, regulating transmit power based on an indication of voltage across the receive coil. Back telemetry was performed if the received voltage was greater than a reference value, so the transmitter increased transmit power unless bits were received from the implant. The design focused on using primarily off-the-shelf components.

Waters et al. [25] designed a system to adjust transmit power based on reflected voltage at the transmitter to compensate for coupling variations in a magnetic resonance system. They later proposed an auto-tuning algorithm to adjust transmit power based on detected load power, and measured the temperature of the receive coil as a measure of efficiency [14]. Back telemetry of load power was mentioned as future work.

2.2.4.2 Adaptive Antennas  In an effort to compensate for changes in alignment of the antennas, which can lead to changes in coupling and power transfer, transmitters have been designed with location and focusing capabilities. The concept is related to power regulation, but the regulation can involve multiple antennas or control of multiple feeds to a single antenna.

McMenamin et al. [97] presented a system that adjusted powering of antennas in animal cages for bio-telemetry experiments based on the animal’s detected position. The goal was to focus wireless electromagnetic energy to power a mobile telemetry unit worn by the animal. The floor of the animal cage consisted of an array of overlapping planar spiral coils, and the magnetic field within the cage was focused by only providing power to the coils closest to the detected position of the mobile telemetry unit. A small magnetic tracer was embedded in the mobile telemetry unit and an array of magnetic field sensors in the cage floor was
used to detect the unit’s location. Ho et al. [17] manipulated the field pattern itself to maximize efficiency by implementing a transmitter with dynamic focusing capability. The transmitter was designed to power a miniature (2 mm × 3.5 mm) implant in the midfield (5 cm separation) for pacemaker or cortical implant applications. Through control of the phases of the antenna feeds, the field was focused at an implant in various locations. The focusing was adjusted using an optical indication of received power as feedback for the purpose of the experiment. A method of back telemetry to practically adjust the focusing was not discussed.

2.2.4.3 Frequency Tuning  Frequency tuning has been utilized to correct for changes in antenna impedance due to the surrounding environment, including maintaining operation at a resonance frequency. Frequency tuning can be difficult due to FCC regulations on frequency bands [85, 92]. It is mainly used in inductively coupled systems, where the coils can be designed such that the coupled system does not radiate significantly, but the operation is highly affected by coil separation distance [69]. A review of automatic frequency control techniques is provided in [98], and several examples of frequency tuning in transcutaneous systems are presented here.

Ko et al. [78] used a voltage-controlled oscillator to adjust the transmit frequency in response to detuning of an animal’s cage for bio-telemetry experiments. The goal was to maintain operation at the resonant frequency of the transmitter circuit including the animal’s cage, in order to power a battery-less implant for chronic animal telemetry. Detuning due to the animal’s movement was detected by measuring reflected voltage at the transmitter, and the transmitter frequency was adjusted to minimize reflections. Phase-lock loop techniques were proposed as a strategy for tuning the receiving unit in response to changes in the transmitter frequency.

Fernald et al. [79] also used a voltage-controlled oscillator, and adjusted the transmit frequency in response to changes in the antenna resonant frequency. The application was a general-purpose implant for animal telemetry experiments. The system performed a resonant search, sweeping the frequency and monitoring the voltage amplitude across the transmit antenna. When the voltage swing reached a threshold value, the frequency was fixed and the
system began transmitting at that frequency. Baker and Sarpeshkar [74] presented a class-E controller to compensate for changes in coupling, comparing the transmit resonator voltage to a reference voltage and controlling the amplifier switching, with the goal of maintaining link efficiency with robustness to changes in coil separation. They also present a feedback control analysis of a coupled system for wireless electromagnetic power, and an experimental investigation of the effects of coil separation on peak efficiency. Experimental tests of their switching control system showed less than 16% variation in rectified output voltage over coil separations of 1–10 mm.

Ahn and Hong [36] adjusted operating frequency to maintain constant output voltage with coupling and load variations. The goal was to implement a low-power solution without requiring complex active circuits or external components. The switching frequency of a self-oscillating class-D power amplifier was controlled based on feedback from the drain of the switch transistor. The frequency-tuning system was demonstrated to have relatively constant output voltage over load variations and distances up to 12 mm, and able to maintain constant voltage over greater distances at higher load resistance.

Wang et al. [45] implemented a zero voltage switching follower design to compensate for coupling variations in inductively coupled systems, with the goal of providing power to an implantable heart pump. Switching frequency control was performed based on voltage feedback at the transmitter, in order to maintain middle zero voltage switching operation. The output power and efficiency were measured with and without frequency control, and the tuned system was able to deliver 10 W at greater coil separation and higher efficiency than without frequency control.
**2.2.4.4 Impedance Tuning**  Frequency tuning has been paired with automatic impedance matching to achieve resonance of an antenna while maintaining matching to a feed line. Van Schuylenbergh and Puers [42] adjusted the transmit frequency using a voltage controlled oscillator based on feedback of the dc input voltage of the internal regulator, communicated from the receiver using a third external sensing coil. Variable impedance matching at the transmit coil was also implemented based on the tuned transmit frequency, using a phase comparator to detect detuning of the coil. Hirata et al. [83] used two varactor diodes to tune the resonance frequency of a coil and to match the coil impedance to a transmission line. The matching was evaluated by sensing the phase change of reflected voltage at the transmitter.

Automatic impedance matching has also been implemented in systems operating at a fixed frequency to compensate for impedance variations that detune the system and decrease power transfer and efficiency. Impedance mismatches can be the result of antenna position or impedance changes due to tissue parameter variations or movement. It has been suggested that although frequency tuning is easier to accomplish, higher efficiencies are possible with impedance matching [69]. Adaptive matching and frequency tuning are two methods of adjusting the system to operate at a desired matching state; one adjusts the impedance of the system to achieve matching at a fixed operating frequency, while the other adjusts the frequency to operate at a matched system impedance. This is illustrated in Figure 2.5, with matching indicated as minimizing the reflection coefficient ($\Gamma_{im}$).

Si et al. [15, 84] implemented switched capacitors to control effective capacitance in a push-pull resonant converter to achieve resonance at a given reference frequency, to compensate for changes due to load or circuit parameters. Rodes et al. [95] developed a tuning system for adjusting capacitance of a half-bridge voltage-mode resonant converter. The work referenced [15, 84], stating that the half-bridge voltage mode resonant converter is better suited to medical applications due to the current-fed push-pull converter’s inductances and the error terms in the switched capacitor transfer functions.

Waters et al. [85] demonstrated automatic impedance matching to match transmit and receive coil resonator input impedances to the source and load impedances, to accommodate changes in coupling of a magnetic resonant system due to changes in distance between the
resonators. The demonstrated systems use extracted S-parameters to calculate the states of tunable matching networks. In the first case, the ideal matching impedance is calculated from the S-parameter matrix; in the second case, the matching state is found using an optimization algorithm, calculating the system efficiency from the S-parameters. The use of a network analyzer to extract S-parameters precludes direct application of this method in a real-time implantable system.

Chan Wai Po et al. developed a method of calculating complex antenna impedance based on detected voltage on capacitors, and using the calculated impedance to determine and set tunable reactive components in a matching network [33, 81, 90, 94]. Implementations included MEMS variable inductors, switched capacitor networks, and varactors. The system was designed to compensate for impedance changes in pacemaker applications due to tissue differences, patient position, or nearby objects. Their one-iteration method reduces the time required for tuning, but requires vector calculations. Tuning power requirements are not provided, but it is proposed that the method reduces power requirements due to reduced tuning time. The method is applicable to both the transmitter and receiver, stated as the first system able to match both in a single process [33]. The transmit antenna impedance is matched to a power amplifier, while the receive antenna impedance is matched to a low noise amplifier.
Park and Ghovanloo [93] controlled for impedance variations due to the environment surrounding an intraoral sensing device. Switches in a CLC pi matching network were controlled by a microcontroller and set according to the output power monitored via a power detector. The system swept 16 possible settings and monitored output power, then set the configuration to the optimal switch setting, in a process taking 30 ms and repeated every second. However, the tuning was acknowledged to potentially impede real time operation due to interrupting communication of the intraoral device to an external receiver. The application was not specifically an implantable device, but was designed with size constraints to be worn on a dental retainer and addressed many of the same issues of energy transfer through tissue.

O’Driscoll et al. investigated the effects of misalignment, implantation depth, and tissue composition on system impedances in an implantable device, with focus on devices for neural recording [65, 72, 82, 91]. The impedance was adjusted using a switched capacitor array, based on feedback of the voltage across the resistive load. The algorithm was a hybrid of gradient and binary search, where only the sign of the gradient was calculated to save power. The capacitor array and control were implemented on-chip, with the adaptation algorithm off-chip. The method is applicable to matching at both the transmitter and receiver.

Carta et al. [80] designed a self-tuning inductively coupled system to compensate for misalignments and distance changes, using a switched capacitor bank at the transmitter. The capacitor values were switched based on the voltage at the receiver, in a two stage process. First, the capacitance combinations were varied to determine the highest induced voltage at the receiver. Then, the capacitance was adjusted to maintain the voltage level determined in the first step.

Zargham and Gulak [73] developed a system for maximizing efficiency by tuning the load impedance to an optimal load. A switched capacitor array was controlled according to a sign-based gradient descent algorithm to maximize the power delivered to the rectifier load at the receiver. The implant receiver coil and circuitry were implemented on a single die in CMOS.
2.3 ELECTROMAGNETIC TISSUE MODELS

2.3.1 Tissue Properties and Variations

This section reviews measurements of tissue properties reported in the literature and discusses mechanisms of tissue property variations. It is well-established that tissue properties vary over frequency [48], but the focus of this background is on reports of tissue properties/parameters and how they vary among people and within a single person over time. A system is typically designed with a target operating frequency, but tissue properties are functions of tissue structure and composition as well as frequency. There is a spread in the values of tissue properties even at a single frequency, as noted in [48], and the goal of this section is to elucidate the causes of that spread. Figure 2.7 illustrates tissue properties that will be discussed in this work in terms of their variability and underlying causes of variability.
Dielectric properties of any material are prone to change with temperature, orientation, mixture, pressure, and molecular structure [99]. Tissue variations include changes in macroscopic tissue structure or body composition (tissue geometry, fat content, etc.), changes in water content and perfusion, and smaller-scale changes in ion or protein concentrations, cell size, and pH, among other mechanisms [100–103].

2.3.1.1 Measurements of Tissue Properties  Practical measurement issues must be considered when defining and comparing tissue properties. Although reported measurements of some tissues are available, there is a lack of in vivo measurements of human tissues due to obvious difficulties with measuring any but the most surface-level of tissues (such as the skin or tongue). Instead, tissue properties are typically estimated from measurements of animal tissue, or measurements on excised human tissue with corrections for differences in temperature [48, 104].
The series of papers by Gabriel et al. [48, 49, 60] is widely cited as a reference for tissue properties in the electromagnetics literature, along with the associated database containing model-calculated values of permittivity and conductivity of tissues at each frequency [105]. However, the tissue properties typically used represent an empirical model, fitted to literature data and measurements, and additional consideration should be given to the variability of properties at a target frequency. Gabriel et al. present a review of measured tissue properties in the literature that were used to fit their empirical model. Tissues that have been extensively studied (e.g., muscle) are likely more indicative of realistic variability among samples, but the values must be compared with attention to the measurement method and tissue sample characteristics.

Variations in reported properties can occur due to measurement uncertainty, differences in measurement methods, and sample preparation [48, 104, 106]. Measurement uncertainty is higher for higher permittivity values, because there is less change in measured reflection coefficient for variations in material permittivity [99].

Tissue measurements in the literature are often expressed in terms of impedance (or individually resistance, capacitance, or phase angle as components of impedance). Impedance is a function of dielectric properties as well as tissue geometry, and impedance measurements will be discussed where relevant.

2.3.1.2 Tissue Variability and Mechanisms There are obvious tissue variations among different body areas, including differences in tissue thicknesses, and the presence of different tissues at the abdomen compared to the head or the eye, for example [107, 108]. Differences in tissues are also expected among patients, such as differences in fat content and muscle mass [109]. Tissue structure differences are also relevant when using animal models. As mentioned previously, due to the difficulty of measuring human tissues, tissue properties have been drawn from measurements of animal tissues, both in vivo and ex vivo. For example, the empirical model by Gabriel et al. [60] was built including measurements of animal tissue where in vivo human tissue measurements were scarce or not available.
Although differences in tissue thicknesses or shape would ideally not influence bulk tissue properties, tissue has an inhomogeneous cellular structure, and therefore changes in effective conductivity and permittivity may occur for different tissue geometries. This leads into a necessary discussion of the relationship between tissue structure and tissue dielectric properties (summarized in Table 2.2). Tissue can be classified into different types based on overall structure (e.g., skin, fat, muscle). The dielectric properties characterizing different tissue types are functions of this smaller-scale structure [110]. Schwan [111] classifies tissues into muscular or fat/bone tissue based on the ratio of water to protein content, with fat/bone tissue having a dielectric constant affected more strongly by small amounts of water and therefore being more variable from one sample to another.

Cellular variations contribute directly to tissue property variations. Tissue is inhomogeneous, and can only be approximated as homogeneous depending on the scale of interest (e.g., the electromagnetic operating frequency) [106]. Increased resistance and capacitance is associated with extracellular matrices (ECM) and additional cells [109]. At a cellular level, muscle tissue is of particular interest due to its anisotropic structure; higher resistance is measured perpendicular to muscle fibers [109]. Jilani et al. [102] saw a minimal change in resonance frequency with muscle anisotropic variation, but greater dielectric sensitivity with measurements perpendicular to muscle fibers.

Studies of tissue water content have been extensively reported [46, 50, 53, 101–103, 109, 112, 113]. As evidenced by differences in water content among tissue types and differences in measured parameters reported in [48, 49], changes in tissue water content affect tissue dielectric properties. Although more hydration without a change in ion concentration decreases conductivity, tissues with greater body water content are higher conductivity, because body water comprises ions [60, 101, 102]. Changes in moisture levels lead to shifts in resonant frequency or quality factor, which are functions of conductivity and permittivity [112].

Total body water has been observed to decrease with age, with the decrease attributed to changes in cell size, structure, water content, and free versus bound water [50, 53, 101, 104]. These mechanisms are exemplified during human brain development, where brain growth increases both the number of cells and their weight, and myelination during the first two years reduces overall ion concentration [101]. Due to age-related water loss, dielectric constant
Table 2.2: Example factors that influence tissue electromagnetic behavior.

<table>
<thead>
<tr>
<th>Cell Structure</th>
<th>Concentrations</th>
<th>Water Content</th>
<th>Pathology</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cell size, density</td>
<td>Glucose</td>
<td>Tissue ICF</td>
<td>Edema, cell</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>injury</td>
</tr>
<tr>
<td>Membrane permeability</td>
<td>Proteins</td>
<td>Age</td>
<td>Cancer</td>
</tr>
<tr>
<td>ECM</td>
<td>Ions</td>
<td>Hydration</td>
<td>Bleeding</td>
</tr>
<tr>
<td>Anisotropy</td>
<td>Hematocrit</td>
<td>Perfusion</td>
<td>Scar tissue</td>
</tr>
</tbody>
</table>

Decreases at all frequencies and loss factor increases below 4 GHz [102]. The effect of changes in body water on tissue properties appears to be higher at higher frequencies [108], as Balidemaj et al. [104] did not see a significant change in conductivity at 128 MHz in their group of 20 subjects aged 30 to 86.

Bioimpedance measurements in combination with patient body measurements (e.g., height, weight) have been used to calculate estimates of parameters such as extracellular fluid (ECF) volume or perfusion [103, 113]. Variations in impedance have been reported associated with tissue perfusion, where phase angle changes with blood volume [103]. Changes in membrane permeability affect intracellular fluid (ICF) and extracellular fluid composition, and substance concentrations in intra- and extracellular fluids contribute to tissue properties, such as blood glucose and hematocrit [103], muscle pH [102], and protein and ion concentrations [100].

Tissue properties vary with temperature, which contributes to differences between properties measured ex vivo and in vivo. Balidemaj et al. [104] extrapolated conductivity measurements at room temperature to estimate conductivity at body temperature by adjusting values 2% per degree Celsius [46]. Hyperthermia studies have investigated tissue property changes with temperature in vivo, to address how changes in tissue properties will affect electromagnetic field distributions during hyperthermia treatments [114].
Figure 2.8: Tissue parameters that have been related to increases or decreases in tissue dielectric properties of conductivity and permittivity.

Pathological changes in tissue properties have also been studied, with the goal of diagnosing conditions based on measured changes in tissue properties. Cancerous tissues have been observed to have higher conductivity [46, 104], and higher permittivity [115, 116] compared to healthy tissue. Internal bleeding can be detected by measuring tissue properties [117]. Lower impedance is associated with edema or injury due to accumulation of extracellular fluid [103, 109]. Neuromuscular diseases are correlated with higher resistivity, and reductions in phase are correlated with disease progression [103]. Higher resistance and decreased reactance in muscle is associated with atrophy (due to smaller muscle) [109]. Increases in resistance are related to epithelial cell growth, and increased reactance is related to cell mass in wound healing [103]. Variable factors that have been associated with increases or decreases in conductivity and permittivity are summarized in Figure 2.8.

2.3.2 Tissue Variations and Electromagnetic Safety

Much of the information in this section is drawn from literature on hyperthermia treatment, MRI, and measurement of tissue parameters that can indicate pathological conditions (e.g., greater water content due to swelling). Information from these studies regarding relationships between absorption/heating and tissue parameters can then be applied to evaluating
electromagnetic exposure and the safety of wireless transcutaneous systems. This section begins with a discussion of electromagnetic absorption and tissue heating mechanisms and models, followed by a discussion of the effects of tissue variability on SAR and temperature rise.

2.3.2.1 Mechanisms of Energy Absorption and Tissue Heating Absorption of electromagnetic energy in tissue can lead to tissue heating, and potentially tissue damage. Heating in wireless electromagnetic systems is most often evaluated in terms of specific absorption rate (SAR). SAR is defined according to IEEE Standard 1528-2013 [118] as in Equation 2.5, where $E$ is electric field intensity, $\rho$ is density, and $\sigma$ is conductivity. SAR can also be estimated in terms of the temperature rise over a given exposure time, as in Equation 2.6, where $T$ is temperature, $t$ is exposure time, and $c$ is specific heat.

$$SAR = \int \frac{\sigma(r)|E(r)|^2}{\rho(r)} d(r)$$  \hspace{1cm} (2.5)

$$SAR = c \frac{\Delta T}{\Delta t}$$  \hspace{1cm} (2.6)

Several organizations have established SAR guidelines for electromagnetic exposure. Because SAR represents absorption of energy per unit mass, standards specify the mass over which to calculate average SAR, where X-g SAR denotes averaging over X g of tissue mass [119]. IEEE Standard C95.1 specifies peak 10-g SAR < 2.0 W/kg and peak 1-g SAR < 1.6 W/kg [120].

Current standards require a homogeneous tissue-equivalent liquid phantom to evaluate SAR. IEEE Standard 1528-2013 recommendations for SAR evaluation are with regard to exposures for wireless communications devices [118], and [121] notes that SAR guidelines are specified in terms of far field parameters. Compared to far-field studies, there are relatively few studies on SAR and heat transfer due to near-field exposures [121].

It should be noted that SAR may not directly correspond to tissue heating [114, 122], because there can occur mm-sized “hot spots” that contribute to high local SAR, and potentially even smaller hot “nano spots” that are below the resolution of typical SAR calculations [123]. Reflections at tissue interfaces occur due to differences in intrinsic impedance.
Figure 2.9: Illustration of SAR averaging volumes, hot spots contributing to high local SAR, and reflections at tissue interfaces causing standing waves.

\[ \eta = \sqrt{\mu/\epsilon} \] [124]). These reflections can cause heating due to standing waves within tissue layers, which can be a concern particularly at higher frequencies where tissue thicknesses are larger in proportion to the wavelength. For example, Schwan [111] suggested frequencies below 1 GHz to avoid standing waves in fat thicknesses up to one inch, to achieve targeted deep tissue heating under layers of subcutaneous fat. This is in the context of hyperthermia treatment, however the concept is applicable to a situation where there is a desired region of power delivery to an implant antenna, but standing waves result in disproportionate field amplitudes and heating in surrounding tissue layers. Figure 2.9 summarizes concerns related to SAR calculations.

Thermal conduction can also contribute to tissue heating, for example through contact with the antenna, contact with implanted circuitry, or heat transfer from tissue to tissue [125]. This is not typically accounted for in SAR analyses. If waves penetrate to subcutaneous fat, the poor thermal conductivity between fat and the body surface leads to greater tissue heating [111].

Heat can be dissipated through the body’s thermoregulatory responses. Heat dissipation is less of a concern in electromagnetic systems because it counteracts temperature increases in tissue, and electromagnetic analyses are typically focused on “worst-case” tissue heating. However, thermoregulatory responses are still relevant to the present discussion on heating mechanisms. Heat loss is defined in terms of exposed body surface area - therefore depend-
ing on conditions including clothing, temperature, air movement, and humidity - and heat dissipation is more effective with shallower penetration of the EM waves (smaller skin depth) [111, 126]. Temperature increases will be greater in tissues with low blood flow; consequently, unpredictability of perfusion and other thermal parameters means that SAR does not always correlate directly with temperature [114, 122].

Heat accumulation and loss in tissue have been described in equation form: the bioheat equation describes the temperature rise in tissue as a function of heat conduction, blood perfusion, and microwave heating [125, 127, 128]. Equation 2.7, used in [127], is based on the Pennes bioheat equation [128], where \( \rho \) is density, \( c \) is specific heat capacity, \( k \) is thermal conductivity, \( \omega \) is blood perfusion rate, \( Q \) is metabolic heat generation rate, \( S \) is SAR, and the subscript \( b \) denotes a blood property.

\[
\rho c \frac{\partial T}{\partial t} = \nabla \cdot (k \nabla) + \rho Q + \rho S - \rho_b c_b \rho_b \omega (T - T_b) \tag{2.7}
\]

With reference to the heating mechanisms discussed above, it can be concluded that SAR estimated from a measured temperature rise over time (Equation 2.6) may actually underestimate the absorption of electromagnetic field energy, due to the body’s heat dissipation mechanisms \textit{in vivo}, or due to heat loss through external surfaces of an \textit{ex vivo} sample or phantom. Similarly, SAR estimated from measurements of the electric field (Equation 2.5) will depend on accuracy and homogeneity of a sample’s conductivity and density as well.

### 2.3.2.2 SAR Measurements

Wang et al. [119] provides a succinct description of SAR measurement techniques, which can be categorized into measurement of electric field or temperature increase. In the electric probe method, a robot arm moves an electric-field probe in a controlled pattern in a liquid phantom, to determine peak local and average SAR. In the thermographic method, a thermographic camera records temperature distribution in phantoms over a specified plane. This requires defining a split plane in the phantom, but is better than the electric-field probe method for measuring boundaries or solid phantoms [119].
Paulides et al. [114] used the electric probe method, scanning the 3D electric field distribution to verify simulations for hyperthermia treatments. IEEE Standard C95.3 attempts to standardize experimental SAR evaluation for human exposures [129]. The complexity of measuring electric field distributions and temperature leads most studies to estimate SAR through simulation [114, 119].

2.3.2.3 Effects of Property Variations on SAR Several studies have investigated body-scale structure variations and SAR, such as body part positions (effects of movement or distance from radiating antennas) [130], inclusion of surrounding body segments in simulation (effects of simplified simulation models) [118, 131], and fluid flow or perfusion [121]. The focus of this review is more on the effects of variability among patients and over time, including variations in dielectric properties.

The effects of dielectric property variations on SAR are not necessarily intuitive, and often complicated by other factors. Keshvari et al. [101] computed SAR with varying dielectric parameters in detailed human models and a simplified muscle model, and observed that there was not always higher absorption in tissues with higher body water, and not always higher SAR with higher conductivity and permittivity. They concluded that “SAR variation depends not only on the increase in the dielectric values but also on ... anatomical shape of the models and the ratio of dielectric values in different tissue layers” [101].

In general, it has been shown that variations in tissue properties tend to affect maximum (local) SAR more than average SAR [106, 108, 130, 131]. Gajsek et al. [106] investigated effects of published dielectric parameter variability on local and whole-body SAR. They concluded that there is no universal approach to predicting relative changes in local SAR and effects of dielectric parameter variations should be validated for each case. Keshvari et al. [101] saw greater effects of SAR averaging volume at higher frequencies (>900 MHz).

Dielectric property variations have less effect on SAR at higher frequencies. Gajsek et al. [106] saw less change in local SAR with dielectric parameter changes at higher frequencies (2 GHz), and Keshvari et al. [101] observed that SAR variation with changes in dielectric values was smaller for 1.8 GHz and 2.45 GHz compared to 900 MHz.
SAR is more affected by changes in conductivity than permittivity [106, 132]. Gajsek et al. [106] saw significant variation in localized SAR when varying muscle parameters from 0.5 to 2.0 times the parameters reported by Gabriel et al. [49], only minor changes seen with manipulating other tissue values such as fat, skin, or bone marrow. Monebhurrum et al. [132] found that maximum average simulated SAR is more sensitive to changes in conductivity than permittivity.

As stated earlier, SAR does not always directly correspond to tissue heating [114, 122]. There are relatively few studies of heat transfer specific to wireless transcutaneous systems (in comparison to studies of SAR), and even fewer on the effects of tissue property variations on tissue heating [121]. Most studies of tissue heating related to electromagnetic exposure are in the literature on hyperthermia treatment, where the goal is to selectively damage pathological tissues through targeted heating [111, 114]. Schwan [111] discussed effects of fat and skin thickness on heat absorption, concluding that absorption depends on fat thickness but that effects of skin thickness are relatively small due to skin electrical thickness below 3 GHz.

### 2.3.3 Electromagnetic Models of Tissue

Analytical, simulation, and experimental tissue models are essential for evaluating the safety of wireless transcutaneous systems in the design phase. This section reviews models used in the literature, focusing mainly on those used for evaluating wireless transcutaneous systems, but also drawing from literature on dosimetry or hyperthermia. The following is a brief review of existing models, with a discussion of which could be used to test effects of tissue variability in wireless transcutaneous systems.

In general, existing models are limited by imprecision in anatomy, tissue interfaces, and tissue properties [127]. Additionally, there are tradeoffs in modeling precision and simulation complexity, and difficulties constructing accurate experimental phantoms.

#### 2.3.3.1 Analytical Models

McAdams and Jossinet [133] provide a critical review of tissue circuit models, which is summarized here as its concepts are essential to evaluating electrical representations of tissue. Most circuit models include combinations of resistances
and capacitances to represent the tissue medium, where the component values are functions of bulk tissue properties and tissue geometry.

In the Fricke model of tissue, shown in Figure 2.10A, $R_1$ represents resistance of the suspending medium (extracellular fluid), $R_2$ represents cell interiors (intracellular fluid), and $C$ represents cell membrane capacitance [102, 133]. Intra- and extracellular fluids act as resistances due to their containing ions that conduct current and contribute to losses at higher frequencies. Cell membranes act as capacitors due to their structure as a dielectric material between two current-conducting electrolytes [102]. McAdams and Jossinet [133] argue that the Fricke model best represents suspensions of red blood cells rather than other tissues, because it lacks an explanation for constant phase angle or the known frequency dependence of tissue resistance and capacitance.

The Debye model of tissue is based on molecular-level polarization mechanisms in tissue, and describes the dielectric behavior of a dilute suspension of free dipoles. The equivalent circuit model is a capacitor in series with a resistor both in parallel with a high frequency capacitor, as shown in Figure 2.10B [133]. The Debye model has been used to fit measured tissue data and predict tissue properties across frequency [46].
The Lapicque model (Figure 2.10C) includes similar components to the previously discussed models, with \( R_1 \) representing intracellular fluid resistance, and \( R_2 \) and \( C \) representing membrane resistance and capacitance, respectively [133]. More complex tissue circuit models have been developed building on the Lapicque model, including a constant-phase-angle frequency-dependent impedance, \( Z \) (Figure 2.10C and D) [133]. With appropriate definitions of the resistances and the impedance \( Z \), these models become circuit representations of the Cole-Cole equation.

The Cole-Cole equation is an empirical model of tissue properties that provides a good fit to experimental data [60, 133]. Perhaps the most widely cited source of tissue properties, the aforementioned series of papers by Gabriel et al., fitted a Cole-Cole model to measurements of tissue properties over frequency [48, 49, 60]. However, [133] cautions against overreliance on the “relaxation” concept to relate the Cole-Cole function to the physics of tissue properties. While expressions of tissue properties as functions of frequency are common, there is a comparative lack of expressions of tissue properties as functions of tissue parameters.

2.3.3.2 Simulation Models Many simulation models of tissue have been constructed based on measurements and imaging of real tissue samples or body parts. Because tissue structure is heterogeneous, homogeneous models and phantoms have been used for reduced complexity [132, 134]. Depending on the frequency, homogeneous models may not accurately represent reflections at interfaces. This, in combination with uncertainty of dielectric properties, can lead to either under- or overestimating SAR, depending on the frequency [135]. Layered models are a step up in complexity from homogeneous models and include discrete layers of tissue types which are homogeneous within a layer, representing skin, fat, muscle, etc. The results of [101] support the use of layered models, especially at higher frequencies and when estimating SAR, because of how tissue composition affects field distribution.

More complex simulation models have been constructed based on imaging, such as CT scans or MRI. Data from the Visible Human project of the National Library of Medicine [136] was used in [106, 137], and other imaging-derived models include the Virtual Family [138, 139], and several detailed body part models [101, 108, 140]. This level of segmentation detail is most important for accurate prediction of hot spots and maximum SAR [114, 122]. There
have been more recent efforts toward semi-automatic methods of importing and updating simulation model parameters from imaging data. Paulides et al. [114, 122] used a modified semi-automatic segmentation of tissues based on CT scans, using information from previous segmentations and combining with MRI scans. Dahdouh et al. [135] built upon such work with semi-automatic methods for transferring anatomy from a detailed model to a minimal model with simplified structure, using a Euclidean distance metric to compare positions of landmarks on models.

As simulation models become increasingly accurate, they also become increasingly specific to the anatomy used to generate the model. Detailed models vary widely in dimensions [134], and abnormalities may be present in very accurate representative models [108]. Differences make comparisons difficult among studies, whether the differences are in overall geometry or limb positioning or smaller-scale structure [135].

The choice of dielectric properties is another source of variability among studies and potential inaccuracy to real operation. The choice of dielectric properties is complicated by limits on model complexity as well as uncertainty in reported dielectric property measurements. Homogeneous models typically use average dielectric parameters or parameters chosen to conservatively estimate absorption in heterogeneous tissue [108]. An effective dielectric constant can be used to represent multi-layered tissue, but it is highly dependent on tissue thicknesses and orientation [107, 117]. By this method, effective permittivity of a mixture of materials is skewed toward the material with higher permittivity [141]. Balidemaj et al. [104] note that some studies have used Cole-Cole model parameters for the bladder wall to represent the whole bladder, which does not sufficiently represent the higher conductivity of urine inside the bladder. Additionally, they indicate that measured conductivity of muscle was 14% higher than currently used in simulation models and measured bladder conductivity was an order of magnitude higher than used in many models [104]. Where human tissue properties are underreported, data from animal models are also common: Dahdouh et al. [135] used tissue properties interpolated from pigs, sheep, or rats in addition to measured properties of human tissues.
Table 2.3: Comparison of excised tissue and tissue phantoms.

<table>
<thead>
<tr>
<th>Excised Tissue</th>
<th>Phantoms</th>
</tr>
</thead>
<tbody>
<tr>
<td>+ Structural complexity</td>
<td>+ Repeatability</td>
</tr>
<tr>
<td>+ Representative properties</td>
<td>+ Controllable properties</td>
</tr>
<tr>
<td>– Sample variability</td>
<td>+ Stability</td>
</tr>
<tr>
<td>– Sample abnormalities</td>
<td>– Structurally simplistic</td>
</tr>
<tr>
<td>– Properties very sensitive to time, temperature</td>
<td>– Frequency limitations</td>
</tr>
</tbody>
</table>

2.3.3.3 Phantoms  Physical tissue models, or “phantoms”, are used for experimental validation of transcutaneous systems. In general, phantoms should represent the dielectric permittivity, conductivity and losses at the frequency of interest, in addition to preferably mimicking tissue structure and shape.

Some studies have used animal tissues (beef or pig skin, for example) to represent structural complexities of tissue. However, properties are unpredictable among samples and may vary from typical values of human tissues. Additionally, the properties of biological tissue are temperature- and time-dependent (particularly for excised tissue where the sample can dehydrate over time). An advantage of fabricated phantoms is the ability to design for properties at room temperature that mimic tissues at body temperature. A summary of excised tissue and phantom characteristics is provided in Table 2.3.

This section discusses tissue phantoms that are designed to mimic human tissue properties in vivo (and the associated phantom “recipes”) in terms of ease of use, repeatability, and representation of tissue electromagnetic behavior.

Lazebnik et al. [142] reviewed gel phantom formulations along with their valid frequency ranges, and developed and characterized formulations of a wideband gel phantom with varying amounts of oil to adjust phantom properties. The phantom conductivity and permittivity were characterized for each amount of oil in the formulation, and the phantoms properties were shown to be stable up to nine weeks after fabricating. Porter et al. [143] used these formulations to build layered phantoms with properties mimicking skin, fat, and tumor tissues in the breast. Bakar et al. [144] adjusted the Lazebnik formulation to substi-
tute formaldehyde. However, the fabrication of this phantom is somewhat complex, and the fat formulation shows less agreement with literature values [143]. Additionally, there may be concerns with the structural stability of the phantoms as the ratios of ingredients are changed.

Wang et al. [119] reviewed brain-equivalent and skull-equivalent agar phantoms. The brain-equivalent phantom comprised deionized water, agar, sodium chloride, sodium azide (as a preservative), TX-151 (for stickiness), and polyethylene powder. The skull-equivalent phantom comprised silicone emulsion, agar, glycerol (solvent), TX-151, and polyethylene powder. Polyethylene powder controls permittivity, while the salt controls conductivity, and the correct proportions of ingredients in each phantom depend on frequency.

IEEE Standard C95.3 references several formulations for muscle and other high-water-content tissues and other gel phantoms of varying viscosity, utilizing polyethylene powder to adjust permittivity and saline concentration to set conductivity [129, 145–147].

Liquid phantoms represent tissue as homogeneous. Merli [134] used homogeneous liquid phantoms with muscle-like dielectric properties. Kibret et al. [148] used a vinyl cylinder filled with saline to represent the body at low frequency (<100 MHz). Wang et al. [119] reviewed an averaged tissue equivalent phantom for the body, composed of sugar, sodium chloride, deionized water, hydroxyethyl cellulose, bactericide, diethylene glycol butyl ether, Triton X-100, diacetin, and 1,2-propanediol. The SAM phantom, defined in IEEE Standard 1528-2013 [118] and used in [149], is an anthropomorphic-shaped phantom made of a low-permittivity, low-loss plastic or fiberglass shell, filled with homogeneous tissue-equivalent liquid. Most of the rationale for the SAM phantom geometry and dimensions is based on wireless handset literature, and dimensions are selected from an anthropometric database of 1774 US Army males.

Homogeneous phantom properties must be chosen based on the heterogeneous tissue that they are to represent, and are typically chosen to give a conservative estimate of absorption compared to real tissue. However, depending on the frequency, homogeneous phantoms may not accurately model standing waves within heterogeneous tissue, which can lead to greater heating. Table 2.4 summarizes general characteristics of both gel and liquid phantoms.
Table 2.4: Comparison of gel and liquid phantoms.

<table>
<thead>
<tr>
<th>Gel</th>
<th>Liquid</th>
</tr>
</thead>
<tbody>
<tr>
<td>+ Semi-solid structure</td>
<td>+ Simple fabrication</td>
</tr>
<tr>
<td>+ Layered phantoms</td>
<td>+ Easily variable properties</td>
</tr>
<tr>
<td>– More complex fabrication, cure time</td>
<td>– Containers, compartments for heterogeneous</td>
</tr>
<tr>
<td></td>
<td>phantoms</td>
</tr>
<tr>
<td>– Adjusting properties can cause structural instability</td>
<td>– Frequency limitations</td>
</tr>
</tbody>
</table>

Phantom dimensions are an important consideration in addition to phantom structure and properties [124]. IEEE Standard 1528 addresses advantages of anthropomorphic phantoms over flat or spherical phantoms for representing antenna loading/impedance and energy coupling into the phantom.

In general, phantoms are advantageous in that they can be used to test a physical system and corroborate simulation results. However, compared to simulation models, it is relatively difficult to fabricate phantoms that are complex and precise to the desired anatomy. The need for structural complexity has motivated the use of excised tissue in experimental validation, although this strategy presents its own problems, including unpredictability, aging and temperature effects, as mentioned previously.

As noted in [118], a “90th-percentile head”, one possessing all of the 90th-percentile dimensions, does not exist. This supports the need for phantoms representing tissue variability that can be used in evaluating transcutaneous systems. A very precise phantom may be less important than a phantom that can be used to test the behavior of a system over a range of material properties. This is particularly relevant with recent research on adaptive transcutaneous systems [1].

Tissue variability and modeling is an ongoing challenge in evaluating systems for wireless transcutaneous powering. The complexity of tissue structure makes predicting tissue properties difficult, as explored in bioimpedance studies and evidenced by ranges of reported tissue properties in the literature. As such, variation in properties must be considered when
evaluating electromagnetic safety of wireless powering. Simulation models are increasingly precise and sophisticated, but can be limited by computational resources and specificity to an individual, as well as segmentation into tissue types and the choice of dielectric properties. Experimental validation depends on the use of excised tissue or fabrication of physical phantoms, which are limited by inaccuracies to human tissue properties in vivo and difficulties representing precise geometries, respectively. It is important to consider representation of tissue variability in the design and evaluation of systems for transcutaneous powering, and to either compensate (adapt) or design for robustness to variations. A summary of studies of tissue variability is presented in Table 2.5.
Table 2.5: Summary of reports and analyses of variability in tissue electromagnetic properties.

<table>
<thead>
<tr>
<th>Source</th>
<th>Tissue</th>
<th>Frequencies</th>
<th>Models/Methods</th>
</tr>
</thead>
<tbody>
<tr>
<td>[48, 49, 60]</td>
<td>30 tissues (multiple species)</td>
<td>10 Hz - 20 GHz</td>
<td>measured properties, Cole-Cole</td>
</tr>
<tr>
<td>[131]</td>
<td>eye, head (12 tissues)</td>
<td>6 - 30 GHz</td>
<td>Debye, FDTD</td>
</tr>
<tr>
<td>[113]</td>
<td>skin</td>
<td>5 - 748 kHz</td>
<td>measured properties, Cole</td>
</tr>
<tr>
<td>[116]</td>
<td>breast (6 tissues)</td>
<td>488 kHz - 1 MHz</td>
<td>measured properties, Cole, Fricke</td>
</tr>
<tr>
<td>[50]</td>
<td>10 tissues (rat)</td>
<td>130 MHz - 10 GHz</td>
<td>measured properties, Cole-Cole</td>
</tr>
<tr>
<td>[106]</td>
<td>body (39 tissues)</td>
<td>70 MHz - 2.06 GHz</td>
<td>FDTD</td>
</tr>
<tr>
<td>[132]</td>
<td>head (10 tissues and homogeneous)</td>
<td>900, 1800 MHz</td>
<td>Visible Human, FDTD</td>
</tr>
<tr>
<td>[137]</td>
<td>body</td>
<td>400, 900, 2400 MHz</td>
<td>Visible Human, FDTD</td>
</tr>
<tr>
<td>[138]</td>
<td>body (51 tissues)</td>
<td>30 MHz - 3 GHz</td>
<td>FDTD</td>
</tr>
<tr>
<td>[101]</td>
<td>eye, head (15 tissues)</td>
<td>900, 1800, 2450 MHz</td>
<td>FDTD</td>
</tr>
<tr>
<td>[53]</td>
<td>head (16 tissues, porcine properties)</td>
<td>50 MHz - 20 GHz</td>
<td>measured properties, FDTD</td>
</tr>
<tr>
<td>[108]</td>
<td>eye, head (14 tissues)</td>
<td>0.9 - 5.8 GHz</td>
<td>FDTD</td>
</tr>
<tr>
<td>[102, 112]</td>
<td>muscle (chicken)</td>
<td>500 MHz - 20 GHz</td>
<td>Fricke, FEA</td>
</tr>
<tr>
<td>[130]</td>
<td>eye (6 tissues)</td>
<td>0.9 - 10 GHz</td>
<td>FDTD</td>
</tr>
<tr>
<td>[140]</td>
<td>skin (stratum corneum, epidermis, dermis, sweat ducts)</td>
<td>0.8 - 1.2 THz</td>
<td>FEA</td>
</tr>
<tr>
<td>[104]</td>
<td>muscle, bladder, cervix</td>
<td>128 MHz</td>
<td>measured properties, Cole-Cole</td>
</tr>
<tr>
<td>[135]</td>
<td>body (14 tissues)</td>
<td>2.1, 2.6 GHz</td>
<td>FDTD</td>
</tr>
<tr>
<td>[117]</td>
<td>head</td>
<td>0.75 - 2.55 GHz</td>
<td>FEA, phantom</td>
</tr>
</tbody>
</table>
3.0 AREAS OF INVESTIGATION

3.1 TRANSMISSION MECHANISMS IN VARIABLE TISSUE

3.1.1 Motivation

Transcutaneous systems in the literature are typically designed with a particular frequency or frequency range in mind. There has been further investigation into optimal frequencies that balance power transfer and energy absorption in tissue, using loop antennas [29]. The goal of this work is to examine transmission mechanisms through tissue for a dipole system, building on prior work at low and high frequency.

3.1.2 Plan of Action

Maximum power gain will be determined through tissues with different dielectric properties at various frequencies representing a range from low to ultra-high frequency (125 kHz, 1 MHz, 13.56 MHz, 403 MHz, 915 MHz). The power gain will be calculated from simulation in ANSYS HFSS, assuming simultaneous conjugate matching. This represents best case power transfer, and serves as an initial investigation into whether certain tissue properties are ideal for transcutaneous systems. The range of tissue properties at each frequency will be chosen based on the reported values across tissue types, as given in [60].
3.2 ANTENNA TOPOLOGIES FOR VARIABLE TISSUE

3.2.1 Motivation

Loop antennas are most widely used in the implantable device literature, with power transfer occurring via inductive coupling in the near field at low frequencies. Recent literature has investigated dipole topologies for operation in proximity to metallic implants, utilizing capacitive coupling. The goal of this work is to compare the effects of tissue variability on different antenna topologies for transcutaneous systems, and to relate these effects to power transfer mechanisms in tissue.

3.2.2 Plan of Action

Planar antenna topologies will be compared in terms of power gain in a transcutaneous system and vulnerability to changes in the tissue medium. Power gain will be determined through simulation in ANSYS HFSS. The analysis will be performed with loop and dipole antenna topologies to compare the effects on primarily magnetic- and primarily electric-field power transfer. First, four antenna topologies will be simulated with varying dimensions of the implant and external antennas, to determine the dimensions that result in peak power transfer through a fixed-property tissue medium. Then, two antenna topologies will be chosen to examine the effects of tissue variations. The implantation depth, tissue layers, and tissue properties will be varied to determine the effects of tissue variability on power transfer, and the effects of impedance mismatch in a system with fixed matching networks.

3.3 TISSUE VARIABILITY AND OPTIMAL FREQUENCY

3.3.1 Motivation

Much of the literature in the area of tissue dielectric properties has been reported in the context of electromagnetic dosimetry [48], while the majority of literature on wireless transcuta-
neous devices does not address reported variations in tissue properties or how this variability affects transmission and absorption other than effects on antenna and load impedances. Adaptive designs have thus far focused on changes in antenna separation or alignment, antenna or load impedance, or resonant frequency, and less on variability in the tissue medium. Variations in tissue properties will affect the optimal frequency that maximizes power transfer within limitations on absorption in tissue. Understanding of potential changes in optimal frequency and absorption is necessary for maintaining the function of a wireless implanted device while maintaining safe energy transmission through tissue.

3.3.2 Plan of Action

Optimal frequency is defined as the transmission frequency that achieves maximum power to an implanted device within safety limitations on absorption of energy in tissue. This work will investigate the effects of tissue variations on optimal frequency in loop and dipole antenna systems, building on work using loop antennas reported in [29]. The basis for expected variations in dielectric properties and tissue structure will be drawn from existing literature on tissue properties, which has remained largely separate from the transcutaneous device literature aside from representative measured values (primarily the empirical Cole-Cole model in [60]). The effects on optimal frequency will be quantified in terms of power gain and specific absorption rate (SAR).

3.4 TISSUE PHANTOMS AND MODELS OF VARIABLE TISSUE

3.4.1 Motivation

Common practice in the wireless implantable device literature is to use simplified layered or homogeneous tissue models to design a transcutaneous system, and to choose a fixed representative tissue model. Anatomically accurate models have been developed based on imaging, but even these models are limited by specificity to the reference images used to develop the model. Therefore, variable tissue phantom representations are necessary for practical test-
ing and evaluation of transcutaneous systems and electromagnetic energy transfer through tissue, particularly at high frequencies.

### 3.4.2 Plan of Action

Based on a review of the literature on tissue properties, electromagnetic absorption, and transmission mechanisms in tissue, tissue models will be investigated experimentally and in simulation as representations of variable tissue properties. This will include fabrication of gelatinous or semisolid phantoms, which provide structural stability and allow heterogeneous phantom fabrication. Concentrations of materials in the phantom formulation will be varied to control conductivity and permittivity, and the dielectric properties of phantom formulations will be measured with a dielectric probe.

### 3.5 REPRESENTING AND SENSING VARIABLE TISSUE PROPERTIES

#### 3.5.1 Motivation

Tissue variability has been quantified in terms of measurements of conductivity and permittivity across frequency, as reported in [48, 49]. There is less documentation of variability in tissue properties as a function of smaller-scale tissue structure, and the existing literature is focused on variability in tissue hydration or fat thickness, often without relation to electromagnetic (dielectric) properties. An understanding of the mechanisms of tissue variability is essential to designing transcutaneous systems that function similarly across patients and over time.

#### 3.5.2 Plan of Action

Due to the difficulty of *in vivo* measurements and quantification of small-scale tissue components, we will utilize homogeneous and heterogeneous tissue phantoms to relate formulations and structure to bulk dielectric properties. These models of tissue structure will be investi-
gated to better represent tissue inhomogeneities that become significant at higher frequencies due to potential reflections. Structural and material parameters with the greatest effect on dielectric properties, particularly those that increase absorption, will be determined. Additionally, a method of sensing changes in tissue properties will be explored utilizing simulations and experimental phantoms.

3.6 PASSIVE IMPLANTABLE DEVICE ADAPTATION

3.6.1 Motivation

Adaptation at both the transmitter and the implant is necessary to achieve maximum efficiency in wireless transcutaneous systems. Adaptation at the receiver is limited by constraints on size and power with the use of passive implanted devices. Systems demonstrated in the literature have included on-chip tuning components, but are still limited by adaptation algorithm complexity and tuning time. There is a need for low-power control and sensing at the implanted device to achieve tuning at the implant.

3.6.2 Plan of Action

The goal of this work is to adjust the effective area of a passive implanted device, based on sensing and tuning at the receiver, processing at the transmitter, and tuning commands from the transmitter. This strategy addresses problems with performing the adaptation algorithm at the implant, which increases the power requirements of the implant. Effective area will be defined for midfield systems as the power delivered to the load relative to the power density at the receiver. Communicating a sensed parameter to the transmitter and performing any calculations externally reduces the time duration of transmitting energy through tissue to power the implant, thereby decreasing tissue absorption and heating.
4.0 TRANSMISSION MECHANISMS IN VARIABLE TISSUE

The content of this chapter has been previously published in and reprinted with permission from [150]. ©2016 IEEE. K. Bocan and E. Sejdić. Transmission mechanisms with variable tissue properties in a paired electrode system for transcutaneous power. 2016 IEEE International Symposium on Circuits and Systems (ISCAS), pages 2739–2742. DOI: https://doi.org/10.1109/ISCAS.2016.7539159

4.1 MOTIVATION

Much of the challenge of wireless powering stems from the need to deliver power through tissue. Tissue properties affect electromagnetic power transfer such that antenna systems must be specially designed to transmit effectively through tissue [46, 47, 60, 151]. Biological tissue behaves as a lossy dielectric material, in which transmission occurs through displacement current and conduction current [29, 46, 60, 151]. The contributions of displacement current and conduction current are directly related to tissue dielectric properties of conductivity and permittivity, both of which vary with frequency.

In addition to their frequency-dependence, tissue properties at a single frequency vary among locations on the body, among individuals, and over time, due to differences in cellular structure and water content [46–53]. Understanding transmission mechanisms at different frequencies and their relation to tissue dielectric properties is essential to designing efficient power delivery to an implanted device. Additionally, it is important to determine the effects of variations in tissue properties on the function of a system, in order to predict how the system will perform in real applications.
Inductively coupled loops or coils are common in implantable devices. However, the presence of metallic surfaces can interfere with the formation of magnetic fields in the configuration necessary for inductive coupling, also interfering with backscattering communication. A touch probe volume conduction system was developed by Liu et al. and demonstrated to function in proximity to metallic surfaces, for the purpose of identifying orthopedic implants with implanted radio frequency identification (RFID) tags [57, 152]. The touch probe system was designed to utilize pairs of electrodes for power transfer from the skin surface to the implant, and the term “volume conduction” was used with reference to prior work utilizing ionic conductivity of tissue for power transfer. However, since the initial work, the system was observed to function through air, indicating that ionic conduction was not the primary power transfer mechanism. The purpose of this work was to further investigate the power transfer mechanisms in this system, with broader applications to dipole and capacitively coupled transcutaneous systems.

Biological tissue is classified as a lossy dielectric, with dielectric properties arising from its fluid content and cellular structure. Tissue properties are frequency dependent, with permittivity shown to decrease and conductivity shown to increase with frequency [48, 49]. In addition, tissue properties vary with tissue structure and water content [46, 50, 53]. There is greater variation among the dielectric properties of tissue types at low frequency, and these dielectric properties are directly related to the system impedance.

Transmission through lossy dielectrics occurs through conduction current and displacement current. An electric field applied to a dielectric causes polarization, and induces displacement flux and conduction current [75]. The contributions of displacement current and conduction current vary depending on the frequency of the applied field and the properties of the medium. In tissue, conduction current dominates at low frequencies and is highly dependent on ionic conductivity. As frequency increases, the contribution of displacement current becomes more significant [29, 153]. The transmission mechanisms are directly related to tissue dielectric properties of conductivity and permittivity. Wireless systems transmitting through tissue must therefore be designed based on the operating frequency and expected tissue parameters.
4.2 METHODS

Simulations were performed to test the effects of tissue dielectric properties on transmission at frequencies covering RF bands from low frequency to ultra-high frequency: 125 kHz, 1 MHz, 13.56 MHz, 403 MHz, and 915 MHz. The simulations were performed in ANSYS HFSS, a three-dimensional electromagnetic solver utilizing the finite element method with adaptive meshing. A model was constructed in HFSS with two electrode-based antennas separated by 1 cm of tissue. This electrode-based antenna topology is of interest because such a design has been successfully implemented at 125 kHz, 13.56 MHz, and 915 MHz in previous work [154, 155]. In the model, the external antenna contacted the outer surface of the tissue, and the implanted antenna was embedded within the tissue, representing powering and communication between an external transmitter and a passive implanted receiver.

The tissue medium was a homogeneous slab defined in terms of conductivity, permittivity, and loss tangent. The values of tissue conductivity and permittivity were varied in simulation according to the range of reported physiological tissue properties, as listed in Table 4.1 [60]. Maximum power gain between the external and the implanted antennas was calculated for every combination of conductivity and permittivity as a metric of comparison among model configurations. The system was analyzed as a two-port network with ports at the external and implanted antennas as indicated in Figure 4.1, and the network scat-
Table 4.1: Ranges of tissue conductivity ($\sigma$) and relative permittivity ($\epsilon_r$) swept in simulation at each frequency ($f$)

<table>
<thead>
<tr>
<th>$f$</th>
<th>$\sigma$ (S/m)</th>
<th>$\epsilon_r$</th>
</tr>
</thead>
<tbody>
<tr>
<td>915 MHz</td>
<td>0.04 - 2.5</td>
<td>5 - 71</td>
</tr>
<tr>
<td>403 MHz</td>
<td>0.02 - 25</td>
<td>5 - 75</td>
</tr>
<tr>
<td>13.56 MHz</td>
<td>0.01 - 2.0</td>
<td>5 - 400</td>
</tr>
<tr>
<td>1 MHz</td>
<td>0.004 - 2.0</td>
<td>20 - 6000</td>
</tr>
<tr>
<td>125 kHz</td>
<td>0.02 - 2.0</td>
<td>60 - 14000</td>
</tr>
</tbody>
</table>

Scattering parameters ([S]) were used to calculate maximum power gain assuming simultaneous conjugate matching. The tissue thickness between the external and implanted antennas was kept constant while the tissue properties were varied. The dimensions of the implanted and external antennas were also unchanged when varying tissue conductivity and permittivity.

According to the complex propagation constant, the ratio of conduction current to displacement current is represented by the loss tangent, given in Equation 6.2 [75, 151]. Based on this ratio, changes in power gain were related to the contributions of conduction current and displacement current at each frequency.

$$\tan \delta = \frac{\sigma}{\omega \epsilon}$$

(4.1)

A transcutaneous system can also be modeled as a linear RLC circuit. The quality factor is then given by Equation 4.2, representing a ratio of reactive oscillatory power to dissipated power [75].

$$Q = \frac{2\pi f L}{R} = \frac{1}{2\pi f R C}$$

(4.2)
4.3 RESULTS AND DISCUSSION

Maximum power gain was calculated at each combination of conductivity and permittivity at frequencies of 915 MHz, 403 MHz, 13.56 MHz, 1 MHz, and 125 kHz. The results indicate the best-case power transfer across a range of physiological tissue properties at each frequency, that is, the power transfer assuming simultaneous conjugate matching for each configuration.

At each frequency, as the conductivity approached zero, the power gain approached one, characteristic of a lossless medium. Additionally, at each frequency there was a peak in maximum power gain at intermediate conductivity, and a plateau at higher conductivities. The peak occurred at a higher conductivity with higher permittivity. As tissue permittivity increases, the electrical size of the implanted antenna also increases. This effect corresponds to the peak gain occurring at a higher conductivity at higher permittivity.

The peak in maximum power gain was determined to be related to the loss tangent. The loss tangent represents a ratio of conductivity to permittivity, equivalently a ratio of conduction current to displacement current. The power gain at each frequency versus loss tangent is shown in Figure 4.2 and 4.3.

At all simulated frequencies, there was a region of peak maximum power gain around a loss tangent of 0.3. At lower frequencies, the maximum power gain was higher and the power gain remained high over a wider range of the loss tangent. The reported properties of skin at 915 MHz fall within the region of high power gain, indicating that 915 MHz is well-suited for powering subcutaneous implants.

Viewing the system as a series RLC circuit, the peak power gain is related to the quality factor \(Q\) according to the definition in Equation 4.2 [75]. This is a simplified view of the system as the field interactions are not directly analogous to an RLC circuit, but the model can be used to represent energy distribution in electromagnetic power transfer. High antenna \(Q\) represents greater energy in the near field. At high \(Q\), less input power is needed to achieve the same near field strength [156].

The antennas in this work are operating at a proximity such that near fields are significant in power transfer. This, along with the peak gain at high \(Q\), suggests capacitive coupling, and explains the ability of the touch probe system in [57] to function in proximity to metallic...
implants even without significant ionic conduction. Note that the loss tangent represented here is the loss tangent of the tissue medium, while the $Q$ is that including the source and load matching networks.

At loss tangent values greater than 0.3 the power gain plateaued at each frequency, and the power gain of the plateau was greater at lower frequencies. This effect indicates that losses due to conductivity are more significant than the contribution of conduction current at higher frequencies. A relatively small contribution of conduction current at higher frequency indicated that contact between the external antenna and the tissue is not essential to transmission of power. This was confirmed by adding a 1-cm air gap between the external antenna and the tissue surface and recalculating power gain with simultaneous conjugate matching (Figure 4.4).
Figure 4.3: Gain (dB) as a function of loss tangent of the tissue medium for each frequency, with color indicating quality factor including matching networks.

With the inclusion of an air gap, the power gain at 125 kHz dropped below -100 dB or required inductances greater than hundreds of mH for conjugate matching. This result suggests that conduction current is the primary transmission mechanism at 125 kHz and therefore skin contact is necessary for power transfer. At 1 MHz and 13.56 MHz, the power gain with an air gap decreased and the power gain profile across the range of loss tangents became comparable to higher frequencies, with a sharper peak around a loss tangent of 0.3.
Figure 4.4: Power gain as a function of loss tangent at 125 kHz (○), 1 MHz (+), 13.56 MHz (×), 403 MHz (△), and 915 MHz (●) with a 1-cm air gap between the external antenna and the tissue surface.

This result is indicative of displacement current. At 403 MHz and 915 MHz, the power gain was relatively unaffected by an air gap. The results suggest that contact between the external antenna and the skin may improve maximum power gain at higher frequencies, while transmission at 125 kHz is highly dependent on the interface between the external antenna and the tissue.

Implantation depth is known to affect received power, but it is not a controllable design variable, as it is typically determined by the tissue thickness at the implantation site [155]. Therefore, the tissue thickness was held constant in the current study. The tissue was also assumed homogeneous to simplify analysis of the effects of variations in dielectric properties. Because powering and communicating with an implanted device necessitates transmitting through heterogeneous tissue layers (e.g., skin, fat, and muscle) depending on the location on the body, experiments with multiple tissue layers could further explore the effects of more
complex tissue composition on power transmission. In the current study, the antenna area in contact with the tissue surface was held constant to focus on the effect of tissue properties and mechanisms at different frequencies. The antenna design was based on previous work at low and ultra-high frequency, and reflects the determination of operating frequency based on practical size constraints on the external and implanted antennas. Implantable antennas are size constrained for patient comfort and safety, and the choice of operating frequency is often a function of these constraints. Therefore, maintaining the electrical size of the antennas was neither practical nor relevant to the goals of the current study. The effect of additional antenna geometries and skin contact area within size constraints could be explored in future studies.

Compensation for environmental variations in antenna systems is being investigated through implementation of adaptive (or tunable) systems. Such systems have been demonstrated to adapt to variations in antenna positioning and variations in system impedance with the goal of maintaining power transfer or maximizing efficiency. Maximum power gains in the current study can be viewed as the ideal performance of the paired-electrode system with adaptive impedance matching in response to variations in tissue properties, where the system achieves simultaneous conjugate matching for each set of properties. The results also indicate the limitations of power transfer through variable tissue even with adaptive impedance matching.

4.4 CONCLUSION

Tissue properties vary with frequency and tissue structure, and the effect of these variations must be considered when designing systems to transmit through tissue. The purpose of this work was to investigate transmission mechanisms through tissue in a touch probe system at various frequencies and to study the effects of varying tissue properties on power transmission. A touch probe antenna system was simulated at 125 kHz, 1 MHz, 13.56 MHz, 403 MHz, and 915 MHz, and maximum power gain was calculated across the range of physiological tissue dielectric properties at each frequency. The results indicate: (1) lower frequencies
depend more on conduction current and therefore a good interface with the tissue surface, and (2) a higher $Q$ and therefore higher achievable gain is possible for certain configurations of tissue properties and frequency. The results of this study provide insight into the mechanisms of transmission through tissue at various frequencies and tissue environments for dipole and capacitively coupled transcutaneous systems.
5.0 ANTENNA TOPOLOGIES AND VARIABLE TISSUE


5.1 MOTIVATION

Passive wireless implantable medical devices provide opportunities for improved patient monitoring, documentation, and treatment. Powered transcutaneously by electromagnetic waves, passive implantable devices enable miniature, battery-less implants, but consequently depend on reliable power transfer through tissue. Passive wireless implantable devices have no implanted battery, instead harvesting energy from incident electromagnetic waves, allowing dramatic miniaturization and extended lifetime of the implant [158–161]. Additionally, passive devices are particularly well-suited for periodic monitoring in combination with implanted biosensors, which only need to be powered when obtaining a sensor reading [162, 163]. The function of passive implantable devices is dependent on the generation of electromagnetic fields at the implantation site and the efficient capture of energy using implantable antennas.

A practical transcutaneous system must be expected to function reliably for each patient and across various patients. However, biological tissue is a variable and often unpredictable medium for electromagnetic power transmission [48]. Depending on the application, a transcutaneous system may need to function at different locations in the body where there are differences in tissue structure (e.g., reading from implanted biosensors at various body loca-
tions). For example, skin and fat will be encountered for subcutaneous implants in the leg and the arm, while power transfer through the skull must be considered for cortical implants. Even the same body area is expected to have different composition among individuals, due to variations in characteristics such as body fat content and muscle size [52]. Tissue properties also vary within a single individual over time due to changes in body fat and fluid content with age or behavior [53, 164, 165]. All such tissue variations affect electromagnetic power transfer to a passive implantable device and therefore dictate the functionality of a wireless transcutaneous system.

While previous studies have investigated the effect of tissue and tissue variations on implantable antenna properties including input impedance, they are mostly dedicated to examining the implantable antenna alone or they focus on a single antenna topology (typically loop antennas) [29, 57, 150, 152, 166–170]. In transcutaneous systems where the external antenna is in proximity to the implant, the external and implanted antenna are not isolated. IEEE Standard C95.1 defines the far field region boundary as $2D^2/\lambda$, where $D$ is the antenna dimension and $\lambda$ is the electromagnetic wavelength [120]. The field region and power transfer mechanisms of an antenna system are therefore dependent on the operating frequency, the antenna dimensions, the transmission medium, and the antenna separation distance.

When the two antennas are not isolated, they must be analyzed simultaneously to account for loading effects of the implant in addition to effects of the tissue medium [29, 150, 170]. Mark et al. performed such a two-antenna analysis, investigating variability and uncertainty in thickness and dielectric properties of tissues in the head, determining the maximum achievable gain across a frequency range of 100 MHz to several GHz using loop antennas [170]. However, maximum achievable gain assumes simultaneous conjugate matching to maximize power transfer, while the power gain of a two-antenna system is ultimately sensitive to mismatch due to changes in the system impedance [1]. It is therefore important to quantify the effects of tissue variability on a transcutaneous system, to ensure the system continues to function efficiently and safely in variable tissue environments.
The goals of this study were to compare antenna topologies in terms of transcutaneous power gain and to examine the effects of tissue variability. The choice of antenna topology is integral to the function of a system, and the optimal choice of topology can improve power transfer while also simplifying the design of impedance matching networks.

In the first part of this work, four printed antenna topologies were evaluated in terms of their function in a UHF system for transcutaneous power transfer to an implanted device: planar dipole, meandered dipole, single-turn square loop, and three-turn square loop. These topologies were chosen for their relative ease of fabrication and their potential for use in thin, miniature implantable devices.

In the second part of this study, the power gain of transcutaneous systems was calculated with varying tissue characteristics for dipole and loop antenna systems. Maximum power gain for each configuration was compared to power gain with mismatch due to tissue variability. Optimizing matching networks for a particular tissue composition mimics designing a transcutaneous system based on the expected properties of the physiological implant location. Varying the tissue composition and tissue properties then represents variations that will be encountered using such a system in practice.

5.2 METHODS

Throughout this work, the power gains of transcutaneous systems were calculated from simulations in ANSYS HFSS 15.0, a 3-D electromagnetic field solver utilizing the finite element method. Each modeled system consisted of an external antenna and an implanted antenna separated by tissue, analyzed as a two-port network.

Scattering parameters (S-parameters) obtained from simulation were used to calculate power gain with conjugate impedance matching and with impedance mismatch. Simulations were performed at 915 MHz to utilize the UHF ISM band for wireless communications and the sub-GHz range recommended for efficient midfield wireless power transfer to miniature implants [29].
The dimensions of the implant antenna in this work were constrained within 1 cm by 1 cm, and the antenna separations range from 3.5 mm to 16.1 mm. Therefore, the operating field regions in this work include both the reactive near field and radiative near field regions, and necessitate full-wave simulation [29, 120].

5.2.1 Part I: Antenna Topologies and Dimensions

In the first part of this study, the simulated tissue was a simplified layered model consisting of skin, fat, and muscle, similar to that used in [29, 170, 171]. The thickness of each layer was modeled with reference to values measured for the arm: 2.2 mm skin, 10.8 mm fat, and 35 mm muscle [164]. Tissue properties were defined according to measured values for skin, fat, and muscle [49]. For simulations in tissue, the antenna was positioned between the fat and muscle layers as it would be for a subcutaneous implant, as shown in Figure 5.1, resulting in an antenna separation of 13 mm [172].

Fabricated antennas were tested experimentally using tissue phantoms, and antenna parameters were measured using a vector network analyzer (VNA) (Agilent 8753ES S-Parameter Network Analyzer). Layered tissue phantoms were constructed according to procedures and formulations in [142, 143], with layer dielectric properties similar to human skin, fat, and muscle, and layer thicknesses based on measured values for the arm (to match the simulation model). During measurements, the implanted antenna was positioned between the layers of fat and muscle phantom, to replicate the positioning of the implant in simulation.

To first validate the simulation model, one-port simulations of implanted antennas were performed in the tissue model and in air and compared to one-port measurements on fabricated antennas. Initially examining only the implanted antenna decreased the simulation complexity and simplified the measurements necessary to verify the simulation model. Values of the input port reflection coefficient \( S_{11} \) were obtained from one-port simulations of the two simplest antenna topologies: a planar dipole and a single-turn loop, shown in Figure 5.1. The input port reflection coefficient is directly related to the input impedance of the antenna, which is a function of the antenna size and topology and the surrounding media.
Simulated and measured $S_{11}$ were therefore used to evaluate the discrepancies between the simulation model and fabricated antennas of the same topology and dimensions.

A feed gap of 1 mm was chosen for both the dipoles and the loops, due to the dimensions of standard connectors that would be used later for measurements. The dipole length and trace width and the loop size and trace width (as labeled in Figure 5.1) were varied in simulation. The same sets of antenna dimensions were simulated in tissue and air to allow use of the same fabricated antennas for measurements in tissue and air. The simulated dimensions of the dipole and loop were chosen such that the dipole length and loop perimeter extended up to at least the first expected resonance in fat, with fat having the lowest permittivity and therefore longest wavelength. The dimensions were also set according to printed circuit board manufacturing specifications and to prevent any overlap of the traces. The increment size of the swept dimensions was chosen to observe trends in $S_{11}$ over the full range of
simulated dimensions, including the expected resonances in tissue (see Appendix D). The simulated values of $S_{11}$ were compared with VNA measurements of the fabricated antennas, using layered tissue phantoms [142, 143]. Each of the fabricated geometries corresponded to one of the simulated configurations. The fabricated geometries were chosen to compare trends observed with changing each dimension in simulation. The expected resonances were in agreement between simulation and measurement to within 2 cm dipole length and 1 cm loop size, with differences attributed to phantom dielectric properties (see Appendix D).

Next, both implanted and external antennas were simulated with the tissue model as a two-antenna system and analyzed as a two-port network to compare the performance of each antenna topology. This was necessary as simulations of the implanted antenna alone do not adequately inform about the behavior of the antenna in a transcutaneous UHF system including both implanted and external antennas. The proximity of the implanted antenna in such a system affects the fields of the external antenna.

Two-antenna simulations were performed to evaluate systems of planar dipoles, meandered dipoles, single-turn loops, and three-turn loops as shown in Figure 5.1. The two-antenna simulations included an external antenna and an implanted antenna separated by layers of tissue, modeling transcutaneous power transfer to a subcutaneous implanted device. The antenna positions were as depicted in Figure 5.1: the implanted antenna was positioned under the layers of skin and fat and on top of a layer of muscle, while the external antenna was positioned on the external surface of the skin, such that the layers of skin and fat separated the two antennas.

The implanted and external antennas were of the same topology within each simulated two-antenna system. The dimensions of the antennas were varied in simulation to compare power gain across different configurations of implanted and external antenna dimensions. The feed gaps and trace widths of the antennas were chosen with the same rationale previously discussed for one-port simulations and measurements. A maximum size limit of 1 cm square was chosen for the implanted antenna, a size comparable to other works on implantable antennas with dimensions ranging from 1 mm by 1 mm to 3 cm by 2 cm [17, 20, 59, 63, 173–175]. The size constraints on an external antenna are more relaxed, but the transmitter size was kept under 3 cm square to constrain the number of simulation iterations. The tissue
geometry and composition were held constant while the antenna dimensions were varied. The following dimensions of each antenna topology were varied in simulation (as labeled in Figure 5.1): the meander height and trace width of the meander dipole, the loop size and trace width of the three-turn loop, the length and trace width of the planar dipole, and the loop size and trace width of the single-turn loop. The dimensions were swept in simulation within the previously stated size limits for the implant and the external antenna (see values in Appendix F).

The antennas were simulated at every combination of implant and external dimensions, such that a single implant size and trace width was simulated with each of the sizes and trace widths of the external antenna. This resulted in a total of 720 configurations of the planar dipole system, 360 configurations of the meandered dipole system, 360 configurations of the single-turn loop system, and 360 configurations of the three-turn loop system. More configurations were possible for the dipole due to the lack of restrictions on trace width that were necessary to prevent overlapping of traces in the meandered dipole and the loop topologies.

S-parameters were used to calculate maximum power gain for each system. The maximum power gain for each set of dimensions was calculated as the power gain assuming simultaneous conjugate matching at the source and the load (see Appendix B). The peak power gain was then determined as the greatest maximum power gain for a given topology across all of the combinations of external and implanted antenna dimensions. The peak gain represents the maximum power transfer between an external antenna of up to 3 cm square and a subcutaneous implant of up to 1 cm square, given a particular antenna topology and operating frequency.

As implementing a complete physical system requires realizable impedance matching networks, the impedance values required for conjugate matching were also considered when comparing antenna topologies.
Figure 5.2: Tissue models of five different implantation locations and antenna topologies used in simulation: (A) Tissue layers and thicknesses at each location. (B) Implant locations: under skin and fat in the arm, thigh, abdomen, and head (scalp), under the skull in the head (cortex), with the indicated tissue thicknesses and antenna positioning. (C) Single-turn loop and meandered dipole antenna topologies. The dimensions of the antennas were chosen to achieve peak gain with an implanted antenna of size less than 1 cm by 1 cm and an external antenna of size less than approximately 3 cm by 3 cm.

5.2.2 Part II: Tissue Variability

The tissue variability analysis was performed using single-turn square loops or meandered dipoles for the external and implanted antennas, based on the results of the first part of the study. The dimensions of the antennas, shown in Figure 5.2, were held constant to isolate the effects of changing the tissue characteristics.
The analysis was performed as follows: first, maximum power gain with simultaneous conjugate matching was calculated at one of five implant locations shown in Figure 5.2 (arm, thigh, scalp, cortex, or abdomen); next, the antenna system location was varied over the five locations with fixed matching networks, and the power gain at each location was calculated using the S-parameters from simulation and the impedances of the fixed matching networks.

The implant antenna was positioned under skin and fat in the arm, thigh, abdomen, and scalp locations, and under the skull in the cortex location (shown in Figure 5.2). The external antenna was always positioned in contact with the external skin surface. For each configuration, power gain without conjugate matching was compared to maximum power gain achievable with simultaneous conjugate matching. The effect of changing tissue dielectric properties was also investigated by comparing the gain at each location with dry skin and wet skin, based on reported properties [49].

Tissue thicknesses and properties were varied to cover a range of tissue locations measured in the literature representing potential locations of transcutaneous systems (see Appendix E for a discussion of tissue geometry). The tissue composition in the arm and thigh was simplified to layers of skin, fat, muscle, and bone. Tissue layers used to represent the abdomen were skin, fat, muscle, and body fluid. Tissue layers used for the head were skin, fat, skull, gray matter, and white matter. Tissue layer thicknesses were modeled according to measured values reported throughout the literature [52, 164, 176–180]. Dielectric properties of each tissue were defined according to reported values [49]. Each location and the associated tissue layer thicknesses are shown in Figure 5.2. The model of the abdomen included body fluid behind the muscle extended to terminate the model, to simulate the abdominal cavity. The model of the head included brain tissue behind the skull with white matter extended to terminate the model [181].
5.3 RESULTS

5.3.1 Part I: Antenna Topologies and Dimensions

The results of the antenna topology comparison are visually summarized in Figure 5.3. For the planar dipole, a peak power gain of -23.19 dB occurred at an implanted dipole length of 1 cm and trace width of 1 cm, and an external dipole length of 3 cm and trace width of 1 cm. For the meandered dipole, a peak power gain of -21.61 dB occurred at an implant meander height of 1 cm and trace width of 0.1 cm, and an external meander height of 1 cm and trace width of 0.5 cm.

For the single-turn loop, a peak power gain of -15.27 dB occurred at an implanted loop size of 0.8 cm and trace width of 0.1 cm, and external loop size 1 cm and trace width of 0.4 cm. The implant loop size at the point of peak power gain was slightly smaller than external, such that the loop trace was aligned along the center of the wider external loop trace. For the three-turn loop, a peak power gain of -42.68 dB occurred at an implant loop size of 0.3 cm and trace width of 0.03 cm, and external loop size of 2 cm and trace width of 0.05 cm.

Assuming a power of 1 W delivered at the source, the results imply a peak received power of 29.7 mW for single-turn loops, 4.8 mW and 6.9 mW for planar dipoles and meandered dipoles, respectively, and 53.9 µW for three-turn loops. This obviously does not account for absorption limitations and assumes a fixed operating frequency, and therefore primarily indicates how the topologies compare in terms of peak power gain and trends associated with changing dimensions.

The maximum power gain of the planar dipole increased with greater length and trace width of the external and implanted dipole. The power gain of the meandered dipole system was greatest when the external and internal dipole meander heights were equal. The power gain increased with greater trace width of both the external and implanted dipole. The meandered dipole length was a function of the trace width, so this result is analogous to the planar dipole system.
Figure 5.3: Peak power gains for each topology and associated antenna dimensions and impedances required for conjugate matching. PD = planar dipole; MD = meandered dipole; 1L = single-turn loop; 3L = three-turn loop; \( G_p \) = peak power gain; \( P_R \) = power received at the implant with 1 W delivered at the source; \( Z_S, Z_L \) = impedance looking toward the source or load required for conjugate matching; \( S/L/H_I, S/L/H_E \) = implant (I) or external (E) loop size (S), dipole length (L), or meander height (H) that resulted in peak power gain; \( T W_I, T W_E \) = implant (I) or external (E) trace width (TW) that resulted in peak power gain.

The power gain of the single-turn loop system was greatest for loops of approximately the same size, and increased with greater trace width of both the implanted and external loops. Power gain of the three-turn loop system increased with greater trace width of both the external and internal loops, and was greatest for the smallest implanted loop size and an external loop size of 2 cm (see Appendix D for more details on the results of Part I).
Figure 5.4: Power gain (G) of the single-turn loop system and meandered dipole system through each tissue location with matching networks optimized to each tissue location. Ab = Abdomen; Th = Thigh; Co = Cortex; Sc = Scalp.

5.3.2 Part II: Tissue Variability

The results of the tissue variability analysis are summarized in Figures 5.4a, 5.4b, 5.5a, and 5.5b. For both antenna systems, the highest maximum gains were possible through the scalp, and the lowest maximum gains occurred through the abdomen. For the single-turn loop system, there was a consistent trend of decreased maximum gain with increased implantation depth. For the meandered dipole system, the greatest power gains occurred through the scalp, followed by the cortex, arm, thigh, and abdomen. This corresponds to decreased gain with increased depth except for the arm and thigh. The meandered dipole was determined to be more affected than the single-turn loop by differences in the tissue geometry. This is a potential explanation for the lack of a consistent trend of decreased gain with increased depth for the dipole system.
Figure 5.5: Power gain (G) of (a) the loop system, (b) dipole system, and (c) gain difference between the loop and dipole systems, through wet skin (W) or dry skin (D) with matching networks optimized to wet skin or dry skin at each tissue location. Ab = Abdomen; Th = Thigh; Co = Cortex; Sc = Scalp. Vertical axes indicate the tissue that matching networks were optimized to; Horizontal axes indicate the tissue through which power gain was calculated.

A comparison of the loop and dipole power gain in each configuration is shown in Figure 5.6 and Figure 5.5c. The loop antenna system afforded higher maximum gain than the dipole system through all but the thickest tissue (abdomen), consistent with loop antennas being most effective in the near field and the observation that magnetic field strength decreases with greater separation of the loop antennas [166]. From the first part of this study, it was determined that increasing the size of the external loop antenna at this separation did not improve the gain, while increasing the length of the external dipole increased the gain for the same size of implanted antenna. Therefore, the dipole system presents an advantage for greater implantation depth. It is expected that further increase in the external dipole length
could allow greater implant depths, within safety limitations on tissue absorption. The gain of the loop system varied over a wider range than the dipole system when the tissue location was varied. The power gain of the dipole system was therefore more consistent through variable tissue, but was generally lower than the loop system. Similar to the results across body locations, the loop system generally provided higher gains while the dipole system provided greater consistency in the presence of dielectric property variations.

The lowest gains with non-optimal matching networks through the abdomen, arm, thigh, and scalp were seen when the matching was optimized to the cortex location. Conversely, when matching was optimized to the abdomen, arm, or thigh, the gains through the arm and thigh were greater than the gain at the cortical location even though the implantation depth is less at the cortex. These effects can be attributed to the difference in tissue composition: at the abdomen, arm, and thigh locations there were layers of skin and fat between the antennas, while at the cortex location there were layers of skin, fat, and bone.

At the greater loop separation in the abdomen, the power gain was more sensitive to changes in tissue properties, indicating that at greater implantation depths the power transfer of the loop system was not solely dependent on the magnetic field. With the dipole system, there was not a trend of increased gain sensitivity to tissue dielectric property variations with implantation depth. In fact, gain was most consistent through the abdomen, the location with the lowest achievable gain with optimal matching.

For both antenna systems, higher gains were possible through wet skin, presumably due to higher permittivity of wet skin and therefore electrically larger antennas. However, the gain was more sensitive to changes in properties when matching was optimized for wet skin, likely due to a combination of mismatch and electrically smaller antennas in dry skin.

5.4 DISCUSSION

Tissue variability is an important consideration for robust implantable medical devices utilizing transcutaneous power. Tissue structure and properties have been documented to vary among patients, across locations of the body, and over time [48, 52, 53, 164, 165]. The re-
Figure 5.6: Gain difference between the loop and dipole systems through each tissue location. Ab = Abdomen; Th = Thigh; Co = Cortex; Sc = Scalp.

Results of this study indicate that power gain is highly dependent on the antenna topology and impedance matching, including impedance mismatch due to tissue variations, and maximum power gain for a given system is determined by implantation depth, antenna size, and tissue characteristics.

For the initial antenna simulations using a planar tissue model and tissue thicknesses representing the arm, the single-turn loop showed the greatest peak power gain. However, the small real impedances required for conjugate matching could prove difficult to achieve with matching networks in a physical system. The dipole topologies offered greater real impedances for matching. Comparing the dipole topologies, the meandered dipole showed higher peak gain than the planar dipole, with similar impedances required for matching. The multiple-turn loop provided even higher impedances for conjugate matching, however
the maximum achievable gain was orders of magnitude less than that of the other topologies. In applications where the size of the external antenna is not constrained, these results show that greater power gain can be achieved for the dipole systems by increasing the external dipole length beyond the length of the implanted dipole. In the loop systems investigated in this work, further increasing the size of the external loop antenna did not afford increased power gain. Additionally, increasing the width of the planar dipole increased the power gain of the system, while increasing the meander height of the meandered dipole did not improve the gain. For an application with unconstrained external antenna dimensions, the planar dipole topology affords the greatest peak power gain for a fixed implant size. Further work includes determining the achievable power delivery within constraints on energy absorption in tissue.

The performance of each antenna topology can be explained by examining the fields of each system [56, 182]. For example, the greatest power gain for the loop systems was achieved when the magnetic field vectors were most perpendicular to the plane of the implant loop, consistent with power transfer in loop systems occurring primarily through inductive coupling. The increase in gain with both dipole length and width indicate combined radiative and reactive mechanisms of power transfer. The dipole antennas in this work couple capacitively due to their proximity, and the meandered dipoles achieve power transfer through a combination of capacitive and inductive coupling [182]. Other works have used dipoles in similar proximity, referring to the antenna systems as electrodes due to the coupling method of power transfer [57].

Optimizing to achieve maximum power gain through a particular tissue composition is a reasonable starting point for designing a transcutaneous system, but in practice the system will encounter variable tissue configurations that will affect the power delivered. Ideally, an adaptive system would be implemented to adjust the matching network to each tissue configuration, although the power gain of even an adaptive system will be subject to limits related to implantation depth and tissue characteristics [1]. Additionally, the power limitations of passive implantable devices may limit adaptive matching at the implant. It is therefore important to quantify the effects of non-optimal impedance matching due to expected environmental variations, as explored in this study.
In this study, antenna dimensions were optimized using one tissue model, and then applied to various tissue configurations to examine effects of tissue variability. The topologies used in the current study were chosen to be representative of power transfer through capacitive or inductive coupling, and the effects of tissue variability are expected to be similar for antennas utilizing similar power transfer mechanisms in the near- and mid-field.

The choice of antenna topology based on power gain through a fixed tissue structure presents its own issues in that the chosen topology may not present the same power gain advantages through another tissue configuration. That is, the choice of antenna topology (and dimensions) can be biased by the choice of tissue structure to design and evaluate potential topologies. In this work, although the loop antenna system showed the highest peak gain in the first part of the study, the meandered dipole system showed similar or higher gain for some tissue configurations and impedance mismatch. In particular, the dipole system surpassed the gain of loop antenna system for larger antenna separations. Therefore, quantifying the effects of tissue variability is not only important for a system to operate despite tissue variability among patients and over time, but also to account for differences between the design environment and the environment in practice.

In this study, changes in tissue properties and composition caused the greatest mismatch losses, while tissue thickness and geometry determined the maximum achievable gain. The gain with mismatch is expected to be lower through a given tissue composition if the matching networks have not been optimized to a similar tissue composition, as evidenced by the lower gains through subcutaneous locations when matching networks were optimized for a cortical implant. Shallow implantation depth and higher permittivity present favorable conditions for power transfer due to limited attenuation and larger electrical size of the antennas, although the results of this study suggest that designing matching networks for these conditions will lead to the system gain being more sensitive to environmental variations in practice.

Based on the tissue variability analysis in this work, the antenna topology and matching networks can be designed to achieve more consistency across tissues (with lower gain) or higher gain through certain tissues (with more gain variability), depending on the power transfer mechanisms. This is analogous to designing a wide- or narrow-band antenna. For example, if a device is expected to be used at several body locations with different tissue
compositions, and particularly through thicker tissue such as the abdomen, a dipole system is more desirable. If power delivery is to be maximized and the implantation depth is comparable to the size of the antennas, a loop antenna system is likely to be more effective.

Variations in gain due to frequency are expected to be similar to the variations in gain due to changes in tissue properties, because both frequency and tissue properties affect antenna electrical size and wavelength within the tissue. The meandered dipole is a more wideband antenna than the loop, and therefore expected to be less affected by changes in frequency as it is less affected by changes in tissue properties [182].

5.5 CONCLUSION

This study investigated the effects of varying tissue structure and properties in wireless transcutaneous systems, by evaluating maximum power gain and power gain with impedance mismatch. Four antenna topologies were first evaluated in terms of peak power gain through a given tissue configuration based on measured tissues in the arm, over a range of external and implanted antenna dimensions. A single-turn loop and a meandered dipole system were then evaluated with varying tissue structure representing different locations on the body, and varying tissue properties representing fluctuations in tissue water content. The results indicate that a single-turn loop antenna topology provides the highest peak gain with the smallest implanted antenna, while dipole systems provide higher real impedance for conjugate matching and the ability to increase gain with a larger external antenna. At close antenna separations, the loop system was shown to provide higher power gain than the meandered dipole system. At antenna separations greater than the loop dimensions, the dipole system achieved higher gain, and the power gain of the dipole system was overall less sensitive to changes in tissue structure and properties. The results suggest that through choices of matching networks and antenna topologies, a system can be designed to maximize peak power gain for a narrow range of tissue properties, or to achieve greater consistency with lower peak power gain through variable tissue.
6.0 EFFICIENCY IN VARIABLE ENVIRONMENTS


6.1 MOTIVATION

Transcutaneous power transfer enables wireless powering of implantable medical devices, allowing miniaturization of the implant and extending the implant lifetime. Electromagnetic (EM) wireless power transfer has been widely investigated and demonstrated to provide sufficient power to implanted devices at high efficiency [59, 163]. The primary challenge of EM transcutaneous power transfer is to power an implanted device within safety limits on absorption of energy in tissue.

Several standards exist for evaluating electromagnetic exposure associated with wireless powering through or near tissue [120, 129]. These standards specify safety limitations for specific absorption rate (SAR) as well as methodology and experimental phantoms for evaluating electromagnetic exposure at frequencies of interest. Dielectric properties of conductivity and permittivity describe EM behavior of tissue and associated absorption mechanisms related to ionic mobility and molecular or interfacial polarization [110, 111]. Absorption in tissue occurs through different loss mechanisms depending on the EM frequency, and the dielectric properties of the tissue are therefore frequency dependent [48, 49]. Electromagnetic losses generate heat in the tissue, which is typically estimated by SAR. Local SAR, averaged
over standard tissue volumes, represents local heating due to the EM field patterns in tissue, which depend on EM operating frequency, tissue properties and structure, and antenna dimensions, separation, and topology [3].

Efficiency metrics have been defined to evaluate transcutaneous power transfer, and to compare performance across systems [1]. Poon et al. presented a definition of efficiency as received power relative to absorbed power, and derived an optimal frequency that maximized efficiency to a mm-sized loop antenna implant for different tissue types (skin, fat, muscle, etc.) [29]. This definition of efficiency is distinguished from definitions of efficiency relative to system input power as it prioritizes power delivery relative to absorption in tissue, and therefore power delivery within limitations on SAR.

Even with a generalized performance metric such as efficiency, the number of variables that contribute to SAR and transcutaneous power transfer complicates transcutaneous system design. A significant concern in wireless powering systems is the sensitivity of the system to changes in the environment in terms of both safety and system performance [1]. In addition to functioning within safety limitations at the nominal environmental configuration, the system should continue to satisfy safety regulations even in the presence of environmental variability. The impact of variability will depend on the characteristics of a given system [157]. Therefore, analysis of the system sensitivity is as important to the system design as verifying performance at nominal environmental parameters.

Here, we investigate the sensitivity of two antenna systems – single-turn square loops and meandered dipoles – to changes in tissue dielectric properties, examining power gain, SAR, and power transfer efficiency. We focus on ultra-high frequency (UHF) transcutaneous systems, based on previous work indicating high efficiency in this frequency range for miniature implant antennas in weakly-coupled antenna systems, but also susceptibility to environmental variations [17, 29, 57, 184]. This work expands on other studies of environmental variability and EM transcutaneous powering (reviewed in [1, 3]) by exploring effects of antenna topology and varying tissue dielectric properties in a layered tissue model to examine effects on local SAR.
6.2 METHODS

Efficiency was investigated for two antenna systems representing power transfer through a variable tissue medium. Efficiency was determined from power gain and SAR, calculated from simulation of antenna systems and tissue in ANSYS HFSS (ANSYS Electronics Desktop 2017.0). Two antenna topologies were modeled in systems consisting of transmit and receive antennas: single-turn square loop antennas (loop system), and meandered dipole antennas (dipole system). The antenna dimensions in each system were optimized in previous work for maximum power gain through tissues representative of the upper arm [157]. The tissue variability model in the current work was a similar model consisting of layers of skin, fat, and muscle, with thicknesses corresponding to the upper arm [164]. In each antenna system, the external antenna was placed in contact with the external tissue surface (skin), and the implanted antenna was positioned between the layers of fat and muscle, representing a subcutaneous implant powered by an external transmitter.

Each tissue layer was modeled with dielectric properties of conductivity and permittivity. The conductivity and permittivity of each tissue layer were varied over ranges reported by Gabriel et al. [48, 49]. “Conductivity” as used here (and in [49, 60]) represents total losses due to ionic conductivity and dielectric polarization mechanisms, and therefore includes the imaginary part of complex permittivity. Consequently, “permittivity” as used here refers to relative permittivity, or the real part of complex permittivity.

To simulate a system designed to operate in a UHF frequency band, the nominal frequency was fixed at 915 MHz, and the property ranges used were those reported in [48, 49] at 1 GHz. The conductivity and permittivity of the fat and muscle layers were varied according to the minimum and maximum in [48, 49] and the nominal value predicted by the empirical Cole-Cole model [60, 105], and the skin parameters were held constant at their nominal values, as listed in Table 6.1. It is important to note that the nominal values as predicted by the model fit are not necessarily equidistant from the minimum and maximum measurement values. Percentage-wise variations of +/- 10% around the nominal model values were also included in the parametric sweeps of tissue properties.
As this was a primarily investigative study, tissue properties were varied independently to cover many combinations of permittivity and conductivity. While varying the tissue properties, the tissue layer thicknesses were kept constant and similar to the thicknesses used in previous work to optimize the dimensions of the antennas for maximum power gain (2.2 mm skin, 10.8 mm fat, 100 mm muscle to terminate the model) [157]. The effect of antenna encapsulation was also investigated by adding a 10 µm layer of parylene to the antennas in simulation [174, 175], and the tissue properties were varied over the same ranges listed in Table 6.1.

For each configuration of antennas and tissue properties, maximum power gain and maximum 1-g SAR were determined from simulation and used to calculate efficiency as a metric of comparison among configurations. Maximum power gain was calculated from the two-port scattering parameters (S-parameters) assuming simultaneous conjugate matching [75]. Maximum 1-g SAR was determined from simulation by defining port impedances as those required for simultaneous conjugate matching, and setting a power of 1 W at the input port to make a consistent comparison across configurations. Setting complex port impedance necessitates the use of power S-parameters [185, 186], which were determined from simulation. With port impedances set for conjugate matching, power gain was calculated as $|S'_{21}|^2$, where $S'_{21}$ is the power wave transmission parameter [186].

Efficiency was calculated from power gain and SAR according to Equation 6.1, where $G_{max}$ is maximum power gain, $P_s$ is input power (source power, here 1 W), $SAR_{max}$ is maximum average SAR in W/kg (here, 1-g average SAR), and $M_{avg}$ is the SAR averaging mass in kg (here, 1 g). The efficiency was calculated and compared for variations in tissue properties.
properties for the loop and dipole systems. In this work we define efficiency as relative to maximum 1-g SAR to focus on SAR “hot spots” [123]. In general, local SAR is most affected by tissue parameters at higher frequencies (in contrast to global SAR or total absorption) [3], and the use of 1-g SAR is also comparable to standards establishing limits on average SAR [120].

\[ \eta = \frac{G_{max} P_s}{SAR_{max} M_{avg}} \] (6.1)

In addition to the analysis previously described, where maximum power gain was calculated for each configuration of tissue properties, a sensitivity analysis was performed assuming fixed impedance matching to nominal tissue properties. The sensitivity analysis was performed in HFSS, including permittivity and conductivity of skin, fat, and muscle.

6.3 RESULTS

The simulated gain and maximum SAR were first plotted across all tissue variations to generally compare the two antenna systems and to examine effects of encapsulation in parylene. Figure 6.1 shows that, overall, encapsulating improves gain and reduces SAR for both antenna systems. The dipole shows larger SAR than the loop, but also higher gain for most tissue configurations, presumably due to the larger transmit antenna. It should be noted that when these dimensions were optimized for power gain in prior work, it was found that increasing the size of the external loop in this configuration does not improve power gain, and this represents a limitation of the power transfer mechanisms of the antennas [157].

Each data point in Figure 6.1 represents one configuration of tissue properties - one set of values for fat conductivity, fat permittivity, muscle conductivity, and muscle permittivity. To reduce dimensionality in visualizing the results, the conductivity and permittivity of each tissue were combined into the loss tangent \((\tan \delta)\), as defined by Equation 6.2, where \(\sigma\) is conductivity, \(f\) is frequency, \(\epsilon_r\) is relative permittivity, and \(\epsilon_0\) is free space permittivity.
The gain, maximum 1-g SAR, and efficiency (as defined in Equation 6.1) are shown for each value of muscle and fat loss tangent in Figure 6.2. The trends in gain, SAR, and efficiency were determined to be consistent for the 10% variation and the larger ranges of tissue properties, therefore only the larger ranges and nominal values are included in Figure 6.2 and subsequent visualizations. Additionally, encapsulation was determined to increase gain and decrease SAR (thereby increasing efficiency) for all cases, without affecting the overall trends with tissue properties (Figure 6.1). Therefore, only results for encapsulated antennas are included in subsequent figures.
As shown in Figure 6.2, the loop power gain showed a peak at low muscle loss tangent and an intermediate value of fat loss tangent. Examining the trends with conductivity and permittivity individually (rather than combined as the loss tangent), gain was higher with lower fat and muscle conductivity and permittivity. Lower permittivity increases loss tangent, so this is why peak gain occurred at an intermediate value of the loss tangent, corresponding to low values of both conductivity and permittivity.

SAR in the loop system depended primarily on fat properties, even though gain depended on both fat and muscle. An investigation of SAR versus fat conductivity and permittivity indicated that SAR was highest for high fat conductivity and permittivity, even though high permittivity decreases the loss tangent. This is consistent with power gain being higher for low permittivity, because there is less power absorbed in the tissue. The region of highest efficiency for the loop system corresponded to the region of highest power gain, as this fell within the region of lowest SAR.

For the dipole system, power gain was greatest for low loss tangents in both fat and muscle, and decreased with higher loss tangent in muscle, but did not steadily decrease with higher fat loss tangent. SAR in the dipole system depended on both fat and muscle loss tangent (in contrast to the loop system depending on only fat), showing peak SAR with the greatest loss tangents, but otherwise complex dependence on fat and muscle properties. The dependence on muscle properties was dependent on the relative value of the fat permittivity: for higher values of fat permittivity, the dipole showed higher SAR with higher muscle conductivity; for lower fat permittivity, SAR was more affected by muscle permittivity than conductivity, with higher SAR for lower muscle permittivity (higher muscle loss tangent). Similar to the loop, the region of high dipole efficiency corresponded to the region of highest power gain, although for the dipole system the high efficiency region did not directly correspond to the region of lowest SAR. Both systems were more affected by variations in fat properties than muscle, presumably because fat separates the antennas whereas muscle is behind the receive antenna. Even so, the dipole SAR was more affected by muscle properties than the loop system.
Figure 6.2: Maximum power gain, maximum 1-g SAR (in W/kg), and efficiency for each configuration of tissue properties, represented by the loss tangent ($\tan\delta$) of fat and muscle.
The power gain shown in Figure 6.2 assumes simultaneous conjugate matching for each set of properties. Load and source impedances required for conjugate matching are shown in Figure 6.3. Comparing load and source impedance variation, there is a greater range of impedances required for conjugate matching at the load than at the source. Comparing real and imaginary parts of impedance, the range of imaginary impedance is greater than real impedance for both the dipole and loop (except the source impedance for the unencapsulated dipole). Comparing the dipole and loop antenna systems, the loop system varies over a smaller range of impedance. Additionally, the loop requires capacitive matching impedances, whereas the dipole requires inductive matching impedances.

Encapsulating appears to slightly increase both real and imaginary impedance for the dipole. For the loop, encapsulating slightly decreases the real impedance and slightly decreases the magnitude of the imaginary impedance (to a smaller negative value). This shows that the boost in efficiency provided by encapsulation is achieved without substantial effects on antenna impedance.

Figure 6.4 shows the results of the sensitivity analysis assuming fixed impedance matching to nominal tissue properties. The results are consistent with the previous analysis assuming adaptive simultaneous conjugate matching, with fat properties having the greatest effect on efficiency. From this analysis it is evident that the skin properties affect the outputs more than the muscle properties. There is a primarily second-order dependence on fat conductivity for both the loop and dipole systems.

6.4 DISCUSSION

Efficiency is paramount in transcutaneous power transfer to maximize the power delivered to the implant within safety limitations. However, a system that is optimized for performance using nominal environmental parameters may still be sensitive to parameter variations, and therefore sensitivity analyses are integral to designing and evaluating transcutaneous powering systems.
Figure 6.3: Impedances required for simultaneous conjugate matching at the source ($Z_S$) and load ($Z_L$) for the loop and dipole systems, with or without encapsulation in parylene.

Changes in tissue properties will affect energy distribution and absorption in tissue, as well as antenna impedance, impacting power transfer and efficiency. Electrically small antennas such as those used in implantable devices tend to have large reactance and small resistance (high quality factor and consequently low bandwidth), and are therefore especially susceptible to variability. The results of this work indicate that variability in tissue properties can cause an order of magnitude change in power gain and efficiency in transcutaneous systems, where efficiency is referenced to maximum local SAR.

Conventional antenna theory addresses inductively-coupled loop systems and dipoles as radiating far-field antennas. However, in small-antenna transcutaneous systems operating at frequencies in the hundreds of MHz and above, electromagnetic power transfer mechanisms are more complex, and the application of many simplifying near- and far-field assumptions
Figure 6.4: Results of sensitivity analysis of the encapsulated loop and dipole systems, showing effects of tissue conductivity ($\sigma$) and permittivity ($\epsilon_r$) on efficiency, assuming fixed impedance matching to nominal tissue properties.

breaks down. Due to the dimensions and proximity of the antennas in this work, both antenna systems transfer power through a combination of capacitive and inductive mechanisms [157]. In the loop system, because the antenna separation is on the order of the largest antenna dimension, inductive coupling is decreased. The wide antenna traces allow improved power transfer through capacitive coupling, and the contribution of the electric field means that power transfer will be more affected by tissue dielectric properties. In the dipole system, the meandering of the dipole traces introduces more complexity into the electromagnetic near fields compared to a straight dipole. These mechanisms are crucial to our analysis of SAR and power gain in the two antenna systems in this work.
Overall, the dipole system showed higher SAR, but also achieved higher power gain and therefore higher efficiency than the loop system for the same input power. The efficiency of both antenna systems was affected by changes tissue properties, with trends in gain and SAR relating to EM fields and transmission mechanisms in each system.

The loop SAR was only dependent on fat properties, whereas SAR in the dipole system depended on both fat and muscle. The relative sizes of the dipole transmit and receive antennas were more disparate than the loop system, and the external dipole antenna was larger than the external loop, which likely caused greater field penetration into the muscle tissue. The results suggest that as EM fields permeate multiple tissue layers, SAR becomes dependent on the relative properties of the tissue layers and not only the medium separating the antennas.

The power gain in both systems generally increased with lower conductivity, as is expected with decreased loss in the tissue. Where the dipole gain was highest with low tissue loss tangents, the loop gain did not directly correspond to loss tangent and was actually highest with lower permittivity of fat and muscle. Lower permittivity increases the wavelength and decreases the electrical size of the antennas, possibly improving gain by effectively decreasing the electrical distance between the loop antennas.

Tissue configurations with high maximum SAR did not directly correspond to configurations with low power gain, indicating that gain and maximum SAR are not inversely proportional. This is consistent with the conclusion that maximum SAR depends on values and ratios of tissue properties among smaller-scale tissue structure, whereas gain is a function of all loss mechanisms in the system as well as the antenna coupling [3]. Therefore, the efficiency of the system and the effects of changes in tissue properties are not necessarily predictable solely from changes in the system gain, and it is important to also consider maximum local SAR as it is determined by tissue properties and EM field patterns.

In a practical implementation, achieving maximum power gain for each tissue configuration requires load and source impedance tuning. The matching impedances needed for the dipole are inductances on the order of nanohenries, and those for the loop are capacitances on the order of picofarads. A capacitive load impedance is more common for passively powered implants due to capacitances in energy harvesting circuitry, therefore the loop system may
be better suited for matching to such circuitry on an implant [182]. However, the greater impedance of the dipole antennas may be more practical for impedance matching.

The sensitivity analysis showed second-order dependence of efficiency on tissue properties, indicative of an optimal set of properties, and consistent with previous works on optimal frequency [29, 34]. A change in optimal frequency is manifested as a change in the system impedance, evidenced by the dual approaches of tuning frequency or impedance to track optimal system performance [1].

Different definitions of efficiency mean that care must be taken in comparing results among studies. The power gain calculated in this work represents transducer gain, and is used to calculate the power delivered to a conjugate-matched load on the receive antenna relative to the input power. Therefore, this power gain may correspond to a power efficiency calculated relative to the input power defined in other work. Additionally, because this is a two-port power gain in a coupled system, it is difficult to compare to radiation efficiency for a single antenna as reported in some work [174]. Poon et al. defined power gain similarly to this work, and showed a matched power gain on the order of 0.001 near 1 GHz [29]. The power gains in this work (0.01 maximum for the loop, 0.04 maximum for the dipole) were higher presumably due to the larger implant antenna (1 x 1 cm) and smaller antenna separation (1 cm) compared to [29] (mm-sized implant antenna, 2 cm separation).

The efficiency calculated in this work relative to SAR is comparable to the power transfer efficiency defined in [29], although our efficiency is relative to a local absorption based on the standard 1-g SAR rather than total absorption. The maximum SAR values calculated in this work are similar to values in [174], with most SAR values being lower due to the smaller size of the implant antenna (32 x 24 mm in [174] versus 10 x 10 mm in this work). More complete comparisons of work in this area would be possible with sensitivity analyses of systems with different antenna dimensions and operating frequencies.

The tissue properties in this work were varied with reference to the literature review by Gabriel et al. [48]. Animal tissue data was included in their literature review, and therefore the ranges used in this work are presumed to represent a worst-case variation in tissue properties. There is a lack of literature reporting measured variability in tissue properties among samples of human tissue, and the work by Gabriel et al. represents one of
the most comprehensive reports of measured tissue properties. However, more measurements of variability among people are needed to better understand tissue variability that would be expected in practice.

6.5 CONCLUSION

The efficiency of wireless electromagnetic power transfer is highly dependent on environmental parameters, and transcutaneous power transfer to implantable devices occurs in a dynamic biological environment. As such, it is important to quantify sensitivity of transcutaneous systems to environmental variability in addition to evaluating performance at nominal values. In this work, we investigated effects of variability in tissue dielectric properties on power gain and SAR in two weakly-coupled antenna systems: single-turn square loop and meandered dipole. The dipole system showed higher efficiency relative to maximum local SAR in tissue. The sensitivity of each system to changes in tissue properties is a function of the power transfer mechanisms of each antenna topology, and sensitivity analyses are therefore an important step in robust transcutaneous system design.
7.0 VARIABLE TISSUE MODELING AND SENSING


7.1 MOTIVATION

Wireless transcutaneous electromagnetic systems must reliably and safely transfer power through a dynamic tissue medium [1]. Electromagnetically, tissue is a lossy dielectric with electromagnetic behavior and power transmission mechanisms described by permittivity and conductivity [56, 126]. Tissue dielectric properties are functions of smaller-scale tissue structure: conductivity describes charge conduction within a material in response to an applied field, where ions act as charge carriers and produce current; permittivity describes polarization within a material in response to an applied field, where cell membranes and other interfaces lead to charge buildup and capacitance, or at higher frequencies (>100 MHz) where polar molecules (such as water) align with an applied field [46–48, 110, 126]. Electronic and atomic polarization occurs at even higher frequencies [99]. Mathematically, permittivity can be represented as a complex quantity to include effects on field amplitude and losses associated with polarization mechanisms. However, losses due to polarization and ionic mobility are often combined into a single loss term.
There are many reports of measured tissue properties, either as conductivity and permittivity or as resistance and phase angle (capacitance) [3]. These measurements are typically over a range of frequencies to examine resonance and dispersion behavior. Gabriel et al. [48, 49, 60] performed a comprehensive review of measurements and developed an empirical model that has since been widely used for nominal property values in the wireless electromagnetic powering literature. From the reported measurements of tissues, it is clear that tissue properties vary among samples even at a single frequency due to differences in structure and composition [48, 49, 104, 111]. Gabriel et al. [48] estimated a variation of of 5-10% above 100 MHz, but Paulides et al. [127] note a variation of 30% or more in breast and brain tissues, and Balidemaj et al. [104] found 14% higher conductivity in their in vivo measurements of muscle tissue. Therefore, designing a system for nominal tissue properties using the empirical model is a starting point, but additional consideration of variability is necessary for a comprehensive evaluation of the potential impacts on system performance and safety.

With the exception of measurements relative to frequency or temperature, there are relatively few reports quantifying tissue structure and composition as they correspond to measured dielectric properties [46, 47, 60, 110]. Finding literature relating tissue property measurements and tissue parameter values is difficult presumably due to the difficulty of in vivo measurements of human tissue properties, and alternately the difficulty of quantifying tissue parameters without breaking down a tissue sample into its material constituents.

General statements can be made about the relationship between dielectric properties and parameters such as water content (e.g., bone, fat, muscle, skin) or cellular structure (e.g., gray matter versus white matter, blood versus tissues) [48, 110, 126]. One of the main resources for such relationships is the bioimpedance literature, where tissue properties are used as indicators of tissue parameters such as water content [103, 111]. Schwan et al. [111] discussed a model to estimate protein-bound water from constituent dielectric properties. Schepps and Foster [46] used the Fricke mixture model and Debye functions to fit measurements of dielectric properties versus frequency for tissues of varying water content (tumor, liver, spleen, fat, and muscle).
The representation of small-scale tissue structure with bulk dielectric properties is integral to modeling heterogeneous tissue with simplified structures, using discrete layers or homogeneous models. Mixture theory defines “apparent” (or “effective”) properties of heterogeneous materials based on their components [103, 124, 141, 188]. In the context of wireless transcutaneous power transfer, the operating frequency determines the scale of interactions that can be simplified with the use of tissue layers and apparent properties, and whether homogeneous models can accurately model electromagnetic exposure metrics such as specific absorption rate (SAR) [101, 129].

Calculating and modeling apparent properties are applicable to tissue models in general, and particularly experimental validation of transcutaneous power transfer using tissue phantoms, such as those developed by Lazebnik et al. [142] and used in later work [143, 144]. Although Lazebnik et al. [142] and Porter et al. [143] presented measurements of their phantoms across frequency, and Lazebnik et al. [142] compared the phantom properties to the empirical model by Gabriel et al. [60], these previous works have not addressed apparent dielectric properties using layered phantoms.

In this work, we first represent the relationship between tissue dielectric properties and smaller-scale structure utilizing tissue phantoms, measuring complex dielectric parameters over a range of frequencies. We then investigate apparent properties of layered phantoms with varying layer thicknesses in simulation, using measured phantom properties and two antenna topologies: a square single-turn loop and a meandered dipole. Finally, we propose a method of sensing changes in the tissue medium using the same antenna that would be used to transmit power to a miniature implanted device. The overall goals of this work are to examine tissue models in the context of tissue variability, and to show how robustness to variability can be integrated into transcutaneous system design.

### 7.2 METHODS

Gelatinous phantoms with various dielectric properties were fabricated using simplified formulations based on work by Lazebnik et al. [142]. The formulations are shown in Table 7.1
Table 7.1: Ingredient ratios of each phantom, representing two formulations for each oil ratio.

<table>
<thead>
<tr>
<th>Oil Ratio</th>
<th>Water [mL]</th>
<th>Gelatin [g]</th>
<th>Oil [mL]</th>
<th>Detergent [mL]</th>
</tr>
</thead>
<tbody>
<tr>
<td>10.00%</td>
<td>85.50</td>
<td>15.30</td>
<td>10.00</td>
<td>0.56 / 5.60</td>
</tr>
<tr>
<td>15.00%</td>
<td>80.75</td>
<td>14.45</td>
<td>15.00</td>
<td>0.84 / 8.40</td>
</tr>
<tr>
<td>20.00%</td>
<td>76.00</td>
<td>13.60</td>
<td>20.00</td>
<td>1.12 / 11.20</td>
</tr>
<tr>
<td>25.00%</td>
<td>71.25</td>
<td>12.75</td>
<td>25.00</td>
<td>1.40 / 14.00</td>
</tr>
<tr>
<td>30.00%</td>
<td>66.50</td>
<td>11.90</td>
<td>30.00</td>
<td>1.68 / 16.80</td>
</tr>
<tr>
<td>50.00%</td>
<td>47.50</td>
<td>8.50</td>
<td>50.00</td>
<td>2.80 / 28.00</td>
</tr>
<tr>
<td>70.00%</td>
<td>28.50</td>
<td>5.10</td>
<td>70.00</td>
<td>3.92 / 39.20</td>
</tr>
<tr>
<td>75.00%</td>
<td>23.75</td>
<td>4.25</td>
<td>75.00</td>
<td>4.20 / 42.00</td>
</tr>
<tr>
<td>80.00%</td>
<td>19.00</td>
<td>3.40</td>
<td>80.00</td>
<td>4.48 / 44.80</td>
</tr>
</tbody>
</table>

for different ratios of oil/gelatin mixture, where the amount of detergent was also varied to examine whether this provided further control over the phantom properties. After formulating according to the procedure in Lazebnik et al. [142] and allowing the phantoms to solidify for 24 hours, the dielectric properties of each phantom were measured with a dielectric probe (SPEAG DAK 3.5, see Appendix G) across the frequency range of 400 MHz to 2 GHz to examine the ultra-high frequency (UHF) range for mid-field wireless powering [59].

The phantoms were fabricated in cylindrical “puck” shapes, of sufficient thickness to appear infinite to the dielectric probe, verified by measuring the dielectric properties with and without a short placed behind the phantom and verifying that the measured properties were the same [99].

The measured phantom properties were used to investigate apparent dielectric properties in simulation, by comparing layered and homogeneous tissue models. Antennas and tissue models of similar geometry to the measured phantoms were simulated in ANSYS HFSS (ANSYS Electronics Desktop 2017.0). Homogeneous and layered tissue was modeled with dielectric properties of each tissue layer defined in terms of conductivity and permittivity. A dipole or loop transmit antenna was placed in contact with the external tissue surface, with antenna dimensions optimized in prior work [157]. A 10 µm layer of parylene was included in the simulation to encapsulate each antenna [174, 175].
Table 7.2: Skin and fat thicknesses simulated in the layered tissue model.

<table>
<thead>
<tr>
<th>Skin [mm]</th>
<th>Fat [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.23</td>
<td>10.77</td>
</tr>
<tr>
<td>1.87</td>
<td>10.35</td>
</tr>
<tr>
<td>2.15</td>
<td>13.92</td>
</tr>
<tr>
<td>2.41</td>
<td>15.45</td>
</tr>
</tbody>
</table>

The dielectric properties of the homogeneous models and each layer of the layered tissue model were defined using measured properties of the phantoms most closely matching skin, fat, and muscle tissues, and in the layered model the thickness of skin and fat layers were varied according to values reported in [164], as listed in Table 7.2. The tissue thicknesses were varied independently over all possible combinations. The overall thickness of the model was kept constant at 25 mm by setting the appropriate muscle layer thickness behind skin and fat, to match the fabricated phantom thicknesses.

The layered and homogeneous models were compared in terms of maximum 1-g average SAR [132] and transmit antenna input impedance [107], to investigate whether layering phantoms is a feasible way of verifying antenna system performance with variable tissue properties.

To investigate a method of sensing changes in dielectric properties, the input impedance of fabricated transmit antennas was measured using a vector network analyzer (Agilent 8753ES S-parameter VNA) and the fabricated phantoms of various formulations. A differential probe was used to measure differential input impedance of fabricated single-turn loop and meandered dipole antennas when in proximity to each phantom (see Appendix H). The change in input impedance was examined as a possible method of sensing changes in properties, similar to using a dielectric resonator [107].
Figure 7.1: Measured relative permittivity ($\varepsilon_r$), conductivity ($\sigma$), and loss tangent ($\tan \delta$) of each phantom formulation compared to Cole-Cole model values for fat, muscle, and skin [60, 105]. Phantom formulations are labeled with the oil ratio [142], where an asterisk indicates the modified formulation with more detergent.

7.3 RESULTS

Measured phantom dielectric properties are shown in Figure 7.1, alongside Cole-Cole empirical model values [60, 105]. The conductivity here represents a combined loss term due to ionic mobility and polarization, and is therefore frequency dependent. Higher percent oil clearly decreases the phantom relative permittivity, and decreases the slope of conductivity versus frequency, consistent with lower polarization losses. The addition of more surfactant increases conductivity closer to the empirical model values for muscle and skin, while causing the permittivity and conductivity to vary over a smaller range across formulations.
Table 7.3: Dielectric properties of phantoms closest to muscle, dry skin, wet skin, and fat at 915 MHz

<table>
<thead>
<tr>
<th>Oil Ratio</th>
<th>Detergent [mL]</th>
<th>Relative Permittivity</th>
<th>Conductivity [S/m]</th>
<th>Tissue Match</th>
</tr>
</thead>
<tbody>
<tr>
<td>15%</td>
<td>0.84</td>
<td>53.83</td>
<td>0.701</td>
<td>Muscle</td>
</tr>
<tr>
<td>25%</td>
<td>1.40</td>
<td>46.25</td>
<td>0.741</td>
<td>Skin - Dry</td>
</tr>
<tr>
<td>30%</td>
<td>1.68</td>
<td>44.03</td>
<td>0.775</td>
<td>Skin - Wet</td>
</tr>
<tr>
<td>80%*</td>
<td>4.48</td>
<td>8.84</td>
<td>0.147</td>
<td>Fat</td>
</tr>
</tbody>
</table>

* Modified formulation with more detergent

Physically, the lower oil content phantoms were more solid than the higher oil content phantoms. The addition of more detergent makes the phantom consistency more gelatinous, such that the higher percentage oil and detergent phantoms did not hold their shape outside of a container. However, the high percentage oil and detergent phantom properties did not match any of the empirical model properties (being high conductivity but low permittivity); all of the phantoms matching the model tissue properties held their shape, making them suitable for fabricating layered phantoms.

The properties of phantoms that were closest to empirical model values for fat (80%, original formulation), muscle (15%, modified formulation), wet skin (30%, modified formulation), and dry skin (25%, modified formulation) were used in subsequent simulations of layered and homogeneous tissue.

Comparisons of homogeneous and layered simulated tissue were first performed at a single frequency of 915 MHz, where the properties of each phantom (as used in the simulation models) are as listed in Table 7.3. As previously indicated and consistent with Figure 7.1, permittivity decreases with greater percent oil, while conductivity is not proportional to percent oil due to the effect of the formulation on the slope of conductivity versus frequency.

Maximum 1-g average SAR at 915 MHz is shown in Figure 7.2a for homogeneous and layered simulation models. Fat thickness was varied over a wider range than skin thickness, so the four clusters of points in each of the plots represent increasing fat thickness, and the four points within the cluster represent increasing skin thickness. SAR for the loop tends to
Figure 7.2: SAR and impedance of the loop and dipole antennas in proximity to layered models and homogeneous models with properties of phantoms closest to skin, fat, and muscle at 915 MHz.

SAR increase with skin thickness and is relatively independent of fat thickness. SAR for the dipole is more dependent on fat thickness, decreasing with greater fat thickness and increasing with skin thickness, particularly for thinner fat.

The input impedances of the dipole and loop antennas in proximity to each layered model and homogeneous model at 915 MHz are shown in Figure 7.2b. For the loop at this frequency, the impedance values of the layered model fall between the homogeneous models of fat and wet skin for both antennas. Lighter marker color for the layered models corresponds to greater skin thickness, therefore it appears that greater skin thickness increases real and
imaginary impedance of the loop. For the dipole, similar to the layered model SAR, the layered model impedances are clustered by fat thickness and impedance tends to increase with skin thickness. The impedance values of the dipole for the layered models do not appear to fall along the same trend as the homogeneous properties.

The trend behind the dipole layered model impedances becomes apparent when impedance is viewed across frequency, as in Figure 7.3a. Proximity to different tissue configurations shifts the frequency profile of the antenna impedance, as expected for changes in media near an antenna. Overall, the profiles of the layered models fall between the homogeneous fat model and the homogeneous models of the other tissues. The small size of the loop antenna causes the peak in impedance to fall beyond the highest frequency included in these simulations, but the behavior can be extrapolated from the simulated frequency range and the expected antenna behavior. Note that the loop input impedance is primarily inductive, while the dipole input impedance is mostly capacitive.

Because the trend of impedance with frequency is more informative than values at a single frequency, maximum SAR was also examined at two frequencies to investigate frequency-dependent behavior. Maximum 1-g average SAR values for the layered and homogeneous models are shown in Figure 7.3b at 915 MHz and 2 GHz. The layered models show good agreement with the homogeneous models of skin and muscle for the loop system, but for the dipole system SAR increases by a greater amount with frequency for the layered model than for the homogeneous models. At 2 GHz, the dipole layered model SAR is higher than that predicted by the homogeneous models.

Heatmaps of the spatial SAR distribution for each antenna and frequency are shown in Figure 7.4, for the layered model with nominal tissue thicknesses and the homogeneous model representing wet skin tissue. The wet skin homogeneous model was chosen because it has the highest conductivity of the homogeneous models, to compare to prior work that simplified heterogeneous structure to homogeneous models with high loss parameters to conservatively estimate SAR [108].

Measured differential input impedance is shown in Figure 7.5. The fabricated antennas were measured first with the antenna traces in contact with the phantom, and again with the substrate in contact with the phantom. Encapsulation has been shown to improve power
transfer and decrease SAR, and the substrate-contact condition was intended to mimic a layer of encapsulant protecting the metal traces of the antenna from contact with lossy tissue [183]. It was found that the antenna-contact configuration impedance changed more with the phantom properties, likely due to the combined effects of the antenna being closer to the tissue and the contact with lossy tissue. The loop antenna (with contact) showed smaller variation than the dipole for phantom oil ratios less than 50%, but greater impedance variations for oil ratios of greater than 50%. This is indicative of the loop antenna being more narrowband than the dipole antenna.

Figure 7.3: Simulated input impedance and maximum 1-g SAR of the loop and dipole antennas in proximity to layered and homogeneous tissue models.
7.4 DISCUSSION

Simulated and experimental tissue models are essential to evaluating wireless electromagnetic powering of implantable medical devices, where antenna systems must operate reliably in a dynamic tissue environment. Biological tissue structure is complex, and the use simplified tissue models can facilitate the design and evaluation of transcutaneous powering systems. However, the simplified tissue models must still provide an accurate representation of the environment, especially SAR and power efficiency.

In this work, apparent dielectric properties were investigated as a representation of how bulk properties change as a function of smaller-scale structure, and how various apparent tissue properties can be mimicked with different phantom formulations and layered phantoms. Phantoms were fabricated and measured dielectric properties were then used in simulations of layered and homogeneous tissue.
Figure 7.5: Measured differential input impedance of the loop and dipole antennas in proximity to phantoms of various formulations. Each antenna was either placed with the printed circuit board substrate or the antenna in contact with the phantom.

The measured phantom properties were similar to those reported by Lazebnik et al. [142]. The slope of the conductivity is indicative of the water content, because this conductivity is calculated from imaginary permittivity. Losses increase to a greater extent with frequency for higher water content phantoms (steeper slope in conductivity versus frequency). The addition of more detergent increases phantom conductivity and compresses the range of relative permittivity across varying amounts of oil [189].

While phantom permittivity can be matched to empirical model properties of muscle and skin, the phantom conductivity varies over a smaller range than actual tissue, and even the 80% oil phantom does reach the small model values of fat permittivity and conductivity.
This contributes to a higher loss tangent for the fat phantom than Gabriel’s Cole-Cole model for fat [60, 105]. However, fat tissue has been shown to vary in dielectric properties among measured samples [48]. Perhaps more importantly, the controllable phantom formulation allows representation of various tissue properties that can be used to evaluate the sensitivity of a system to changes in tissue properties.

Maximum average SAR has been used in prior work to compare homogeneous and heterogeneous tissue models [132]. Homogeneous tissue with high dielectric properties has been used to obtain a conservative estimate of SAR [108], but it has also been suggested that SAR does not necessarily scale with higher dielectric properties [101, 183]. The primary concern with modeling heterogeneous tissue as homogeneous is the potential misrepresentation of SAR due to the lack of small-scale tissue structure [123].

In this work, the maximum SAR trend with frequency for the dipole antenna differed between the homogeneous and layered models, whereas the trend for the loop layered model matched the homogeneous models for skin and muscle. The dipole antenna was larger than the loop, therefore the dipole system was presumably more dependent on deeper tissue and layer interfaces including fat and muscle. However, even though the maximum loop SAR corresponded to homogeneous models across frequencies, the SAR distribution was noticeably affected by the presence of layered tissue structure for both antennas. Examining the results of SAR across frequencies in this work indicates that the smaller-scale tissue structure is essential for predicting SAR distribution and trends in maximum SAR. The layered model simulated in this work assumed homogeneity within tissue layers, but even this simplification will not be valid at high enough frequencies.

The results of the impedance analysis in this work show that a homogeneous model designed to match input impedance of heterogeneous tissue at a given frequency may not accurately represent the full frequency profile. Therefore, a simplified homogeneous tissue that has been verified by comparing impedance to a heterogeneous model is limited in utility to a specific tissue geometry, frequency, and antenna dimensions. While this may not seem to be an issue if the analysis is performed at a fixed frequency, the frequency behavior of tissue is directly related to changes in impedance due to environmental parameters at a fixed frequency. This is illustrated by the results of the current work, comparing changes
in dipole impedance for tissue models at 915 MHz versus the impedance over a range of frequencies. This result emphasizes the importance of characterizing tissue models across a range of frequencies to ensure that they accurately model tissue behavior.

The frequency-dependent impedance profile of an antenna enables use of the antenna as a resonator for detecting changes in surrounding media [107]. Here, we focused on impedance variations due to properties of the tissue medium, however, variations in antenna impedance can also occur due to antenna positioning and other environmental changes. Impedance variations indicating environmental changes provide informative feedback in adaptive powering systems [1]. While robustness to environmental changes is desirable in a static system, in an adaptive system changes in the system parameters are not undesirable, provided they can be detected and compensated by tuning properties such as input power, impedance, or frequency. The results of this work are therefore generalizable to adaptive transcutaneous powering, using the transmit antenna as a resonator and impedance variations as feedback for an adaptive tuning scheme.

In general, narrowband antennas such as the loop will provide large changes in impedance over a small range, while wideband antennas will provide smaller changes in impedance over a wider range. The sensing capability of an antenna used as a resonator is higher when in closer proximity to the sensed media. In the context of this work, this means moving the antenna closer or in contact with the tissue surface. However, contact with lossy tissue may then decrease the effectiveness of the antenna for power transfer [183]. Therefore, if using the same antenna as a sensing resonator and as a transmit antenna for wireless powering, using the antenna substrate as a protective dielectric layer on one side of the transmit antenna can provide both sensing capability and power efficiency. Placing the antenna traces in contact with the tissue provides greater sensing capability, whereas placing the dielectric substrate layer between the antenna and the tissue improves powering efficiency. This technique is valid for symmetric antennas such as the loop or dipole presented in this work. Because the antennas in this work were optimized for power transfer, different antenna dimensions may improve the utility of the antennas for sensing.
Tissue models are essential to evaluating the performance and safety of wireless transcutaneous powering systems. Simplified tissue models can represent the complex structure of biological tissue, but the structural simplification must not misrepresent the characteristics of the tissue medium. In this work, we investigated loop and dipole antenna topologies in proximity to tissue models (both simulated and experimental), and compared heterogeneous and homogeneous tissue models in terms of input impedance and SAR. The results indicate the importance of accurately representing frequency-dependent characteristics of tissue even when performing evaluations at a single frequency, in order to accurately model changes in impedance. As frequency increases, heterogeneous tissue structures become more important to SAR calculations, and using a representative worst-case set of properties in a homogeneous model will not accurately reflect the SAR distribution in heterogeneous tissue. Additionally, the same antenna can be used to transmit power and to detect changes in tissue properties via input impedance, with attention to wide- or narrowband characteristics.
8.0 ADAPTATION OF A PASSIVE IMPLANTABLE DEVICE


8.1 MOTIVATION

Antenna design and system evaluation is a challenge in weakly-coupled wireless electromagnetic transcutaneous power transfer due to the dependence on environmental parameters and complex field patterns [1]. Comparing performance among such systems is complicated due to the number of different parameters, and several metrics have been defined and used in the literature, primarily power gain and specific absorption rate (SAR) [118, 120]. SAR and power gain have been combined into efficiencies specifically for implantable systems, where energy absorption in tissue is a limiting factor in power delivery [29].

Evaluating efficiency is necessary during the design of a system, but the evaluation is limited by the use of representative tissue parameters and phantoms [3]. Even very detailed simulation models cannot practically represent all possible variations among patients or over time. To address potentially deleterious effects of environmental variations on transcutaneous powering systems, adaptive techniques have been studied to detect and accommodate unpredictable and variable tissue properties in practice [1].

With passive implantable devices, power is harvested from an external transmitter to allow greater implant miniaturization and lifetime [163]. However, adapting a passive receiver is a challenge due to the limited power requirements. Any useful adaptive tuning method...
must consume minimal power, so that the majority of power is left for the implant to perform its intended function, and so that excess power transmitted for tuning purposes does not contribute to greater tissue energy absorption.

Communication in passive RFID systems is achieved through backscattering, where the incident wave from the transmitter is modulated in some way by the receiver, and the transmitter detects and demodulates the reflected (scattered) wave [182]. This enables passive implantable devices to communicate with the external transmitter via load modulation, often using the same frequency carrier for both powering and communication.

A linear network such as a transcutaneous powering channel can be described by scattering parameters (S-parameters). These can be either power or voltage waves depending on the mathematical definition [185, 191]. Voltage S-parameters can be measured with a vector network analyzer (VNA), which calculates the parameters by measuring incident and reflected waves at each port with known loads connected to other ports of the system.

A two-antenna system consisting of an external transmit antenna and an implanted receive antenna can be represented as a two-port network, including the properties of the tissue medium. The S-parameters of the two-port network can be used to calculate power transfer (gain) and efficiency of the antenna system. S-parameters have been proposed as a performance metric for calculating tuning parameters in adaptive systems: Waters et al. used S-parameters to evaluate their system and tune transmit power, and Chan Wai Po et al. discussed a direct calculation of system tuning parameters [14, 25, 81].

This work expands on efficiency evaluation and adaptive tuning systems by proposing a method of calculating S-parameters utilizing feedback from the implanted receiver. Using principles of network analysis and backscattering communication, we discuss a sensing and tuning method for transcutaneous powering of a passive implantable device based on calculation of the network transmission parameters. The parameters can then be used for evaluating efficiency and calculating tuning states at both ends of the two-port network.
8.2 METHODS

Scattering parameters of two-port systems were calculated from one-port reflection coefficients and switched loads at the second port, assuming a linear reciprocal system such as a transcutaneous antenna system consisting of an external transmit antenna, an implanted receive antenna, and separated by tissue.

Equation 8.1 defines the input reflection coefficient in terms of the scattering parameters of a two-port system [191]. Equation B.3 defines the load reflection coefficient $\Gamma_L$, where $Z_L$ is the load impedance, and $Z_0$ is the reference impedance. For an ideal open connection at the load ($Z_L \to \infty$), there is total reflection ($\Gamma_L = 1$); for an ideal short connection at the load ($Z_L = 0$), there is total reflection with inversion ($\Gamma_L = -1$); for the reference impedance connected to the load, there is zero reflection ($\Gamma_L = 0$).

\[
\Gamma_1 = S_{11} + \frac{S_{21}S_{12}\Gamma_L}{1 - S_{22}\Gamma_L} \quad (8.1)
\]

\[
\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (8.2)
\]

Assuming ideal open and short connections, the input reflection coefficients in each case are defined by Equations 8.3, 8.4, and 8.5.

\[
\Gamma_{1O} = \Gamma_1|_{Z_L=\infty} = S_{11} + \frac{S_{21}S_{12}}{1 - S_{22}} \quad (8.3)
\]

\[
\Gamma_{1S} = \Gamma_1|_{Z_L=0} = S_{11} - \frac{S_{21}S_{12}}{1 + S_{22}} \quad (8.4)
\]

\[
\Gamma_{1L} = \Gamma_1|_{Z_L=Z_0} = S_{11} \quad (8.5)
\]

For a reciprocal system ($S_{21} = S_{12}$), using the principles above, the two-port parameters can be estimated from a one-port reflection coefficient and three load connections: open, short, and a reference impedance.
Rearranging Equation 8.1 and substituting the definitions in Equations 8.3, 8.4, and 8.5 gives Equation 8.6.

\[
\begin{bmatrix}
S_{11} \\
S_{21} \\
S_{22}
\end{bmatrix} =
\begin{bmatrix}
1 & 0 & 0 \\
0 & 1 & \Gamma_{1S} - \Gamma_{1L} \\
0 & -1 & \Gamma_{1L} - \Gamma_{1O}
\end{bmatrix}
^{-1}
\begin{bmatrix}
\Gamma_{1L} \\
\Gamma_{1L} - \Gamma_{1S} \\
\Gamma_{1L} - \Gamma_{1O}
\end{bmatrix}
\]  

(8.6)

If the open and short connected are non-ideal – meaning they have some parasitic capacitance or inductance, respectively – we can define \(\Gamma'_{1O}\) and \(\Gamma'_{1S}\) using the non-ideal open and short reflection coefficients at the load, as shown in Equations 8.7 and 8.8. It follows that the system to calculate the two-port S-parameters using non-ideal connections is given by Equation 8.9. The non-ideal open and short reflection coefficients at the load can be calculated from characterized non-ideal open, short, and reference impedances, using Equation B.3.

\[
\Gamma'_{1O} = \Gamma_{1}|_{\Gamma_L = \Gamma_{LO}}
\]  

(8.7)

\[
\Gamma'_{1S} = \Gamma_{1}|_{\Gamma_L = \Gamma_{LS}}
\]  

(8.8)

\[
\begin{bmatrix}
S_{11} \\
S_{21} \\
S_{22}
\end{bmatrix} =
\begin{bmatrix}
1 & 0 & 0 \\
0 & -\Gamma_{LS} & (\Gamma_{1L} - \Gamma_{1S})\Gamma_{LS} \\
0 & -\Gamma_{LO} & (\Gamma_{1L} - \Gamma_{1O})\Gamma_{LO}
\end{bmatrix}
^{-1}
\begin{bmatrix}
\Gamma_{1L} \\
\Gamma_{1L} - \Gamma'_{1S} \\
\Gamma_{1L} - \Gamma'_{1O}
\end{bmatrix}
\]  

(8.9)

In all cases, the reference impedance (normalization impedance) for the S-parameters is assumed equal to the impedance used when measuring \(\Gamma_{1L}\).

Using the procedure detailed above, the two-port S-parameters \((S_{11}, S_{21}, S_{12}, S_{22})\) were calculated from measurements of the input reflection coefficients with the other port connected to an open circuit, short circuit, and reference impedance.

The calculation of S-parameters from input reflection coefficients was verified for four systems (or DUTs, devices under test): a non-zero length thru connector, a 50-ohm microstrip line, an unmatched dipole system, and a 50-ohm matched dipole system. The systems were
chosen to investigate three conditions of the transmitted power: high power transfer, low power transfer, and intermediate power transfer.

The input reflection coefficients were measured using a vector network analyzer (Agilent 8753ES S-parameter VNA). The open, short, and reference loads were the calibration loads from the VNA calibration kit (Keysight 85033D/E 3.5 mm Calibration Kit), manually connected to the second port of each network while reflection was measured at the input port. The calculated S-parameters were verified by comparing to the two-port S-parameters measured by connecting both ports of the network to the VNA.

The calculated parameters are highly dependent on accurate characterization of the open, short, and reference loads, used for the calculation in Equation 8.9. The open, short, and reference loads were characterized by measuring impedance over the same frequency range as S-parameters were calculated for the two-port systems. Characterization of the open, short, and reference loads is shown in Figure 8.1.
Figure 8.2: Measured (Meas) and calculated (Calc) S-parameters of the calibration kit non-zero-length thru (Calc +/- indicate positive or negative roots of \( S_{21}^2 \)).

8.3 RESULTS

The measured and calculated S-parameters of the thru, microstrip line, unmatched dipole, and 50-ohm dipole are shown in Figures 8.2, 8.3, 8.4, and 8.5. Because the calculation method gives \( S_{21}^2 \), it is not possible from only the one-port reflection coefficient to determine which root is the true value of \( S_{21} \) for each frequency. Both roots are included in the results plots to compare to the actual value measured with both ports of the VNA.

The calculated values of \( S_{11} \) and \( S_{21}^2 \) match closely to the measured values for all of the devices. The calculated value of \( S_{22} \) is more accurate for higher transmission (greater magnitude of \( S_{21} \)), most evident in comparing the calculations for the thru or microstrip line to the unmatched dipole system. The intermediate transmission value of the 50-ohm
Figure 8.3: Measured (Meas) and calculated (Calc) S-parameters of the 50-ohm microstrip line (Calc +/- indicate positive or negative roots of $S_{21}^2$).

dipole system is sufficient to estimate the value of $S_{22}$, as shown in Figure 8.5. For the calibration thru and microstrip line (Figures 8.2 and 8.3), $S_{11}$ and $S_{22}$ are near zero, and $|S_{21}|^2$ is near one across the measured frequency range. This is as expected for low reflection and high transmission. The microstrip line $S_{21}$ phase varies over a greater range than the calibration thru. The unmatched dipole system shows very low transmission, and therefore the calculation of $S_{22}$ is not informative. However, the low power transfer is still accurately indicated by the low calculated value of $S_{21}^2$. The 50-ohm matched dipole system shows intermediate magnitude of $S_{21}^2$, and the value of $S_{22}$ deviates more from the actual value than the other calculated parameters.
Figure 8.4: Measured (Meas) and calculated (Calc) S-parameters of the unmatched dipole system (Calc +/- indicate positive or negative roots of $S_{21}^2$).

The importance of load characterization is illustrated in Figure 8.6, where the calculation of two-port parameters is performed with and without accounting for an adapter connector between the second port of the DUT and the open, short, or reference load.

8.4 DISCUSSION

In adaptive systems, tuning of transmit power, frequency, or impedance is performed to maintain performance of a system in the presence of environmental variations that can impact the system safety and function [1]. Sensing and tuning are performed at the transmitter, receiver, or some combination of sensing and tuning at both the receiver and transmitter.
Figure 8.5: Measured (Meas) and calculated (Calc) S-parameters of the 50-ohm dipole system (Calc +/- indicate positive or negative roots of $S_{21}$).

In this work, we have described a method of calculating the two-port S-parameters of a transcutaneous antenna system, utilizing load modulation feedback from the receiver. The S-parameters can then be used to directly calculate matching impedances for maximum power transfer, or to provide feedback for adaptive tuning algorithms. Maximum power transfer in a two-port network can be achieved through simultaneous conjugate matching [75]. Required impedances at the source and load are calculated from $S_{11}$, $S_{22}$, and $S_{21}^2$ for a reciprocal network. Therefore, the method described in this work can be used to directly calculate simultaneous conjugate matching impedances, enabling rapid tuning of a system into the maximum power transfer state. Additionally, in situations where the ideal tuning state cannot be achieved (whether not achievable with practical component values or not achievable due to the design of impedance matching networks), the calculated S-parameters
Figure 8.6: Measured and calculated S-parameters of the calibration kit thru with and without correction for an adapter at the second port (Calc +/- indicate positive or negative roots of $S_{21}^2$).

can instead be used as feedback for search algorithms. This expands the versatility of the method presented here beyond tuning impedance, to providing feedback for tuning other system parameters essential to maintaining efficiency and safety, such as power or frequency.

In passively-powered systems, the receiver is limited in its power consumption due to harvesting power provided by the transmitter [182]. In implantable systems, the receiver is additionally limited in size [20, 59]. The dual challenges of low power and small size necessitate that any complex calculations are performed at the transmitter, and any tuning
circuitry at the receiver must be designed within the implant’s strict performance constraints. The method discussed in this work provides the distinct advantage that actual calculation of the S-parameters is performed using reflectometry and calculations at the transmitter; at the receiver, the only power necessary is the same as that required for load modulation, comparable to backscattering communication performed on a passive RFID tag [192]. For reading a passive RFID tag, a power in the range of 30 - 100 µW is required at the terminals of the receive antenna, assuming a 30% rectifier efficiency for 10 - 30 µW delivered to the RFID chip [182]. Many works in the transcutaneous powering literature have demonstrated this to be well within the power delivery capabilities of electromagnetic power transfer [20, 59].

The open, short, and reference impedances connected at the receiver while measuring input reflection coefficients at the transmitter must be well characterized in order to accurately back-calculate S-parameters. This characterization is expected to be performed during the design of backscattering load modulation at the receiver.

In a transcutaneous system with a passive receiver, the transmitter must provide power for the duration of tuning time. Excessive tuning time contributes to greater energy absorption and tissue heating. Direct calculation, as enabled by the method discussed here, reduces the time needed for tuning compared to methods that search over a range of configurations of an impedance network, thereby reducing the impact of the tuning algorithm on energy absorption in tissue [65, 81].

One limitation of the method detailed here is the inability to distinguish between the two roots of $S_{21}^2$. However, only $S_{21}^2$ is necessary for calculating simultaneous conjugate matching impedances, so the method as presented here is practically applicable for tuning impedances for maximum power transfer. A second limitation is that if there is minimal power transfer due to attenuation within the network, the calculation of $S_{22}$ will be inaccurate, impeding calculation of matching impedances. However, the use of $S_{21}^2$ is valid regardless of the transmission level, and can be used as a feedback performance metric. In this case, a two-stage tuning approach could use $S_{21}^2$ to perform an initial tuning search until the power transfer is sufficient to use $S_{22}$ to calculate impedances for simultaneous conjugate matching.
8.5 CONCLUSION

We have detailed a method of calculating two-port S-parameters of a transcutaneous antenna system using principles of reflectometry and backscattering communication. The determination of S-parameters enables direct calculation of impedance matching for maximum power transfer, or provides feedback as a measure of power transfer for tuning search algorithms. The method is intended for transcutaneous systems with a passively-powered implant, minimizing power consumption at the receiver by performing calculations at the transmitter, performing only load modulation at the receiver, and minimizing the transmitting time in the case of direct calculation of the tuning state. Next steps include implementing the calculation method in a full system, utilizing load modulation at an implanted passive receiver and reflectometry at an external transmitter.
9.0 RESEARCH SUMMARY

Portions of this chapter have been previously published in and reprinted in accordance with MDPI Open Access Policy from [1]. K. N. Bocan and E. Sejdić. Adaptive transcutaneous power transfer to implantable devices: A state of the art review. Sensors, 16(3), 2016. DOI: http://dx.doi.org/10.3390/s16030393

9.1 CONCLUSIONS

Miniature implantable devices are increasingly desirable to lessen the intrusiveness of the device for the patient and to reduce surgical complexity and infection risk. Electromagnetic energy transfer enables transcutaneous powering and communication with fully-implantable wireless medical devices, lessening the dependence on an implanted battery. The primary goal of transcutaneous energy transfer is to provide sufficient power to the implanted device while minimizing tissue heating due to absorbed energy. This has led to extensive research toward maximizing efficiency, through optimizing operating frequency, impedance matching and power delivery.

A significant complication of transcutaneous energy transfer is the variability of biological tissue in terms of tissue structure and electromagnetic properties. Tissue thickness and properties vary among individuals, among areas on the body and over time. Therefore, optimizing a transcutaneous system for a particular environment does not guarantee consistent operation in a real application. The goals of this research were to investigate the effects of tissue variability and to explore methods of detecting and compensating for these effects in wireless transcutaneous systems.
In this research, we showed that the effects of variable tissue on system metrics such as efficiency and SAR are a function of the antenna topologies and field patterns in a system. Both antenna systems transferred power through a combination of inductive and capacitive coupling mechanisms, and the effects of variability were dependent on the antenna topologies and dimensions. The effects of variability in tissue properties exhibited trends consistent with the existence of optimal frequencies for transmitting through lossy tissue, and emphasized the importance of frequency-dependent characteristics to modeling behavior even at a fixed frequency. We examined the relationship between tissue structure and bulk dielectric properties using different experimental phantom formulations and layered simulation models. We showed the importance of representing heterogeneous tissue at ultra-high frequencies, as homogeneous models did not consistently represent actual efficiency and SAR patterns. Using these variable tissue models, we investigated detection of variable tissue properties through impedance changes at the transmitter. We also discussed a method of calculating system parameters with load modulation feedback from a passively-powered implant. These results are directly relevant to coupled adaptive systems, where sensing and tuning at both the transmitter and receiver are necessary to track optimal performance.

Implementations of adaptive transcutaneous systems have the potential to enable safer and more reliable wireless powering and thereby reduce dependence on implanted batteries, as well as facilitating wireless and remote monitoring via implantable devices. In particular, adaptive devices that are able to compensate for differences in the antenna positioning and system impedances have the potential to provide improved readings from implantable sensors due to more stable powering. Implantable medical devices have already revolutionized treatment and improved quality of life for many patients, and adaptive functionality is positioned to be the next major advance in implantable device technology.

9.2 FUTURE WORK

The future of the field of wireless transcutaneous powering will continue to be heavily cross-disciplinary, combining research on the biological side (bioimpedance, hyperthermia, elec-
tromagnetic dosimetry) with electromagnetics (dielectric materials, antenna theory, wireless power transfer). As exemplified in this work, background in each of these areas is necessary for design and evaluation of systems for wireless transcutaneous powering.

Future research efforts are most needed in the areas of quantifying variability among tissues, as there are currently limited reports of interpatient and time-dependent in vivo variability in tissue dielectric properties. To continue to address safety issues associated with transcutaneous power, future research must continue to address operation of devices within SAR limitations in variable environments. As explored in this work, this requires sensitivity analyses of the effects of variability in tissue structure and properties on system performance metrics such as efficiency, power gain, and SAR. Such analyses should be followed by experimental validation with variable phantoms, where heterogeneous phantoms are necessary to represent behavior at higher frequencies.

The antenna topologies in this work were chosen to represent dipole and loop transmission mechanisms, with dimensions optimized for power gain. However, other antenna dimensions and topologies could likely provide the ability to discern smaller variations in the tissue, and could be optimized for efficiency with regard to reducing SAR. As illustrated in this research, the many factors that contribute to complex field patterns in weakly-coupled systems further motivates the use of sensitivity analyses when comparing systems using different antenna topologies, dimensions, and operating frequencies.

In cases where a system cannot be designed to be sufficiently insensitive to the effects of tissue variations, adaptive methods should be implemented. Although this work addressed methods of sensing variability in the environment through changes in scattering parameters, methods of detecting SAR or tissue heating would provide greater ability to dynamically evaluate transcutaneous system efficiency in terms of power delivered relative to power dissipated in the tissue.
APPENDIX A

ABBREVIATIONS

- C: Capacitance
- CMOS: Complementary metal-oxide semiconductor
- EM: Electromagnetic
- FCC: Federal Communications Commission
- ICNIRP: International Commission on Non-Ionizing Radiation Protection
- IEEE: Institute of Electrical and Electronics Engineers
- ISM band: Industrial, scientific, and medical band
- L: Inductance
- LNA: Low-noise amplifier
- MEMS: Micro electro-mechanical systems
- MICS: Medical Implant Communication Service
- PA: Power amplifier
- PLL: Phase-locked loop
- R: Resistance
- RF: Radio frequency
- Rx: Receiver
- S-parameters: Scattering parameters
- SAR: Specific absorption rate
- Tx: Transmitter
- UHF: Ultra-high frequency
• VNA: Vector network analyzer
• ZVS: Zero voltage switching
APPENDIX B

POWER GAIN CALCULATION

The equations below were used to calculate power gain assuming simultaneous conjugate matching, where $S_{11}$, $S_{21}$, $S_{12}$, and $S_{22}$ are the S-parameters from simulation [75]. Equation B.1 represents maximum power gain ($G_{\text{max}}$) with simultaneous conjugate matching at the source and load, where $\Gamma_S$ is the reflection coefficient looking toward the source and $\Gamma_L$ is the reflection coefficient looking toward the load. Equation B.2 represents the reflection coefficient looking toward the source required for simultaneous conjugate matching, where $B_1$ and $C_1$ are defined by Equations B.4 and B.6, respectively. Equation B.3 represents the reflection coefficient looking toward the load required for simultaneous conjugate matching, where $B_2$ and $C_2$ are defined by Equations B.5 and B.7, respectively. The value $\Delta$ in Equations B.4, B.5, B.6, and B.7 is defined by Equation B.8 [75].

\[
G_{\text{max}} = \frac{1}{1 - |\Gamma_S|^2} \frac{|S_{21}|^2}{1 - |S_{22}\Gamma_L|^2} 
\]

\[
\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} 
\]

\[
\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2} 
\]

\[
B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 
\]
\[ B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 \]  
\hspace{1cm} \text{(B.5)}

\[ C_1 = S_{11} - \Delta S_{22}^* \]  
\hspace{1cm} \text{(B.6)}

\[ C_2 = S_{22} - \Delta S_{11}^* \]  
\hspace{1cm} \text{(B.7)}

\[ \Delta = S_{11} S_{22} - S_{12} S_{21} \]  
\hspace{1cm} \text{(B.8)}
This appendix provides more detail on power S-parameters, as used in Chapter 6.

Scattering parameters (S-parameters) are typically referenced to a real impedance, however they can be referenced or renormalized to complex impedances as well. The reference impedance defines the impedance assumed to result in zero reflections from a port of a network. This is the impedance “connected” to that port when calculating network parameters at other ports. Often the terms “reference impedance” and “characteristic impedance” are used interchangeably when discussion S-parameters, but they represent a mathematical and physical parameter, respectively. If the reference impedance is equal to the characteristic impedance of the system, then the scattering parameter matrix represents the true scattering matrix of the system, otherwise they represent a “pseudo-scattering” matrix [193].

Both power-wave and voltage-wave S-parameters have been defined, although the definition is identical with real reference impedance [194]. Although there is some argument against using power-wave parameters due to the easier measurement of voltage-wave parameters [193], the use of power-wave parameters can be advantageous when discussing power transfer, and is necessary when using complex reference impedances [186]. The concept of zero reflection at a port terminated in the reference impedance is consistent for voltage- and power-wave parameters, although the reflected wave refers to either voltage or power waves in the corresponding case.
This appendix provides more detail on the simulations and measurements discussed in Chapter 5. Dimensions of antennas in one-port simulations and measurements are listed in Tables D1 and D2, respectively.

In simulation in air, a minimum $S_{11}$ of -20.9 dB occurred at a dipole length of 13.5 cm and trace width of 0.1 cm. In simulation in tissue, a minimum $S_{11}$ of -13.4 dB occurred at a dipole length of 6 cm and a trace width of 0.1 cm. The first resonance of a dipole is expected to occur at 0.47 wavelengths [56]. The wavelength in fat at 915 MHz is approximately 14 cm, so the resonance was expected at a dipole length of 6.6 cm. The wavelength in air at 915 MHz is approximately 33 cm, so the resonance was expected at a dipole length of 15.5 cm. The resonant length decreases as the dipole width increases, which explains the smaller resonant length in both air and tissue [56]. The measured $S_{11}$ in air showed a difference of up to 2.79 dB around the resonance point, but the resonance occurred at approximately the same dipole length as in simulation. Otherwise the measurements agreed with the simulated values to within 1 dB. The resonance points were shifted in tissue, indicating that the wavelength in the fat phantom was longer than the wavelength in the simulated fat layer, therefore the associated phantom permittivity was lower than the permittivity used in simulation. The resonance in measurement and simulation occurred within 2 cm of dipole length (Figure D1).

In simulation in air, a minimum $S_{11}$ of -9.86 dB occurred at a loop size of 7 cm and trace width of 0.01 cm. Because 7 cm was the maximum simulated loop size, the minimum does not represent the resonance point, but was used as a reference for experimental comparisons.
Table D1: Planar dipole and single-turn loop dimensions (cm) for one-port simulations in air and in tissue.

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Range</th>
<th>Increment</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Planar Dipole</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Length</td>
<td>0.5 - 14</td>
<td>0.5</td>
</tr>
<tr>
<td>Trace Width</td>
<td>0.1 - 1</td>
<td>0.1</td>
</tr>
<tr>
<td><strong>Single-turn Loop</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Size</td>
<td>1.5 - 7</td>
<td>0.5</td>
</tr>
<tr>
<td>Trace Width</td>
<td>0.01 - 0.5</td>
<td>0.02 (&lt;0.2), 0.1 (&gt;0.2)</td>
</tr>
</tbody>
</table>

In simulation in tissue, a minimum $S_{11}$ of -33.4 dB occurred at a loop size of 4 cm and trace width of 0.3 cm. The first resonance of a loop is expected to occur at a circumference of 1.2 wavelengths [56], so the resonance was expected at a loop size of approximately 4.2 cm in tissue and 9.9 cm in air. The measurements for the loop in air agreed with simulation results to within 2 dB. The resonance in simulation and measurement occurred within 1 cm loop size, although loop $S_{11}$ in simulation was considerably lower than measured (Figure D2).

Table D2: Fabricated planar dipole and single-turn loop geometries tested in air and with tissue phantoms

<table>
<thead>
<tr>
<th>Planar Dipole</th>
<th>Single-Turn Loop</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length (cm)</td>
<td>Trace Width (cm)</td>
</tr>
<tr>
<td>2</td>
<td>0.1</td>
</tr>
<tr>
<td>2</td>
<td>0.3</td>
</tr>
<tr>
<td>2</td>
<td>0.5</td>
</tr>
<tr>
<td>4</td>
<td>0.1</td>
</tr>
<tr>
<td>6</td>
<td>0.1</td>
</tr>
<tr>
<td>8</td>
<td>0.1</td>
</tr>
<tr>
<td>12</td>
<td>0.1</td>
</tr>
</tbody>
</table>
Figure D1: Simulated (o) and measured (x) S11 of dipole antenna in air and tissue. In air, a minimum $S_{11}$ of -20.9 dB occurred at a dipole length of 13.5 cm and trace width of 0.1 cm. In tissue, a minimum $S_{11}$ of -13.4 dB occurred at a dipole length of 6 cm and a trace width of 0.1 cm.
Figure D2: Simulated (o) and measured (x) $S_{11}$ of loop antenna in air and in tissue. In air, a minimum $S_{11}$ of -9.86 dB occurred at a loop size of 7 cm and trace width of 0.01 cm. In tissue, a minimum $S_{11}$ of -33.4 dB occurred at a loop size of 4 cm and trace width of 0.3 cm.
This appendix describes preliminary simulations regarding the work described in Chapter 5. Simulations were initially performed with planar tissue models and curved tissue models at each body location, to determine any effects of tissue geometry. For the single-turn loop, there was a difference of up to 0.83 dB between the power gains calculated with the planar and curved tissue models. The maximum difference occurred for the arm model. For the meandered dipole, there was a maximum difference of 3.61 dB between the power gains calculated with the planar and curved tissue models. The maximum difference occurred for the thigh model. The dipole was therefore more affected by differences in tissue geometry. Because the curved models are better representations of physical tissue geometries, only the results from the curved models were subsequently analyzed in the results presented in Chapter 5.
Table E1: Power gain of the single-turn loop system through each tissue location with planar tissue models. Bold values indicate maximum power gain (with conjugate matching).

<table>
<thead>
<tr>
<th>Power Gain through (dB):</th>
<th>Abdomen</th>
<th>Arm</th>
<th>Thigh</th>
<th>Cortex</th>
<th>Scalp</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abdomen</td>
<td>-20.24</td>
<td>-15.73</td>
<td>-14.63</td>
<td>-17.48</td>
<td>-7.54</td>
</tr>
<tr>
<td>Arm</td>
<td>-20.32</td>
<td>-15.66</td>
<td>-14.51</td>
<td>-17.00</td>
<td>-7.45</td>
</tr>
<tr>
<td>Optimized To:</td>
<td>Thigh</td>
<td>Cortex</td>
<td>Scalp</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Abdomen</td>
<td>-20.38</td>
<td>-15.68</td>
<td>-14.49</td>
<td>-16.73</td>
<td>-7.52</td>
</tr>
<tr>
<td>Arm</td>
<td>-27.78</td>
<td>-22.72</td>
<td>-21.31</td>
<td>-9.95</td>
<td>-8.48</td>
</tr>
<tr>
<td>Cortex</td>
<td>-24.44</td>
<td>-19.90</td>
<td>-18.88</td>
<td>-17.36</td>
<td>-2.16</td>
</tr>
</tbody>
</table>

Table E2: Power gain of the meandered dipole system through each tissue location with planar tissue models. Bold values indicate maximum power gain (with conjugate matching).

<table>
<thead>
<tr>
<th>Power Gain through (dB):</th>
<th>Abdomen</th>
<th>Arm</th>
<th>Thigh</th>
<th>Cortex</th>
<th>Scalp</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abdomen</td>
<td>-19.11</td>
<td>-17.46</td>
<td>-17.25</td>
<td>-18.98</td>
<td>-9.65</td>
</tr>
<tr>
<td>Arm</td>
<td>-19.12</td>
<td>-17.45</td>
<td>-17.24</td>
<td>-19.02</td>
<td>-9.67</td>
</tr>
<tr>
<td>Optimized To:</td>
<td>Thigh</td>
<td>Cortex</td>
<td>Scalp</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Abdomen</td>
<td>-19.13</td>
<td>-17.46</td>
<td>-17.23</td>
<td>-19.01</td>
<td>-9.74</td>
</tr>
</tbody>
</table>
This appendix provides detailed results of simulations optimizing antenna dimensions for power gain using tissue thicknesses representing the arm, as discussed in Chapter 5.

The range of dimensions swept in simulation for the implanted and external antennas are listed in Table F1. The maximum power gains for each configuration of antenna dimensions are shown in Figures F1, F2, F3, and F4.

Table F1: Simulated dimensions (cm) of the external and implanted antenna of each topology in two-antenna simulations.

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Implant Range</th>
<th>Increment</th>
<th>External Values</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Planar Dipole</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Length</td>
<td>0.3 - 1.0</td>
<td>0.1</td>
<td>1.0, 2.0, 3.0</td>
</tr>
<tr>
<td>Trace Width</td>
<td>0.1 - 1.0</td>
<td>0.1</td>
<td>0.1, 0.5, 1.0</td>
</tr>
<tr>
<td><strong>Meandered Dipole</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Length</td>
<td>0.62 - 1.10</td>
<td>0.12</td>
<td>1.1, 2.3, 3.5</td>
</tr>
<tr>
<td>Meander Height</td>
<td>0.3 - 1.0</td>
<td>0.1</td>
<td>1.0, 2.0, 3.0</td>
</tr>
<tr>
<td><strong>Single-turn Loop</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Size</td>
<td>0.3 - 1.0</td>
<td>0.1</td>
<td>1.0, 2.0, 3.0</td>
</tr>
<tr>
<td>Trace Width</td>
<td>0.02 - 0.10</td>
<td>0.02</td>
<td>0.1, 0.3, 0.4</td>
</tr>
<tr>
<td><strong>Three-turn Loop</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Size</td>
<td>0.3 - 1.0</td>
<td>0.1</td>
<td>1.0, 2.0, 3.0</td>
</tr>
<tr>
<td>Trace Width</td>
<td>0.01 - 0.03</td>
<td>0.005</td>
<td>0.01, 0.03, 0.05</td>
</tr>
</tbody>
</table>
Figure F1: Maximum power gain for each configuration of the planar dipole two-antenna system.
Figure F2: Maximum power gain for each configuration of the meandered dipole two-antenna system.
Figure F3: Maximum power gain for each configuration of the single-turn square loop two-antenna system.
Figure F4: Maximum power gain for each configuration of the three-turn square loop two-antenna system.
This appendix describes measurement of tissue dielectric properties using the dielectric probe, as discussed in Chapter 7. The measurement setup is shown in Figure G1. The frequency range was set to 0.4 - 2 GHz and the IF bandwidth was set to 1000 Hz.

The dielectric probe was calibrated according to the manufacturer’s instructions, using an open circuit (nothing contacting the end of the probe), short circuit (a strip of copper tape held against the end of the probe), and a known dielectric load (deionized water). The end of the probe was then placed in contact with each phantom, and the dielectric properties were measured over the frequency range of 0.4 - 2 GHz.

Sufficient thickness of the phantoms was verified by placing a short (a strip of copper) behind the phantom and verifying that the measured dielectric properties were the same with and without the presence of the short.
Figure G1: Measurement of phantom dielectric properties.
This appendix describes the antenna measurement setup for the work described in Chapter 7, and verification of the simulation setup as used in Chapter 6.

All measurements were performed with a vector network analyzer (VNA). For each set of antenna measurements, cable positions were fixed while the measurement setup was calibrated and during measurements to minimize errors due to cable movement. Measurements were performed with VNA settings recommended for passive devices: narrow IF bandwidth, high averaging, and low power. Through measurement comparisons, it was determined that the following settings provided sufficient accuracy: 1000 Hz IF bandwidth, 16 averages, 0 dB power.

Measurements were first performed connecting cables to SMA connectors that were soldered to the antenna terminals. These measurements most closely matched simulations using a wave port and including a model of the SMA connector. However, the impedances measured and simulated with this method were consistent with a single-ended measurement setup, which does not fully represent the desired differential behavior of the antennas. Additionally, the impedances in this configuration were highly dependent on the de-embed distance, which primarily affected the phase of the S-parameters.

To better model the differential antenna behavior, lumped ports were used in simulation, and a differential probe (Figure H1) was constructed to perform differential measurements using two ports of the VNA.
To perform differential measurements, the VNA was calibrated to the end of two coaxial
cables. The differential probe was then attached to the end of the cables, and port extension
was used to move the reference plane to the tips of the probe. The port extensions were
determined before attaching an antenna at the end of the probe, by increasing the port ex-
tension until the S-parameters appeared as an open circuit with only negative phase (Figure
H2a). The port extensions were: 194.88 ps (58.4 mm) for port one, and 189.87 ps (56.9 mm)
for port two. The full measurement setup is shown in Figure H2b.

The differential impedance was calculated from the two-port VNA measurements using
Equation H.1, where $R_0$ was set to the reference impedance of 50 ohms [195, 196].

$$Z_{in} = 2R_0 \frac{(1 - S_{12})(1 - S_{21}) - S_{11}S_{22}}{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}$$  \hspace{1cm} (H.1)

A comparison of the measured differential impedance and the simulated impedance using
a lumped port is shown in Figure H3 for a dipole of length 9.9 cm and trace width 1 mm,
and in Figure H4 for a square loop of size 4 cm and trace width 3 mm.
Figure H2: Port extension and measurement of antenna differential impedance with differential probe and VNA.
Figure H3: Dipole impedance measurement verification

Figure H4: Loop impedance measurement verification
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