Physical Layer Design and Implementation of a Biometric Authentication System using Galvanic Coupling Intra-body Communication

A Thesis Presented
by
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To my son, Christopher.
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James 1:12 (NIV) - ”Blessed is the one who perseveres under trial because, having stood the test, that person will receive the crown of life that the Lord has promised to those who love him”

I first give honor to God, who has bestowed upon me the gifts and the blessings that have made the conclusion of this chapter of my life possible. I have been blessed by an amazing support system, with my wife, and mother of my child, Kego, at the forefront. Her unwavering resolve to uplift my spirits and stand by my side throughout this journey only solidifies why she is the best thing that has ever happened to me. It brings me nothing but joy to have the opportunity to give her the world she undoubtedly deserves. I would also like to thank my father, mother and sister, for providing examples of hard work and sacrifice, self-confidence and someone to set a positive example for, respectively. Lastly, but certainly not least, I would like to thank my advisor, Kaushik Chowdhury and PhD Committee, Milica Stojanovic and Christopher Yu, for believing in a kid whose goal was to be better than his environment dictated. They have helped pushed me to become a better scholar, thinker and doer.
Abstract of the Thesis

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The emerging field of Intra-Body Communication (IBC) promises innovation in both the commercial and medical domain, by initiating data transfer from within and across the body surface to specific target devices. Traditional forms of RF-based communication face drawbacks in terms of high signal absorption within tissues that can potentially result in security and privacy issues, owing to its omnidirectional radiation. Instead, we propose the use of Galvanic Coupling (GC), wherein a weak electrical current (0.5mA) is used, resulting in a significant increase in energy efficiency compared to classical RF. In this thesis, we illustrate an example use case of intra-body communication in the form of a radically different biometric authentication system. Instead of using an over-the-air channel, we demonstrate an alternate method of transmitting data securely using GC that bypasses the vulnerable RF-channel. The GC-method injects weak electrical current into human tissue, primarily propagating through the skin. The proposed design provides impermeability to malicious attacks when sending biometric data (e.g. side-channel sniffing) as the body behaves as a natural waveguide. This dissertation makes the following contributions leading to a proof-of-concept implementation for the aforementioned application: (i) The use of analytical and empirical channel modeling techniques via a multilayered tissue equivalent circuit representing the human arm-wrist-palm propagation path and an electrically equivalent synthetic tissue phantom, respectively. (ii) Utilization of empirically verified and stored channel models for system design and testing with various narrowband and ultra-wideband modulation schemes. (iii) Demonstrating the electrocardiogram (ECG) signal, as one example biometric signatures to identify an individual. (iv) Implementation of a low-power GC transceiver with optimized communication parameters with an end-to-end validation with the transmission of biological data and (v) Demonstration of the eavesdropping susceptibility of GC-signals, and similar body communication techniques, over-the-air and while in direct contact with
the medium and (v) a method for detecting channel degradation as a precursor towards adaptive modulation and transmission of IBC data using the body's naturally changing physiological states. Performance results of the GC-transceiver prototype yield a bit error rate of $10^{-6}$ with a transmit power of -2 dBm, in addition to over 7x reduction of signal radiation outside the body compared to capacitive coupling.
Chapter 1

Introduction

Healthcare today is mired in a web of costly preventive procedures available to a limited segment of society, complex insurance rules, and lifestyle choices, all of which impact timely diagnosis. In fact, healthcare expenditures in the United States have reached over $3 trillion USD a year [1], signaling an urgent need to explore new forms of medial sensor technology that can autonomously identify and react to abnormal conditions in real-time.

Traditionally, medical tests are conducted on-site at a medical facility, which may require the use of wires and leads, leading to intrusive procedures and physical discomfort. Miniaturization and advancements in MEMS technology have given rise to sensors that can be placed unobtrusively in and around the human body. The resulting connected body area network (BAN) imposes stringent design constraints that go beyond those commonly assumed for classical over the air wireless sensor networks (WSNs), particularly with respect to power consumption, size, tissue heating and electromagnetic wave absorption. Understandably, BANs have resulted in new possibilities for preventative medical treatment by recognizing trends in medical data for rapid identification of treatment options or for sending emergency alerts for critical events [2].

- Motivation for intra-body communication: Solely limiting the placement of sensors on the surface of the body (i.e., the network architecture typically assumed within BANs) limits the types of biological information that can be obtained. Furthermore, it can potentially increase patient inconvenience as the number of devices begins to scale and additional restrictions on prolonged signal exposure, heat and movement come into play [3]. As a result, we advocate for a completely new paradigm where the sensors are implanted inside the body, with direct access to the specific target regions of interest. The implant-driven architecture, shown in figure 1.1, is composed of small-scale sensors that gather biological information, actuators responsible for drug delivery functionality, and an
CHAPTER 1. INTRODUCTION

Figure 1.1: Conceptual representation of an intra-body network

Figure 1.2: Number of articles referencing intra-body communication and intra-Body networks

on-body relay node for data aggregation and communication with external networks. Communication among implants, designated as intra-body communication (IBC), can also be used to coordinate activities between different sub-sets of sensors, in terms of performing heterogeneous sensing tasks, local data aggregation and decision making by sharing computational loads.

- **Types of intra-body communication:** The most common form of an IBC link, as seen in figure 1.3, uses classical Radio Frequency (RF) in the form of narrowband (NB) or ultra-wideband (UWB) signals for implant-to-implant or implant-to-relay communication [4]. However, there are other non-traditional forms of wireless communication that have unique properties that may be more suited for IBC. Ultrasound (US), widely known in the medical community for its application in imaging, is a viable option for connecting medical implants with the use of acoustic waves well above the 20 kHz range [5]. Capacitive Coupling (CC) enables the propagation of near electric fields around and through the human body, creating links that span the length of the medium [6]. Magnetic Resonant Coupling (RC), similar to inductive coupling techniques, employs loosely coupled coils wrapped around parts of the body to transmit and receive magnetic energy. This form of IBC takes advantage
-century site. The property of freely allowing the magnetic fields to flow through biological tissues [32].

Galvanic Coupling (GC), injects a weak current on the order of 0.5 mA into human tissue as the main signal that can be modulated to interconnect implants. Leveraging the dielectric properties of human tissue (e.g., conductivity, permittivity, etc.), this waveguide based approach confines the signals within the human body, while ensuring low signal absorption within human tissue [132]. Galvanic coupling is a method that uses the human body as a channel to propagate the electrical signal created by a pair of coupled electrodes. Alternating current is coupled inside the body instead of between the body and the environment. On both the transmission and reception side there are two electrodes; the differential voltage between the two transmission electrodes causes flow of an alternating current through the body to be measured at receiver [131].

Based on information from literature, a qualitative comparison between GC and other IBC
CHAPTER 1. INTRODUCTION

Figure 1.4: Qualitative comparison of design constraints for all IBC methods

technologies is presented in figure 1.4. Given this comparison, it can be seen that Galvanic Coupling promises to provide the best trade-off between data rate, power, attenuation, low interference and system design. For these reasons, among others (introduced in later chapters) Galvanic Coupling is chosen as the enabling method for the development of an application that utilizes IBC.

1.1 Example Applications for Galvanic Coupling

This section presents a series of application scenarios using different characteristic features of GC-IBNs, with different operational requirements. It is important to note that applications presented in this section represent only a small sample size of what exists in the medical field and commercial market. Table 1.1 lists several existing medical applications, summarized from [4], which can benefit greatly from the IBC methods summarized in this chapter. For example, current implementations of deep brain stimulation systems, have limitations in terms of where devices in the brain can be located. This creates the need for a communication channel that propagates information through the inner tissues to a data aggregator (which can be implanted or on-body). Without the use of IBC, the device is only capable of receiving information from parts of the brain reachable via a wired infrastructure, presenting surgical inconvenience for the patient. Advancement in intra-body communication technologies will lead to the creation of a network where several of the applications mentioned in table 1.1 support a system where remote monitoring of biological data and stimulation of tissue/organisms can occur, with little inconvenience to the patient.
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Table 1.1: Telemedicine Applications

<table>
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<th>Telemedicine Application</th>
<th>Target Data Rate</th>
<th>Max. Target BER</th>
<th>Max. Target Latency</th>
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<td>Deep Brain Stimulation</td>
<td>1 Mbps</td>
<td>$10^{-3}$</td>
<td>250 msec</td>
</tr>
<tr>
<td>ECG (12-channel)</td>
<td>72 kbps</td>
<td>$10^{-10}$</td>
<td>250 msec</td>
</tr>
<tr>
<td>EEG (24-channel)</td>
<td>86 kbps</td>
<td>$10^{-10}$</td>
<td>250 msec</td>
</tr>
<tr>
<td>EMG (12-channel)</td>
<td>576 kbps</td>
<td>$10^{-10}$</td>
<td>250 msec</td>
</tr>
<tr>
<td>Video and Medical Imaging</td>
<td>10 Mbps</td>
<td>$10^{-3}$</td>
<td>100 msec</td>
</tr>
<tr>
<td>Drug Delivery</td>
<td>&lt; 1 kbps</td>
<td>$10^{-10}$</td>
<td>250 msec</td>
</tr>
<tr>
<td>Temperature/Glucose Monitoring</td>
<td>&lt; 10 kbps</td>
<td>$10^{-10}$</td>
<td>250 msec</td>
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Figure 1.5: Conceptual diagram of neural stimulation and monitoring system using galvanic coupling and radio frequency

1.1.1 Wireless Neural Stimulation and Monitoring

Electric medicine, or the use of therapeutic electric stimulation in health care applications, is a rapidly growing field. Deep brain stimulation for Parkinson’s disease, cochlear and retinal implants, epidermal stimulation for spinal cord injuries, and vagus nerve stimulation for epileptic seizures are just a few recent applications of electric medicine. Neural stimulation and monitoring can greatly benefit from technological improvements in discovery platforms and clinical applications: biocompatibility/mechanics, invasiveness, resolution, durability and power demands. The use of intra-body communication can directly address some of those improvements in a system utilizing neural stimulation using a combination of high specificity electrodes communicating directly with the nerves in the brain and with a relay node that collects information to be delivered outside the body.

Designing and implementing the neural stimulation and communication system presented
in figure 1.5 can resolve some of the technical issues involved in electrical medicine. This use case involves a minimally invasive approach, since the implanted electrodes will not need frequent changes due to the low power consumption of galvanic coupled communication systems. The relay node can be implantable or placed on the surface of the body, depending on the level of invasiveness desired in each use case. Furthermore, the use of an array of electrodes and beamforming approach designed specifically for use within the brain channel can increase the resolution of data communicated to and from the relay node to the nerves. For the intra-body communication link between the electrodes and the relay node, \[85\], devises a distributed beam-forming approach that allows coordinated transmissions from the implants/electrodes to both the on-skin relay nodes as well as target remote sections of the brain tissues without direct electrode contact. Additionally, the same electrode elements can serve a multitude of purposes; communication, sensing and neural stimulation. In order to examine the behavior of the brain tissue as a channel for galvanic coupling, the typical skin-fat-muscle-bone model needs to be designed, and tailored to mimic the various layers within the brain. The other forms of galvanic coupling channel modeling methods mentioned in this study (introduced in subsequent chapters) can also be used to characterize the human brain medium.

1.1.2 Implantable Continuous Blood Glucose Management System

Galvanic coupling intra-body communication also has the potential to revolutionize treatment mechanisms for traditional diseases such as diabetes. Diabetes is a common health issue that has affected more than 29.1 million Americans \[73\]. Diabetes is caused by lack of insulin secretion in body (Type 1) or the body not being able to properly use the secreted insulin (Type 2).

The glucose concentration in a diabetic patient can be maintained close to the non-diabetic
CHAPTER 1. INTRODUCTION

range by injecting the right amount of insulin into the body whenever required. The typical diabetic treatment option is stressful and tiresome \[74, 75\]; the patient tests the blood glucose level periodically with a glucose meter using the blood samples obtained from finger pricks. Current glucose levels are used to estimate the right dosage of insulin to be injected. This method of insulin calculation is cumbersome, inhibiting positive user experiences. Even if correct calculations are made, the chances of hyperglycemia (excessive glucose) and hypoglycemia (diminished glucose) persists in most people, especially with Type 1 diabetes.

Technology has sought to lessen the burden of diabetic patients by introducing the continuous glucose monitor (CGM) \[76\] and insulin pump. The CGM has a sensor that captures the real-time blood glucose levels. The sensor then transmits the reading to the patient’s mobile device, thus freeing the patient from frequent finger pricking. However, the patient still has to calculate the correct dosage of insulin for a given blood glucose value and operate the insulin pump periodically.

Innovation of automatic insulin pumps may prove to be a major benefit to the diabetic community, performing the insulin quantity calculations autonomously and avoiding the possible human error. In the bionic pancreas proposed in \[77\], the insulin pump runs an application in the patient’s mobile device to make decisions on the amount of insulin to be injected, based on the CGM values transferred to the device by the sensor. Though beneficial, this technique forces the patient to always carry a mobile device as both the CGM and insulin pump communicate via the device. In this scenario, a mobile switch-off situation caused from battery depletion would be catastrophic.

Galvanic coupling can provide an improvement to the current system limitations for existing glucose control systems with a CGM and automatic insulin pump. Currently, the CGM is an inconvenient add-on to the patient’s body: the sensor has a needle inserted subcutaneously with an on-skin transmitter, a set-up that is prone to infection and requires to be replaced every 2-3 days. Apart from showing a large patch on body, the patient can experience itching and bruising at the site of insertion. This inconvenience to the patient is expected to last throughout the entire usage of this system. To reduce this hardship by half, \[78\] and \[79\] proposed implanting the sensor along with the transmitter subcutaneously as a long term relief, eluding any on-surface sensing component. An implanted sensor enables accurate more recent glucose values and the sensors can be implanted with minimally invasive techniques and can be replaced over longer periods of time.

In this topology, represented in figure 1.6, the sensor has to communicate from inside the body to a receiver embedded in the pump attached on the skin. For such intra-body communication, the radio frequency based techniques that are most suitable for the over the air communication, prove to be unsafe owing to high amounts of RF signal absorption by tissues. As a safe alternative, galvanic
coupling based intra-tissue communication can be adopted, using less energy for communication
than its peer techniques.

As an added advantage, this system can form a multi-parameter based closed loop diabetes
management system by using the observations of several other physiological parameters influencing
blood glucose levels. Other factors such as the activity level, sleep level, food intake, perspiration rate,
heart rate, estimations of food intake, varying pH levels, body temperature and ambient temperature
are a few of the human body variables that can cause a fluctuation of glucose levels. The above
mentioned observations can be made possible either with implanted or on-surface sensors and can
be networked towards an adaptive control system with the proposed GC-IBC technology. Such a
system can intelligently and autonomously adapt the data reporting rate of the sensors and direct the
action of the insulin delivery actuators. With the help of machine learning techniques and behavior
predicting algorithms, the possibility of hyperglycemia or hypoglycemia can be completely avoided
enabling the diabetic patients to lead a normal life.

1.2 Selected Application

The research issues addressed in this dissertation are derived from the study of a particular
use case for a galvanic coupling intra-body communication. This section presents the chosen
application that motivates the work presented in the following chapters.

1.2.1 Secure On-Skin Transmission of Body-Generated Password

Securing personal information is a growing concern due to an increase in password pro-
tected mobile devices/services at our disposal. Malicious attempts to crack these secure passwords
allow access to sensitive information, thereby compromising the entire system. Biometric authentica-
tion techniques exist to provide enhanced forms of security (e.g.,to the fingerprint ID on smartphones),
but are also being challenged by unique ways of exploitation. Recalling that the galvanic coupling
IBC technology is energy efficient and robust against external interference, this work envisions
applying the GC technology to specific contributions in the area of password replacement.

The propagation characteristics of galvanic coupling (signal transmission confined within
the body) and unique a biological marker generated by the body (e.g., electrocardiogram signals),
can be combined to create an alternative method of securing our personal devices. The generation
of GC-signals directly within human tissue would yield communication links that are immune to
CHAPTER 1. INTRODUCTION

Figure 1.7: Conceptual diagram of body-generated password using galvanic coupling

vulnerabilities such as eavesdropping and spoofing. The desired use case, illustrated in figure 1.7 requires an individual to wear an on-body device, preferably a wrist-worn wearable. Upon direct contact the unique biological identifier is recorded, processed and communicated from within the body to a receiver located on the device of interest.

• Dissertation Goals and Organization: As shown in figure 1.2 there is a growing interest in this area of implantable sensors, which makes the study of various IBC physical layer solutions timely. This dissertation provides the design and development of a GC-based biometric authentication system (mentioned above), contributing the following analysis leading up to a proof-of-concept hardware and software implementation: (i) comprehensive tables created by studying metrics from current literature that provide at-a-glance overview of the design constraints of IBC systems, (ii) analytical and empirical characterization of GC signal propagation in human tissues, (iii) simulation and run-time testing of various GC communication system designs (heavily influenced from the channel modeling), (iv) an adaptive modulation and transmission scheme based on live channel measurements and (v) a biometric authentication system utilizing the non-radiating GC concept for improved inherent security. The remainder of this dissertation is organized as follows: Chapter 2 provides an extensive literature review for the various components related to GC-IBC system design, Chapter 4 presents the methods employed for understanding the behavior of how the human body behaves as a communication medium under the influence of GC, Chapter 5 illustrates how the channel models are used to design PHY layer communication systems for GC, Chapter 6 details the design and development of the biometric authentication system, Chapter 8 provides the algorithm and performance of an adaptable modulation and transmission scheme to account for changes in the human body channel and Chapters 7 and 9 wrap things up with the conclusions and future work, respectively.
Chapter 2

Related Work

This chapter provides a literature review of various intra-body communication techniques and their metrics on tissue safety, power consumption and channel modeling, with greater detail given to Galvanic Coupling related information. Additionally, communication systems and applications similar to the one chosen for this dissertation will be presented. The information presented in this chapter will be useful in determining the key differences and trade-offs between existing work and the contributions of this dissertation.

2.1 Intra-body Communication Methods

2.1.1 Radio Frequency (RF)

The most salient features of RF-based IBC, narrow band (NB) and ultra-wideband (UWB) channels, are presented first. Here, the RF transceivers, in the form of implanted sensors, communicate by emitting electromagnetic waves in different frequencies.

![Figure 2.1: Radio frequencies used by implantable medical devices](image)

Figure 2.1: Radio frequencies used by implantable medical devices
CHAPTER 2. RELATED WORK

- **RF-Narrowband:** RF-NB utilizes bands known as the Wireless Medical Telemetry Service (WMTS) and the Medical Implant Communications Service (MICS) that were released by the Federal Communications Commission [8]. The bands allocated for WMTS are in the 608-614 MHz (i.e., channel 37 in the digital TV band), 1395-1400 MHz (lower-L) and 1429-1432 MHz (upper-L) range [9]. MICS uses the 401-406 MHz range [10]. The main difference between the two bands is that WMTS is used for remote patient monitoring whereas the main purpose of the MICS band is on-body sensor communication. All the RF frequency bands used in IBN applications are depicted in figure 2.1. MICS has since been renamed as MedRadio and more frequency bands have been added, such as the 2360-2400 MHz band specifically designated for Medical Body Area Networks (MBANs) [10]. More specifically, it is used for low power body sensors networks controlled by a hub device either on or in close proximity to the body. Overall, the MICS band is used by implanted medical devices for diagnostic and therapeutic purposes. Examples of medical devices utilizing the MICS are implanted cardiac pacemakers, defibrillators as well as neuromuscular stimulators. The bandwidth of the channels ranges from 100 kHz to 6 MHz and can be used only by authorized health-care providers.

MICS, WMTS and mostly wireless BAN related applications use the IEEE 802.15.6 standard, defining the PHY and MAC layers for specific medical or non-medical applications [11]. The MICS and WMTS bands are preferred over the unlicensed ISM bands due to the reduced noise and the security provided by the regulatory protection of the former. ISM bands were not originally allocated for communication between medical devices, therefore interference is a persistent issue. The FCC recognized this and allocated bands specifically for communication between medical devices, whether implanted or not [8].

RF-NB technology has been used in a variety of implantable medical devices, such as the Zarlink transceiver, which uses the MICS frequency range to communicate at data rates of 200-800 kbps in the body [8]. Additional experimental work in [12] showed data rates of up to 1 Mbps covering a distance of 20 cm in the body. Similar experimental work in the MICS frequency range showed an attenuation of 60-80 dB for communication links of 15-20 cm in the body [13].

- **RF-Ultra wide Band:** The RF-UWB spectrum utilizes the 3.1-10.6 GHz range. Experimental results have shown that RF-UWB used for communication in the body can reach data rates up to 1 Mbps covering a distance of 12 cm, when operating at a frequency of 4 GHz [14]. A consistent characteristic of RF IBC is the high attenuation levels, both experimentally observed and shown in theory. More specifically, through animal trials, [14] demonstrated that the signal attenuation is above 80 dB through tissues. Even with the high attenuation levels, RF-UWB and RF-NB technology have...
already been used within implantable pacemakers that wirelessly communicate important diagnostic information, such as cardiac rhythm, as well as information about the device status (power level) to the physician. The same technology has aspired researchers to develop implantable neuromodulation systems that communicate diagnostic information, such as the pressure of transporting fluids (e.g., blood). There are two major concerns that arise when considering the use of RF, both stemming from power consumption. The high attenuation requires transmission at greater power levels within the tissues compared to an equivalent coverage distance in free space. This not only poses concerns on tissue heating, but also for energy replenishment of the sensors, once they are implanted.

- **Millimeter wave:** With the recent growth in the number of connected devices and demand for accessible mobile data, fifth generation (5G) mobile networks aim to alleviate the issue of spectrum scarcity by utilizing the bountiful amount of spectrum located in the millimeter wave (mmWave) domain, operating on the band of spectrum between 30 and 300 GHz. Promising to provide an enormous increase in communication capacity, emerging applications in mmWave communication strive provide device-to-device (D2D) inter-connectivity among wearable electronics that require throughput on the order of gigabits per second (e.g., virtual reality, augmented displays). However, some of the characteristics of mmWave communication (large bandwidth, reasonable isolation and dense deployment) can yield major research challenges in terms of high propagation loss, directivity and sensitivity to blockage/interference. The presence of these characteristics can also yield an increase in tissue heating (a phenomenon also prevalent within other RF-based approaches) if directed towards the human body. Studies from indicate that although tissue heating can extend to deeper layers of the body, more than 90% of the transmitted electromagnetic power is absorbed within the epidermis and dermis layers of the skin. These characteristics suggest that mmWave communication is difficult to practically implement within implantable devices. However, there is significant opportunity for on-body scenarios. For example, proposes an on-body mmWave channel that spans a link distance of up to 50 cm with associated maximum attenuation values of approximately 80 dB. RF methods have been used traditionally, as seen throughout this section, for intra-body and on-body medical applications.

However, in recent years, however, development of RF implants have become rather stagnant with an emphasis on alternative methods to communicate through the human body.
CHAPTER 2. RELATED WORK

2.1.2 Ultrasound

Mechanical waves that propagate in frequencies above 20 KHz (the upper limit of human hearing) are also known as ultrasound waves [18]. Ultrasound has been used extensively for underwater communications due to its optimal propagation through media composed of mostly water. For this reason, [5] advocates the use of ultrasound for IBC since the human body consists of 65% water. Ultrasound has been used in medical applications where the transmission of image and telemetry data occurs from a device inside the body to receivers on the outside (e.g., capsule endoscopy). Ultrasound is preferable to RF in this example case of capsule endoscopy due to its higher power efficiency.

Experimental data in [20] reveals that an ultrasound signal with Single Side-Band modulation (SSB) or Double Side-Band modulation (DSB) presents higher frequency and power efficiency than RF signals, through there is an associated trade-off in the low propagation speeds of mechanical waves. The low radiated power allows for a better safety margin, while providing the required image bandwidth for the application. Another phenomenon unique to ultrasound propagation in human tissue, known as cavitation, is the expansion and contraction of gas bubbles due to the varying pressure of an acoustic field. This health effect must also be considered along with traditional safety limits (to be discussed later in Section III) placed on tissue heating.

The most recent and complete prototype for ultrasonic intra-body communication is presented in [67] and achieves a data rate of 90 kbps with a BER of $10^{-6}$. The power consumption of such a system is at 36 mW. The highest data rates achieved in ultrasound appear promising, with a maximum of 28.12 Mbps data rate demonstrated through synthetic phantoms [21]. However, experimental data rate at a power of 40 $\mu W$ was measured at 700 kbps, which is significantly lower [18]. Studies on ultrasounds for communication in the body has shown that attenuation is lower than that of RF communications. At a distance of 20 cm, when operating at 5 MHz, the attenuation is experimentally shown as 25 dB [22], much lower than that for RF. Overall, ultrasound is a promising communication method for IBNs, especially for applications that require high data rate transmission.

2.1.3 Capacitive Coupling

Coupling methods in general are based on the energy transfer between a set of transmitters and receivers to generate an electrical signal that propagated through the human body [23]. The electrical signal generated by coupling methods is low-frequency (under 120 MHz) and low-power (in the order of $\mu W$ compared to traditional electromagnetic signals that go up to several GHz. For
## CHAPTER 2. RELATED WORK

<table>
<thead>
<tr>
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</thead>
<tbody>
<tr>
<td>RF-NB</td>
<td>80</td>
<td>20</td>
<td>401-406 MHz</td>
<td>800 kbps</td>
</tr>
<tr>
<td>RF-UWB</td>
<td>&gt; 80</td>
<td>12</td>
<td>3.1-10.6 GHz</td>
<td>500 Mbps</td>
</tr>
<tr>
<td>mmWave</td>
<td>80</td>
<td>50</td>
<td>30-300 GHz</td>
<td>Range</td>
</tr>
<tr>
<td>US</td>
<td>100</td>
<td>10</td>
<td>1-100 MHz</td>
<td>28.12 Mbps</td>
</tr>
<tr>
<td>CC</td>
<td>65</td>
<td>170</td>
<td>100 kHz-120 MHz</td>
<td>60 Mbps</td>
</tr>
<tr>
<td>GC</td>
<td>65</td>
<td>15</td>
<td>100 kHz-10 MHz</td>
<td>1.56 Mbps</td>
</tr>
<tr>
<td>RC</td>
<td>35</td>
<td>130</td>
<td>DC to 50 MHz</td>
<td>-</td>
</tr>
</tbody>
</table>

**Table 2.1: Overview of IBC Methods and their Characteristics**

this reason, coupling methods have become a popular component of the on-going research on IBC since their low power and low frequency complies with safety considerations and decreases energy consumption [23, 131, 25, 26].

Capacitive coupling occurs when two circuits sharing the same electric field cause a flow of energy from one circuit to the other. In the case of intra-body capacitive coupling, the common electric field of the body and its environment causes an induced current flow from a transmitter to a receiver in the form of electrodes. One transmitting and one receiving electrode are attached to the body, while the other two are floating, acting as ground electrodes. The body acts as a conductor of the electric potential and the ground acts as a return path for the signal [131].

Capacitive coupling has been studied extensively to determine the data rate it can sustain, common levels of channel attenuation and expected levels of power consumption. Recent works on capacitive coupling at 60 MHz exhibits an attenuation of 20-25 dB while covering a distance up
CHAPTER 2. RELATED WORK

to 140 cm in the body \cite{41}. At this operating frequency, the attenuation is lower than that of both RF IBC methods and ultrasound. Additionally, the data rates that have been calculated theoretically in \cite{131}, where promising results indicate a maximum rate of 2 Mbps for the 1-200 MHz range. Recent experimental works show a maximum data rate of 60 Mbps by employing a multi-level coded transmission scheme \cite{40}, while operating within a bandwidth extending from 40-80 MHz. Overall capacitive coupling is a promising method for IBC, although its ability to significantly interact with the surrounding environment can be considered both a positive and negative attribute.

2.1.4 Resonant Coupling

Resonant Coupling uses the properties of electromagnetic resonance to generate a magnetic field throughout the body. RC is employed to create a near-field wireless transmission of electrical energy between two coils wrapped around parts of the body, driving the field propagation. Its potential benefits arise from low power requirement.

The spectrum range most commonly used in RC research extends from DC up to 50 MHz yielding a maximum attenuation of only 8.1 dB for a 40 cm distance covered \cite{45}. The results of RC are very promising but this area is still in a nascent stage in terms of proven experimental data. For example, there are no conclusive results for the data rate achieved by communicating through the body using RC. The wavelength of the magnetic field proposed in the work by Park in \cite{45} is 2.3 m, which can potentially interfere with the magnetic fields of other nearby devices. Also, the interference of the system from other magnetic fields, including electrical machinery, needs to be investigated further.

2.2 Safety Considerations

The ways in which electromagnetic waves interact and affect human tissue have been widely studied. As a result, several standards have been created that set explicit exposure limits in terms of field strength, power density and average time of exposure. The IEEE C95.1 standard \cite{46}, uses the aforementioned variables to determine restrictions over a range of operating frequencies. Specifically, this standard discusses the limits of signal exposure within the body from the perspective of electrostimulation and tissue heating. In the frequency range of 3 kHz to 5 MHz, limitations are primarily set to alleviate hazardous effects related to electrostimulation; in the operating frequency range of 100 kHz to 300 GHz, the dominant effects that need to be accounted for are associated
CHAPTER 2. RELATED WORK

Figure 2.2: Comparison between MPE values for different IBC methods

with tissue heating. A special frequency range (0.1 to 5 MHz), known as the transition region, have set limits that account for both phenomena. Also within this region, tighter bounds are placed on continuous wave signal exposure that have adverse effects on heating, while short isolated pulses (with low duty cycles) have more restrictions on electrostimulation [46]. The Specific Absorption Rate (SAR—Watt/kg) is a measure of the amount of energy absorbed within a given amount of mass. It is calculated from various field strength values and is used to derive another metric known as Maximum Permissible Exposure (MPE). This quantity describes the maximum rms, peak electric or magnetic field strengths, or power density to which a person may be exposed without incurring adverse health effects. These can be derived or estimated from induced electric field, SAR, or power density values. In this work, we will use MPE to compare tissue safety for each IBC method.

Figure 2.2 and Table 2.2 illustrate the differences between the MPE (mWatt/cm²) values associated with the IBC methods within that employ electric, magnetic and electromagnetic-based fields. The values listed within this figure represent the MPE in an uncontrolled environment, at the maximum operating frequency range mentioned for each IBC solution within this survey.

Ultrasound waves are expressed in terms of Intensity, a measure of the energy absorbed through the transfer of mechanical waves into a medium. The limits of ultrasound intensity are set by the Food and Drug Administration and the American Institute of Ultrasound in Medicine [18] and utilize the same units for quantifying MPE, allowing for a better comparison against the other non-acoustic based approaches in this study. However, the ultrasound intensity limits are not

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CHAPTER 2. RELATED WORK

Table 2.2: Maximum Permissible Exposure of IBC Methods

<table>
<thead>
<tr>
<th>IBC Method</th>
<th>Operating Frequency</th>
<th>MPE (mW/cm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF-UWB</td>
<td>10.6 GHz</td>
<td>1</td>
</tr>
<tr>
<td>RF-NB</td>
<td>406 MHz</td>
<td>0.203</td>
</tr>
<tr>
<td>US</td>
<td>5 MHz</td>
<td>730</td>
</tr>
<tr>
<td>CC</td>
<td>200 MHz</td>
<td>0.2</td>
</tr>
<tr>
<td>GC</td>
<td>10 MHz</td>
<td>1.8</td>
</tr>
<tr>
<td>RC</td>
<td>50 MHz</td>
<td>2.02</td>
</tr>
<tr>
<td>mmWave</td>
<td>30 GHz</td>
<td>1</td>
</tr>
</tbody>
</table>

displayed within figure 2.2 but are listed within table 2.2, since the average values show at least 360x increase from the highest MPE values represented in the figure. Therefore, this type of IBC technology creates minimal heat dissipation in the tissue compared to the other approaches based on electromagnetic waves. Table 2.2 also summarizes the MPE for the other IBC methods studied in this survey. For intra-body communication, and the eventual design of intra-body networks, it is important that this crucial design constraint be considered when several sources are transmitting within or on the human body.

2.3 Power Consumption

Table 2.3: Comparison of Power Consumption across IBC Methods

<table>
<thead>
<tr>
<th></th>
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<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>RF-NB</td>
<td>481 µW</td>
<td>2.60</td>
<td>0.45</td>
<td>OOK/FSK</td>
<td>402-405 MHz</td>
<td>ASIC</td>
</tr>
<tr>
<td>RF-UWB</td>
<td>835 µW</td>
<td>1.67</td>
<td>1.20</td>
<td>FSK</td>
<td>3.328-4.608 GHz</td>
<td>ASIC</td>
</tr>
<tr>
<td>US</td>
<td>36 mW</td>
<td>-</td>
<td>-</td>
<td>PPM</td>
<td>-</td>
<td>FPGA</td>
</tr>
<tr>
<td>GC</td>
<td>2 mW (Tx)</td>
<td>1.28</td>
<td>3.30</td>
<td>PPM</td>
<td>65 MHz (BW)</td>
<td>FPGA</td>
</tr>
<tr>
<td>CC</td>
<td>250 µW (Rx)</td>
<td>0.24</td>
<td>1.00</td>
<td>FSK</td>
<td>40-120 MHz</td>
<td>ASIC</td>
</tr>
</tbody>
</table>
CHAPTER 2. RELATED WORK

The powering of biomedical implants enabled with wireless connectivity creates a challenge for IBC in general. When one or more devices must be implanted, it is of utmost importance to minimize the instances of battery replacement/recharge. As this technology continues to progress, ultra-low power transceiver design plays an intricate role in the ubiquitiveness of intra-body networks. The design of these wireless communication systems varies greatly, depending upon which IBC technology is employed and the application (refer to table 1) that is meant to be utilized. Additionally, obtaining a proper comparison of power consumption from system-to-system is not trivial. Several factors about the implementation must be compared (see table 2.3) that provide a detailed understanding of which IBC solution has the potential to offer the most reasonable power budget for implantable operation. This fact also gives way to variance among the power consumption values listed in this survey. Figure 2.3 supports this by displaying the Energy/bit Vs. Power Consumption values for some of the work sampled with respect to each IBC method.

In this section, the most recent power consumption studies conducted in relation to IBNs/WBANs is provided. Table 2.3 lists a comparison of the power consumption, along with other related metrics, summarized from various implementations in the published literature. As of late, implementations have gone beyond classical RF while galvanic and capacitive coupling, as well as ultrasound have begun to emerge. Solutions involving resonant coupling are still in the early stages with limited information on power consumption. Power consumption analysis for mmWave based IBC communication is also notably absent. It can be hypothesized, that due to the the use of large antenna arrays associated with a beamforming approach, sampling of signals in the GHz range and operational data rates in the Gbps range, power consumption for mmWave communication may

![Figure 2.3: Energy/bit vs power consumption](image)

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2.3.1 RF-Ultra wideband

There have been many published works related to low power design of UWB systems for body area/intra-body network applications. The work presented in [48], describes a low-power frequency modulated ultra-wideband transmitter developed in 130nm CMOS technology. The transmitter is designed to operate within the frequency range of approximately 3.3 to 4.6 GHz, and consumes 835W or power from a 1.2V supply while achieving an energy efficiency of 1.67 nJ/bit. In [49], a UWB transceiver designed for high data rate for human body communication is presented. Developed in 65 nm CMOS technology, and designed for operation between 7.25 and 9.5 GHz, it utilizes OOK modulation with a spectrum efficient frequency hopping technique. Results from this transceiver show that it consumes a total of 13.3 mW from a 1 volt supply, with an energy efficiency of 26.6 pJ/bit.

2.3.2 RF-Narrowband

Narrowband transceivers that operate in the MICS band have also been widely studied and developed in ASIC platforms. In [50], combined modulation schemes of OOK (for wake-up and data communication) and FSK (for always-on communication) are implemented in .18 $\mu$m CMOS technology. The wake-up mechanism with OOK modulation is done to reduce power consumption levels, while FSK operation grants greater noise immunity and BER. Power consumption totals in the amount of 481 $\mu$W, with an energy efficiency of 2.6nJ/bit (OOK) and 1nj/bit (FSK). The work done in [51], designed specifically for implantable devices, is also constructed in 0.18 $\mu$m CMOS technology and is supplied with .7 Volts for operation. It is specifically configured to mitigate interference, while employing FSK based modulation. This system is capable of an energy efficiency of 1.96 nJ/bit, while consuming 890 $\mu$W of power.

2.3.3 Capacitive Coupling

Capacitive coupling transceiver design for low power communication has been represented on FPGA platforms [54], but have recently shifted towards the ASIC domain, indicating the growth in attractiveness for alternative IBC solutions. One such example is a crystal-less double FSK transceiver and a collection of other FSK based systems that are designed and studied within [4], respectively. This system was implemented in .18 $\mu$m CMOS technology and operated with a 1 volt
supply. The transceiver consumes a total of 5.4 mW with an energy efficiency of 3.2 nJ/bit. A more recent study on front end design for a capacitive coupling transceiver, detailed in [55] is developed in 65 nano meter CMOS technology and is designed to be suitable for data rates on the order of kbps, operated with a supply of 1.2 volts, with an active power consumption of approximately 250 µW.

2.3.4 Galvanic Coupling

For GC based systems, the most recent works have been implemented on platforms with an FPGA at its core [141]. The work proposed in [53] consists of a system using a pulse position modulation (PPM) scheme, similar to an impulse radio ultra-wideband techniques. An FPGA is also employed to facilitate PPM baseband transmission. Results demonstrate that the transmitter system power consumption is 2.0 mW with a 3.3 volt supply, and an energy efficiency of approximately 1.28 nJ/bit. Results demonstrate a suitable data rate for GC-IBC in the range of 1.56 Mb/s. The PPM transmitter power consumption is 2.0 mW with 3.3 V supply voltage. Having energy efficiency as low as 1.28 nJ/bit provides the potential for enhanced solutions where portable biomedical applications can be applied.

2.4 Channel Modeling Methods

Signals emitted by each intra-body communication technology undergo different channel-induced transformations while propagating on or within the human body. In this section, context is provided about the common channel modeling techniques employed for uncovering signal propagation behavior in human tissues. One representation of this, figure 2.4, depicts the occurrence of several types of channel modeling methods with respect to each IBC technology. The modeling techniques fall under the classification of Analytical, Empirical, Numerical, Statistical, and Hybrid (Analytical + Numerical) methods, for the human body channel under the influence of each IBC method.

2.4.1 RF-Ultra wideband

RF-ultra wideband channel for the human body medium in [65] is modeled analytically by finding closed form expressions via Maxwell’s Equations inside a homogeneous medium. This method captures similar results to the work done in the numerical domain, such as [66], where the dielectric nature of human tissue is considered. In this work, Finite Difference Time Domain analysis
CHAPTER 2. RELATED WORK

is done to gain a better understanding of the intra-body communication from a node embedded in the chest to an external base-station like device. The characterization work presented in \[14\] evaluates and validates the performance of liquid based phantom (similar to muscle tissue) trials by conducting experimental modeling on a live animal (pig) where nodes are surgically implanted. The overall analysis of channel behavior is consistent among the other works presented in this subsection. Other results indicate that multipath fading is present, but due to high path loss beyond a certain range, its effect presents little to no impact on the performance of signal propagation. To exemplify this, the work done in \[14\] indicates that the relatively short multipath delay spread can be approximated using a two-ray propagation model. Additionally, blood and muscle tissues contribute to higher path loss when compared to bone and fat layers within the body.

2.4.2 RF-narrowband

Channel modeling RF-narrowband is conducted via hybrid, empirical and numerical based methods. In \[12\] a path loss model for multi-layered homogeneous tissue layers within a heterogeneous body is presented. Finite Difference Time Domain approach is used in conjunction with equations based on Friis transmission using while accounting for the dielectric nature in tissues. Empirically, work has been done in \[64\] to study the path loss and channel characteristics for nodes implanted within a single tissue layer. A ray-tracing based approach to channel modeling was performed on a homogeneous tissue phantom, also with representative dielectric properties. The Finite Difference Time Domain numerical approach is presented in \[13\], where the RF antenna is also modeled as a part of the channel, for a simplified homogeneous 3-layered tissue structure. Channel behavior uncovered from the hybrid modeling method, details that log normal shadowing is present in the channel and that an increase in tissue conductivity yields degradation of the path loss (similar to RF-UWB).
CHAPTER 2. RELATED WORK

2.4.3 mmWave

In crowded environments such as train cars or airline cabins, human bodies become significant sources of blockage in the mmWave frequencies. Self-blocking can occur where the user’s body location results in restricted access to the local network in addition to causing unwanted interference for other wearable networks. A possible network architecture assumes the user’s smartphone to function as a gateway hub, and the neighboring wearable networks associated with different users are likely uncoordinated. Thus, techniques involving stochastic geometry are used to model and analyze the performance of these networks with a finite number of interferers in a bounded network region. These models incorporate different path-loss and small-scale fading parameters, depending on whether state of the link.

Current models for on or near the human body operation describe the propagation characteristics as quasi-optical. presents an empirical and statistical approach at modeling the mmWave on-body channel. From the empirical model, the separation of short-term and long-term fading, occurs by experimentally employing a temporal averaging window of the order of 0.11 seconds. The data gathered from the body-worn setup is also used to facilitate statistical analysis, uncovering that the short-term fading best fits the Cauchy-Lorenz distribution. In an analytical approach, validated through numerical simulation is used to uncover insight about the overall system performance (e.g., spectral efficiency, SINR coverage probability, blockage probability) in the presence of multiple interfering agents. In this study, the Nakagami distribution is assumed and the human body itself is modeled as a source of link blockage. Despite the recent work related to on-body mmWave channel modeling, significant work is needed to quantify signal transmission characteristics and incorporate the results into practical systems design.

2.4.4 Ultrasound

The work conducted in, uses an ultrasonic testbed to empirically measure the channel. This FPGA-based prototype, with a kidney phantom as the medium for propagation, is supplemented with a statistical model to characterize the behavior of interference. The model uses a new approach known as the M-sampling method, where the interference at the receiver is characterized by taking multiple sample sets at various instances of time within pre-specified interval. In this work, the channel behavior takes on the form of the generalized Nakagami probability distribution. In order to design proper analytical models that represent ultrasound propagation in various media, applies acoustic wave equations (Helmholtz equations) in order to characterize physical parameters such as
pressure of sound waves, amplitude, propagation speed, wave intensity, etc. Numerical methods are further employed to satisfy the need to represent acoustic field behavior in time and space throughout the human body. The work in [63] utilizes what is known at the pseudo-spectral (PS) approach that exploits Fourier series expansions, FFTs and the k-space method (an approximation method to obtain temporal derivatives) to perform operations in the spatial and time domain, respectively. Results from the previously mentioned works, in addition to the research presented in [22], show that the general ultrasonic human body channel undergoes small scale fading and slow propagation speeds, resulting in large multi-path delay spreads. This phenomenon is due to the inhomogeneity of tissue in terms of density and sound velocity, where numerous reflectors and scattering surfaces exist. The channel characteristics of ultrasound inhibit the use traditional passband modulation or continuous wave transmission due to their poor performance in multipath channels. Thus, systems that use pulse based low duty cycled communication or OFDM are examples of desirable solutions. However, OFDM requires high system complexity and yields a peak-to-average power ratio that requires inefficient linear power amplifiers, a less ideal solution where strict power budgets exist.

2.4.5 Capacitive Coupling

Capacitive coupling channel characterization uses circuit models to drive analytical expressions [59] and different subjects to test human body variation during empirical modeling [54]. [59] suggests that numerical models do not accurately account for the electrode-tissue contact impedance, an important factor used in estimation of signal propagation loss. Thus, the authors utilize a hybrid channel modeling approach, where there exists a combined benefits of an analytical electrostatic circuit model and numerical simulation of 3D human body geometries under electromagnetic fields. Each approach accounts for the frequency dependent dielectric properties of human tissue, and reasonable tissue dimensions. This Finite Element Modeling based approach is used to study path loss under different body positions and environments while validating the data from other literature. Capacitive coupling signal propagation characteristics can vary under different environmental conditions, human body posture, and if external interferers are present. Channel gain versus frequency plots provide evidence that frequency selectivity is present within the CC channel, indicating the possibility of multipath fading. However, no additional information is provided (e.g., phase variation, channel impulse response, etc.), making it difficult to set values for the metrics in table 2.4. Additionally, capacitive coupling based communication places tight requirements on the design of ground and signal electrodes and shows high variability in path loss measurement with different
system configurations. [59].

2.4.6 Galvanic Coupling

Galvanic coupling allows modeling the human body channel as a complex network of impedances representing the different tissue layers. For example, in [118], each layer of human tissue is analytically modeled as a combination of electrical circuit that constitute a 2-port network for the computation of signal gain at an output terminal. The implementation of this 2-port circuit model allows for easy modification of parameters (center frequency, electrode separation, electrode dimensions, etc.) and flexibility in terms of transmitter and receiver location. The model also takes into account several paths that the current can travel within the layered human body model, enabling the observation of the scattering parameter and boundary reflection at the electrode-tissue interface. Similarly, in [122], an equivalent circuit representation is used to model the behavior of the human body under the influence of galvanic coupling, and experiments are done on a human upper arm, torso and leg to validate the model. In [123], mathematical modeling and simulation of galvanic coupling with different transmission paths is presented, where the human body (abstracted as concentric cylinders) was transformed into a model represented by the head, arm, torso, and leg, for the layers of skin, fat, muscle, cortical bone, and bone marrow. These methods of channel modeling, although highly complex and well inclusive of the dielectric properties of human tissue, are not well representative of the signal propagation path emphasized in this study. Additionally, uniformity is considered across all signal paths and tissue layers, where in reality this assumption does not hold. Validation of similar analytical models were performed using ANSYS HFSS, enabling full-wave electromagnetic simulations, numerically analyzing the electric field distribution at user defined locations within the tissues using finite element analysis (FEA). An empirical channel model from [56] describes a correlative channel sounding experiment to uncover the channel impulse and frequency response of transmission and reception within the skin and muscle layers of a porcine tissue specimen. Other channel characterization work within [123] and [58] used similar methods at numerical, analytical and empirical modeling to capture the channel behavior, where FEM, tissue equivalent circuits, and porcine tissue or human forearms, were used respectively. Galvanic coupling also adds an inherent benefit of additional security and interference resilience. Signals that are confined within the body cannot be intercepted unless one is in direct contact with the medium. The results from the collection of GC channel modeling portrayed in this survey, indicate that biological channel behavior exhibits no multi-path fading, and thus can be considered equivalent to an AWGN
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channel [56]. The channel modeling conducted in this work focuses on a 3D tissue equivalent circuit model with varying dimensions that adequately reflect the portion of the body used as a communication medium.

2.4.7 Resonant Coupling

Recent work in the area of magnetic resonant coupling human body channel modeling exist mostly in the numerical and empirical domain. The channel studies in [45] and [32] use Finite Element Modeling (via Ansys HFSS) and Finite Difference Time Domain numerical analysis, respectively. These numerical based approaches take advantage of mesh models that account for the frequency dependent dielectric properties and geometries of human tissue. Results indicate that path loss of RC is very low within human tissue, attributed to the low values of tissue magnetic permeability. This concept shows that magnetic fields can freely move throughout the human body. However, channel gain characteristics presented in [45] indicate frequency selectivity that is highly dependent on node placement and body posture. Unfortunately, the insufficient amount of channel modeling data with respect to RC limits the available field entries in Table 2.4.

2.5 Comprehensive Comparison of IBC Methods

A summary of the pros and cons associated with the metrics of each IBC technology are compared in this section. The analysis includes final thoughts on signal attenuation, link distance, achievable data rates, power consumption, channel impairments, tissue safety, along with other parameters listed within Table 2.5 and Figures 2.5 and 1.4.

Although RF-UWB and RF-NB waves have been studied extensively, there are several issues that limit their deployment. RF-UWB requires high energy consumption of 2.6 nJ/bit [53]. As a comparison, coupling methods have an energy efficiency as low as 0.24 nJ/bit. RF operates in high frequencies, 401-406 MHz for NB, the Industrial, Scientific and Medical radio ISM band (2.54 GHz) and 3.1-10.6 GHz for UWB. The health risks related to the high frequency of RF waves in the body are a concern due to tissue-heating caused by the high absorption of the wave-energy [5]. Hence, the MPE of RF methods is lower than that of ultrasound and other coupling methods. Probability of security intrusions and susceptibility to interference of information transmitted using RF through the body are high due to the fact that the waves propagate both in-body and off-body, into the environment [5].
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The emerging mmWave technology presents challenges when it comes to application on WBANs. In order to fully realize the design constraints for wearables, accurate channel models are needed. Models that account for the impact of reflections within the finite bounded region and mitigate the complexity involved in boundary value problems relating to Maxwell’s equations, will be a foundational step. At higher frequencies, energy absorption is increasingly confined to the surface layers of the skin. At frequencies beyond 60 GHz, surface waves play no significant role in propagating over human skin and the generated fields are largely attributed to space waves. Penetration depths at the 60 GHz range are on the order of millimeters, and thus, only the dielectric properties related to the skin need to be considered for modeling scenarios where person-to-person body variability needs to be studied.

Resonant coupling is the least studied method out of all IBC technologies with extensive research underway to fully understand the channel characterization metrics. Though RC is safe for the human body in terms of tissue heating, the interference of the magnetic field created around the body with other magnetic fields in the environment is still an issue that may affect the communication link. Capacitive Coupling, though effective at providing long range IBC and sufficient data rates, does not solve some of the issues that present itself with RF-IBC. As mentioned in section 2.1.3, CC depends on the ground reference to create a capacitive link that extends outside the human body, thereby increasing it’s interference domain. This level of susceptibility makes capacitive
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coupling a less preferable method for IBC, especially as more users are equipped with their own IBN/BANs.

Although high data rates systems have been achieved using ultrasound in the body, practical rates are in the high kbps range [67]. Ultrasound waves with low duty cycles can cover sufficiently long distances in the body without exposing tissues to high temperature. Ultrasound-UWB techniques mitigate the channel impairments and medium access control algorithms built on top of this physical layer solve the interference problem for ultrasound in the presence of multiple users [5].

Galvanic coupling offers moderate transmission distances, lower data rate compared to methods such as CC, US and RF-NB and can safely operate with relatively high limits within the human body. It’s true advantages are its low attenuation and full confinement of signals inside the human body, offering more security and interference-free communication. The simplistic GC-IBC communication channel allows for low complexity in transceiver design, resulting in lower power consumption. Though the data rates are lower compared to other IBC methods, they are still suitable for many telemedicine applications.

Galvanic coupling can be used for communication between skin, muscle and fat tissue and its properties are affected by the tissue layer that is used as a medium as well as the location of the electrodes on the body. Experimental research in [23] has shown that GC links are stronger in muscle-to-muscle communication and, in terms of location, the thorax seems to have the best transmission characteristics [26]. Overall, Galvanic Coupling is a safe and efficient method but relatively new in the field of IBC, with ongoing research on data rate, attenuation, optimal frequency, among others, in order to achieve the most efficient communication throughout the human body. Galvanic coupling is a promising IBC method because, similar to CC, it offers low-power and low-frequency signals. The transmission data rates appear to be lower than those of CC but at the same time, there is no need for a floating ground reference or the environment as a path for the signal to traverse [131]. GC signal frequency ranges from as low as 10 kHz [27] to 100 MHz [28] for most effective communication. Experimental work in [29] has shown maximum data rates up to 1.23 Mbps when transmitting at just 200 kHz with attenuation levels typically at 50 dB when covering distances up to 15 cm [30]. The additional comparisons provided in the previous section, further motivates the use of Galvanic Coupling as an enabling method for IBC. It’s low attenuation, high safety limits, high energy efficiency and confined signal propagation are the main attributes that we exploit in the resulting GC-based system design for the selected application described in Sec. 4.1.
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2.5.1 Nano Communication Networks

Advancements in the field of nanotechnology have spawned revolutionary scenarios in which an alternative form intra-body communication can be realized. Nanonetworks, or Nano-communication networks, utilize device-to-device communication through interconnected nanomachines within the body. These nano-scale devices are responsible for identifying the presence of harmful agents and delivering drug-induced counter measures without the company of a physician [126]. These devices range in size from hundreds of nanometers to a few micrometers at most and are only bestowed with the capability to perform simple tasks such as computing, storing of data, sensing of biological markers and actuation. [127]. Several communication methods have been proposed for use in nanonetworks, but the most promising and widely studied is molecular communication, where molecules are used to encode, transmit and receive data. Communication via nanomachines function alongside the biological operation of the human body. More specifically, the molecule emission process, molecule diffusion process and receptor binding process, all describe the transmitter, channel and receiver functionality of a single link within a nanonetwork, respectively [128]. As a result, the work conducted in [126,127,128] contribute several channel and system models that are based upon actual phenomena that occur in the biochemical realm within the body. Unfortunately, this research is still in an embryonic stage and although current analysis will be useful in designing a complete system, it has not yet yielded a physical implementation in which performance of a communication link can be quantified.

2.5.2 Cognitive Radio for Intra-Body Networks

Cognitive radio (CR) is a highly utilized technology within the wireless network community. It’s presence is geared towards addressing the spectrum scarcity problem by supporting techniques for dynamic spectrum access. It’s functionality relies on the intelligent enabling of unlicensed secondary users to opportunistically access underutilized licensed frequency bands that are designated for primary users. In the context of Intra-body Networks (IBNs), cognitive radio plays an important role in alleviating the spectrum scarcity problem due to increased deployment of IBNs for medical monitoring and telemedicine related applications. Additionally, it is aimed at reducing interference with neighboring medical devices not necessarily equipped with wireless communication capability. The work conducted in [129] surveys various MAC protocol implementations for CR-IBNs, while [130] surveys a broad classification of CR-IBN contributions, including but not limited to, MAC protocols, spectrum sensing and system architectures. In both the aforementioned works, which
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specifically target the use of medical applications, it is assumed that sensor-to-relay and relay-to-
base station data transfer both employ RF-based methods for intra-body communication. Utilizing
RF-based IBC greatly supports the need for CR-IBNs, as the need for EMI mitigation and spectrum
efficiency drastically increases when all devices typically operate in the Industrial, Scientific, and
Medical radio (ISM) band and RF-bands specified for wireless medical telemetry, medical implants
or wearables. Although these recent works focus on the use of CR-IBNs with RF-IBC, the use of
cognitive radio technology can be expanded to other IBC methods whose communication range
can extend beyond the body. As more personal IBNs are deployed, body-to-body interference
management becomes a problem that cognitive radio aims to solve. Additionally, IBN topologies
will continue to leverage the use of an RF-based external gateway node for the transmission of data
to remote monitoring centers. However, the operation of many CR related functions (e.g., channel
sensing, resource allocation and spectrum mobility) depend on the performance of sophisticated
signal processing and upper layer protocols \cite{129}. Therefore, the research surveyed in this work will
not focus on examining the use of CR-IBNs in great detail, as the focus is on simple point-to-point
physical layer characteristics of each IBC method.

2.5.3 Galvanic Coupling Communication Systems

The creation of hardware and software systems that facilitate galvanic coupling is gaining
popularity for their ability to show the practicality of the technology while also validating attempts
at modeling the behavior of the human body channel. Several experimental platforms and testbeds
using GC have been proposed earlier \cite{124,125,131,132}. The work in \cite{124} validates numerical
simulations for a portion of the human body using tissue simulated liquid. Differential binary
phase-shift-keying (DBPSK) modulation is selected given its robustness to amplitude variations and
minimal hardware complexity when compared to coherent schemes. The design emphasizes low
power and circuit miniaturization for long term implantable medical operation. Similarly, in \cite{125}, a
platform is created for testing the affect of various electrode and body placement configurations on
transmission and reception of galvanic coupled signals. This system is designed for communication
on a real human forearm, utilizing a Xilinx SPARTAN FPGA with supporting front-end hardware.
The work conducted in \cite{131} and \cite{132} reveal the design of an UWB system in preference of the
narrowband schemes typically implemented. The former illustrates a proof-of-concept system, while
the latter improves the design and focuses on multiple access related experimentation. Both works
follow the specifications of the physical layer based on the standards outlined in IEEE 802.15.6
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for Wireless Body Area Networks. A carrier-free, pulse position modulation (PPM) transmitter is employed for operation in an FPGA and received signals are detected via an oscilloscope. This design is able to achieve data rates in the Mbps range due to the very large bandwidth requirements of the system. The aforementioned studies show promising developments for GC communication. However, these systems offer macro-scale representations of the transmitter and/or receiver architecture, characteristics not yet applicable for the wearable device arena. Additionally, data communication of pertinent information has not yet been explored. The majority of the empirical analysis is performed for channel behavior verification. In contrast, this work focuses on the sending and receiving of biological data via GC, with a system design focused on operation in the wearable domain.

2.5.4 Classification of ECG signals

Several works have demonstrated that ECG signals have unique biometric features related to the electrical operation of the heart. The features used to identify this uniqueness of the ECG signal are categorized as fiducial, non-fiducial and hybrid. Fiducial features relate to the various impulses in an ECG wave that directly map to physiological operations of the heart, and are commonly observed in the time-domain in the form of amplitude, duration between peaks and the morphology of the signal. Non-fiducial features do not rely on the observance in time of the many waves and complexes that form an ECG waveform. These features are commonly extracted from the frequency domain, whereas hybrid features can be a combination of both non-fiducial and fiducial data. In [113], a back propagation neural network classifier was used on a set of non-fiducial features in the form of Fourier Transform coefficients. In [112], signals were pre-processed for artifact removal, followed by the extraction of several fiducial based timing intervals (e.g., sample distance between the P-R segment and R-T segment). Classification was performed on heartbeats using standard linear discriminant analysis. In [114], a survey is provided that gives an overview of the most common fiducial and non-fiducial features used in the most recent studies, while highlighting the use of various classifiers ranging from distance based methods (e.g., K-Nearest Neighbor) and decision tree based-methods to the most successful algorithms that lie in the neural networks domain.

Some of the highest reported classification accuracies for fiducial and non-fiducial features are 99.2 and 98.8%, respectively [115, 116]. However, these works consider computationally heavier feature selection, dimensionality reduction and neural networking classification techniques. Additionally, complex signal processing for waveform artifact removal are commonly used to achieve the near-ideal signal reconstruction. For this work, we consider general, resource-constrained
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mobile and wearable devices that are limited in battery and computational capability. Most recently, [117], presented a system of continuous ECG authentication using motion sensors with a high accuracy of 98.61%. The algorithm presented in [117] utilizes fiducial features extracted with signal processing techniques for a multi-class scenario, training and testing on subjects within a controlled database. Our work presents an alternative to validating the biometric performance of ECG signals by mimicking a more realistic scenario in which authentication can occur in the mobile/wearable domain.

2.5.5 Touch and Proximity-Based Biometric Authentication Systems

In the commercial space, a product exists, known as the Nymi Band. It is a wearable authenticator that is designed to work with other devices (desktop computers, doors, etc.) and perform authentication based on proximity to the locked device. Similar to what we study in this work, the Nymi band employs Lead I ECG measurements for use as a biometric. Once a user is authenticated, it uses Bluetooth Low Energy and NFC to pair with devices running the Nymi supported application. The Nymi band does not perform true continuous authentication, but only authenticates a user once, then granting the user access to trusted devices. Only upon removal does the device cease functionality and must be re-authenticated [?].

The work done in [148], referred to as the Bioamp, proposes a biometric touch sensing mechanism that enables continuous authentication with commodity devices that interact with the user via touch screen (a table running windows 8 in this study). The bioimpedance of the user is measured from a wrist worn device and is configured as the biometric of choice. This data is modulated through the user’s body using an On-Off Keying approach at capacitive coupling body communication, and upon interaction with the touchscreen, enables the device to identify the user. Consequently, dealing with commodity devices with low update rates limits the achievable data rate for the IBC link. Thus, to compensate, the Bioamp sends the biometric information using Bluetooth and only modulates a unique identifier onto the body. The unique identifier is obtained via Bluetooth feedback from the tablet, once biometric data is received. The tablet, which also performs the authentication by classifying feature vectors via Support Vector Machines, searches through a look-up table to match the identifier with the stored biometric values. Once the band is removed, the process for authenticating a user must be repeated, as the Bioamp halts transmission of impedance valued with the tablet device.

Additionally, the authors in [?] demonstrate the possibility of also using the input to a
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commodity device (e.g., fingerprint sensors and touchpads) to capacitively couple information to wireless receivers that are only in contact with the human body. The design leverages a fingerprint sensor, typically found on most mobile devices, to enable the transmission through the body. A custom receiver solution is constructed via software-defined radio, where the system blocks are flexibly controlled and implemented. Initial tests show achievable bit rates up to 30 bits per second over the body. All of these studies show great potential for using biological data as a biometric and the utility of proximity/touch-based authentication. However, the wireless communication methods employed still operates over-the-air in some fashion. The Nymi band leverages Bluetooth for its communication, while the latter two works attempt to take advantage of capacitive coupling. In both cases, eavesdropping and data interception can still occur. For the capacitive coupling methods, additional interference can be obtained from other environmental factors mentioned previously. Thus, none of these methods are suitable for applications where the goal is to secure implant-to-implant, implant-to-surface and surface-to-surface communication by avoiding the over-the-air medium.
Table 2.4: Channel Characteristics Comparison of IBC Methods

<table>
<thead>
<tr>
<th>IBC Method</th>
<th>Channel Modeling Method</th>
<th>Noise Sources</th>
<th>RMS Delay Spread [s]</th>
<th>Propagation Speed [m/s]</th>
<th>Channel Characteristics</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF-NB[^12]</td>
<td>Analytical</td>
<td>Thermal Noise Ext. Interference</td>
<td>~ 10^{-9}</td>
<td>~ 10^{7}</td>
<td>Log normal Shadowing</td>
</tr>
<tr>
<td>RF-UWB[^64]</td>
<td>Empirical</td>
<td>Thermal Noise Ext. Interference</td>
<td>~ 10^{-9}</td>
<td>~ 10^{7}</td>
<td>Two-Ray Propagation</td>
</tr>
<tr>
<td>mmWave[^87][^17]</td>
<td>Statistical</td>
<td>Thermal Noise Ext. Interference</td>
<td>~ 10^{-9}</td>
<td>~ 10^{8}</td>
<td>Cauchy-Lorenz Fading</td>
</tr>
<tr>
<td>US[^62]</td>
<td>Statistical</td>
<td>Thermal Noise</td>
<td>~ 10^{-6}</td>
<td>~ 10^{3}</td>
<td>Nakagami Fading</td>
</tr>
<tr>
<td>GC[^55]</td>
<td>Empirical</td>
<td>Thermal Noise</td>
<td>~ 10^{-9}</td>
<td>~ 10^{7}</td>
<td>AWGN</td>
</tr>
<tr>
<td>CC[^51]</td>
<td>Empirical</td>
<td>Thermal Noise Ext. Interference</td>
<td>~ 10^{-9}</td>
<td>~ 10^{7}</td>
<td>Frequency Selective</td>
</tr>
<tr>
<td>RC[^45]</td>
<td>Empirical</td>
<td>Ext. Interference</td>
<td>-</td>
<td>~ 10^{7}</td>
<td>Frequency Selective</td>
</tr>
</tbody>
</table>
Table 2.5: Comparison of Attenuation Values across IBC Methods

<table>
<thead>
<tr>
<th>IBC Method</th>
<th>Attenuation [dB]</th>
<th>Distance [cm]</th>
<th>Operating Frequency</th>
<th>Tissue Layer</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF-NB</td>
<td>60 (^{13})</td>
<td>15 (^{13})</td>
<td>400 MHz (^{13})</td>
<td>Skin</td>
</tr>
<tr>
<td>RF-UWB</td>
<td>80 (^{14})</td>
<td>12 (^{14})</td>
<td>4 GHz (^{14})</td>
<td>Skin</td>
</tr>
<tr>
<td>mmWave</td>
<td>40 (^{17})</td>
<td>15 (^{17})</td>
<td>60 GHz (^{17})</td>
<td>Skin</td>
</tr>
<tr>
<td>US</td>
<td>10 (^{5})</td>
<td>10 (^{5})</td>
<td>10 MHz (^{5})</td>
<td>Skin</td>
</tr>
<tr>
<td>CC</td>
<td>24 (^{20})</td>
<td>10 (^{20})</td>
<td>40 MHz (^{20})</td>
<td>Skin</td>
</tr>
<tr>
<td>GC</td>
<td>40 (^{23})</td>
<td>10 (^{23})</td>
<td>40 KHz (^{23})</td>
<td>Skin</td>
</tr>
<tr>
<td>RC</td>
<td>12 (^{20})</td>
<td>15 (^{20})</td>
<td>50 MHz (^{20})</td>
<td>Skin</td>
</tr>
</tbody>
</table>
Chapter 3

Understanding the Behavior of Galvanic Coupling Signal Propagation

3.1 Experimental Model Validation

Understanding the behavior of a wireless channel is paramount in designing capable communication systems that achieve the best possible performance. Based on the analytical modeling conducted in [84], initial experimental trials are conducted with a porcine tissue, to validate the behavior observed. A porcine tissue is selected due to the similarities between its dielectric properties and that of human tissue. This model, illustrated by figure 3.1, accounts for the homogeneous tissue layers of skin, fat, muscle and bone, each of which is modeled as a combination of electrical circuits with impedance values derived from dielectric properties and tissue dimensions. A 2-port network approach is used to compute signal gain and phase while varying certain parameters (center frequency, electrode separation, electrode location, electrode dimensions, etc.) The experimental setup is depicted in figure 3.2. An Analog Discovery™, by Digilent, Inc., was used to generate the sinusoidal function used at the Tx terminal. At its base there is a Spartan-6 Xilinx FPGA. The device uses its own in-house signal generation software, known as WaveForms™, with a sampling rate up to 100 MHz. The use of baluns, (Schaffner IT239), are present in order to isolate the common ground return paths of the transmitter and receiver. Using this software, a continuous sine wave, with a 5 V-pp amplitude, was generated at a center frequency within the allotted spectrum for GC-transmission (100kHz to 1MHz). After measuring the received peak-to-peak voltage, the sine wave was incrementally shifted, in 100KHz spacing, to a different center frequency. These measurements
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Figure 3.1: Tissue Equivalent 3D Multi-layered Circuit Model for Human Forearm

Figure 3.2: Experimental setup, including porcine tissue

were repeated for different distance configurations. The methods of channel modeling referred to in Section 2.5 are also highly complex and well inclusive of the dielectric properties of human tissue that may affect signal propagation. However, they do not make mention of the fundamental properties of wireless channels (e.g., amplitude-fading statistics and multi-path delay spread) that need to be considered in the eventual design of a communication system. Additionally, obtaining results through the experimental method described above yields very limited information. Thus, an improved empirical modeling method is implemented and described in the following subsection. A

3.1.1 Correlative Channel Sounding

Experimental characterization of a human-body-like channel is conducted using a correlative channel sounding method, as opposed to the method selected in section 2.1. When designing and simulating wireless communication systems, it is imperative that accurate channel models are created to fully understand how the signals of interest propagate through the medium. Different types of
channel modeling methods are available for use, and each with their own unique way of uncovering the properties that have an effect on signal performance. The method of channel modeling conducted in the context of this work is known as a Stored Channel Impulse Response method [90]. In this method, a correlative channel sounder (explained below), measures, captures and records the channel impulse response. This approach has two distinct advantages: (1) the measured and stored channel impulse responses are realistic and (2) the stored responses are reproducible, reusable for as long as needed, allowing for the simulation and optimization of different communication systems to take place.

Conceptually, a channel sounding signal is composed of a pulsed transmission that takes places with pre-specified repetition intervals and upon reception of the signals, filters for storage and processing off-site. The type of sounding signal sent is determined by the specific method of channel sounding used [90]. In the rest of this study, the term ‘channel’ exclusively implies the human body channel. Specific to this work, correlative channel sounders are implemented for the sake of exposing channel behavior. These sounders belong to the general category of pulse compression techniques [92]. When a white noise signal \( n(t) \) is applied to the input of a linear system, the output \( w(t) \) can be cross-correlated with a delayed replica of the input \( n(t - T) \), yielding a cross-correlation coefficient that represents a scaled version of the impulse response of the system, \( h(t) \), evaluated at \( T \). Following [92], this can be shown from the definition of the auto-correlation function of the noise \( R_n(\tau) \), as equation 3.1, where \( N_0 \) is the single-sided noise-power spectral density. In the time domain, the system output is given by \( w(t) = h(t) * n(t) \) and it is then cross-correlated with the input to yield equation 3.2, which is proportional to the channel impulse response.

\[
R_n(\tau) = E[n(t) n^*(t - \tau)] = N_0 \delta(\tau)
\]  

\[
E[w(t) n^*(t - \tau)] = E \left[ \int h(\zeta)n(t - \zeta)n^*(t - \tau) d\zeta \right] = \int h(\zeta)R_n(t - \zeta) d\zeta = N_0 h(\tau)
\]  

In practice, the behavior of noise signals, which are added to the channel, must be known at the receiver. Therefore, experimental systems employ deterministic waveforms with characteristics that resemble noise-like behaviors. A widely known type of such a signal is known as binary maximal length pseudorandom noise sequences, or PN sequences, typically generated using shift registers.
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Figure 3.3: Channel sounding generic block diagram.

with the appropriate logic. Using maximal length PN sequences as the signal of interest should yield an auto-correlation function with a high correlation peak at $R_{\tau=0}$ (or zero-shift point), and very low side lobes (high peak-to-off-peak ratio). This property allows for the detection of each multi-path component due to the convolution matched-filter (CMF) in the receiver, correlating the channel output with the originally transmitted PN sequence. Figure 3.3 presents a block diagram depicting a traditional channel sounding architecture.

3.1.1.1 System Overview and Experimental Method

In order to measure the channel impulse responses, PN sequences were transmitted in a single carrier BPSK modulated fashion. The chip duration of the linear generated polynomial PN Sequence of degree $m = 14$, using a linear-feedback shift register, was set to the value of 0.909 microseconds (for a bandwidth of approximately 1.1 MHz). This number allows for measurements to take place in the frequency range used in the galvanic coupling system described in [93], beginning and ending at 100 kHz and 1 MHz, respectively. The lower bound was chosen to avoid interference with medical telemetry devices and human body signals, while the upper bound was set at the frequency where the effects of the human body behaving as a radiating antenna become noticeable, drastically reducing the gain. The carrier frequency of 550 kHz was chosen for modulation as a middle point, in order to observe the entire bandwidth of interest when measuring the channel impulse response.

Initial experiments are carried out on a tissue of a freshly slaughtered swine consisting of skin, fat and muscle layers, with a length of 25.5 cm, a width of 23.5 cm and a varying thickness between 2.5 and 5 cm. The rationale for using a porcine tissue for human-body-like behavior stems from an understanding that the dielectric properties of human skin and porcine skin are very similar.
For future tests, described in proceeding chapters, a more stable synthetic channel is used. The entire experimental setup, including the porcine tissue, is also pictured in Figure 3.2.

The correlative channel sounding test bed implemented utilizes the same setup that can be seen in figure 3.2, albeit with an additional layer of software responsible for transmission and reception of data. The analog discovery device is also enabled to work directly with $R$, but the sampling rate was deemed too low for use in this test. In order to utilize some of the custom signal generation capabilities of the WaveForms software, all signals were generated in and saved into the .csv file format. Each signal is then loaded to WaveForms for reconstruction and generation. It is important to note that experiments were conducted with the use of alligator clips to allow for better contact with the porcine tissue, instead of the electrodes. For power and off-site processing, a personal computer was directly connected with the Analog Discovery device. The received signals are captured in WaveForms in .csv file format and then stored for post processing in . Within , a band pass filter is used to restrict measurements to solely the frequency range of interest. This stage is followed by the convolution matched-filter, then the detection and estimation of the parameters of concern. Transmitter and receiver location, distance and orientation are modified to test received signal strength in various positions. Measurements were taken at distances of 5, 10 and 15 cm in each of the locations and orientations, while keeping a constant electrode separation distance of 5 cm.

## 3.1.1.2 Channel Characterization

Following the acquisition through the WaveForms software, the received signals were post-processed in to obtain the channel frequency response and channel impulse response. Channel characterization work done in concluded from preliminary simulations that the channel of a galvanic coupling based communication scheme is non-frequency selective. In order to verify this claim and that of many other works, the channel impulse and frequency response are presented
in this section. Each of the following channel characteristics will be displayed for different tissue communication scenarios, specifically Muscle to Muscle (MM), Skin to Skin (SS), Muscle to Skin (MS) and Skin to Muscle (SM), with the first layer representing the placement of the transmitter and the second layer mentioned the placement of the receiver. This nomenclature will be used to present results throughout the duration of this work. The measured channel impulse response (CIR) for one tissue communication scenarios (MM) can be seen in Figure 3.5. It can be seen that the high peak-to-off-peak ratio (discussed in Section ??) exists, providing good correlation results from the experiments. All of the CIRs from each communication scenario obtained from the experiments, show a very similar impulse response, indicating the presence of no multi-path in the channel environment.

The corresponding frequency domain representation (Channel Frequency Response), for an assumed transmitter bandwidth of 50 kHz, indicates that the channel is relatively flat within the frequency range of interest. In Figure 3.6 for each tissue communication scenario, the Channel Frequency Response (CFR) exhibits a decreasing gain with frequency. The best channel gain takes place for the communication of MM, with SS having the worst performance in terms of channel gain. It is important to note that within this figure, the CFR for a communication range of 10 cm is presented. Equivalent trends with higher magnitudes for channel gain are presented in the CFRs of each tissue communication scenario captured at shorter distances between the transmitter and receiver. In Figure 3.7, a comparison of the model and experiments used in [93] is presented for center frequency values of 100 kHz and 1 MHz. The difference in channel gain, $\Delta G$, which equals

\[
\tau(s) \times 10^{-5} -4 -2 0 2 4 6
\]

Amplitude (V²)

\[
-0.06 -0.04 -0.02 0 0.02 0.04 0.06 0.08 0.1 0.12 0.14
\]

Figure 3.5: The measured channel impulse response (CIR) for the Muscle to Muscle tissue communication scenario.
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Figure 3.6: Channel Frequency Response (CFR) for all tissue communication scenarios for $d = 10$ cm.

the absolute value of the difference between the experimental channel gain and model channel gain from [93]. The model parameters in this work were tailored to mimic the experimental setup as best as possible. The results, presented for all tissue layer communication scenarios except SM (an experimental outlier in this comparison), indicate that a good matching exists between model and experiments (no more than 10 dB of difference among the different scenarios explored). This difference can be attributed to the lack of the 2-port circuit model’s ability to account for the exposed muscle tissue in the experiments. The Skin to Skin based communication comparison has a very low difference, which may be due the model accounting for the loss at the Skin to Air interface. For the other layers, perfect isolation is assumed. Though this cannot be avoided in our experimental setup, comparison in terms of channel gain show promising results.

3.1.1.3 Path Loss Model Fitting

The human tissue can be characterized as a lossy dielectric propagation medium [95]. As such, the energy of the wave attenuates by a factor $e^{2\alpha d}$, and specifically, the power per unit area flowing past the point $d$ is given [96] by equation 3.3, where $d$ is the linear distance between transmitter and receiver. The magnitude $\mathcal{P}(0)$ is the power per unit area flowing at $d = 0$. If both transmitter and receiver have the same effective area, then the gain can be computed by way of equation 3.4.

\[
\mathcal{P}(d) = \mathcal{P}(0)e^{-2\alpha d} \quad (3.3)
\]
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Figure 3.7: Comparison between channel gain measurements obtained from experiments and the model created in [93] for $d = 10$ cm

<table>
<thead>
<tr>
<th>Layer</th>
<th>$P(0)$</th>
<th>95% confidence bounds</th>
<th>$\alpha$</th>
<th>95% confidence bounds</th>
</tr>
</thead>
<tbody>
<tr>
<td>MM</td>
<td>0.03961</td>
<td>(0.02348, 0.05574)</td>
<td>24.2450</td>
<td>(29.6850, 18.8100)</td>
</tr>
<tr>
<td>MS</td>
<td>0.06061</td>
<td>(0.02769, 0.09353)</td>
<td>24.1250</td>
<td>(31.3650, 16.8800)</td>
</tr>
<tr>
<td>SM</td>
<td>0.01177</td>
<td>(0.007756, 0.01578)</td>
<td>22.9400</td>
<td>(27.4100, 18.4750)</td>
</tr>
<tr>
<td>SS</td>
<td>0.01297</td>
<td>(0.006081, 0.01985)</td>
<td>29.5050</td>
<td>(37.0200, 21.9950)</td>
</tr>
</tbody>
</table>

$A_{dB}(d) = -10 \log_{10} \left( \frac{P(d)}{P(0)} \right) = 20 \log_{10}(e) \alpha d = 8.686 \alpha d$ (3.4)

Based on several measurements in the porcine tissue, the parameters for an exponential fit are determined in equation 3.3. This model provides a simple approximation that is suited for most system design problems. Figure 3.8 presents the set of measurements and their fitting model, that have been constrained to the range of 3 to 15 cm, where it shows a better adherence and is also the scope of our system. The parameters $P(0)$ and $\alpha$, the attenuation coefficient, were obtained for different layers and are summarized in Table 3.1. From Figure 3.8 we observe that the worst case of attenuation at 15 cm is near $-60$ dB. Assuming a transmit power of 0 dBm, a sensitivity of approximately $-30$ dBm in the receiver side would be required.
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Figure 3.8: The measurements and the fitted exponential model for the gain on different layers.

3.1.1.4 Noise Analysis and Capacity Estimation

Another set of measurements were taken in the porcine tissue for the assessment of the noise characteristics, including probability distribution and spectral power.

- **Noise Characteristics** The results show that the noise’s probability density function fits well as a normal distribution. The frequency analysis presents a fairly flat power spectral density with a noise power spectral density dependent on the layer of tissue and in the order of $N_{0,MM} = -107.0 \text{ dBm}$, $N_{0,SS} = -105.5 \text{ dBm}$, and $N_{0,MS} = N_{0,SM} = -102.4 \text{ dBm}$. Therefore, the channel is taken as a zero-mean Additive White Gaussian Noise (AWGN) and treated as such for channel capacity estimation.

- **Channel Capacity** For an AWGN channel we employ the well known Shannon-Hartley formula given by (3.5) to make an estimate of the maximum achievable capacity of the system. The calculations are made using the measured received power $P_{RX}$ for several locations and for a signal covering the whole 900 kHz bandwidth.

$$C = BW \cdot \log_2 \left( 1 + \frac{P_{RX}}{N_0 \cdot BW} \right)$$  \hspace{1cm} (3.5)

Figure 3.9 shows the results for different Tx–Rx combinations whereas Figure 3.10 presents one comparison example of the channel capacity estimation from the experimental data and results from the 2-port circuit model, presented for a center frequency of 100 kHz and the noise levels mentioned previously. Results indicate a similar range of values for capacity estimation, even in the presence of the differences among tissue exposure to the environment that was not modeled in [93].
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Figure 3.9: Channel capacity estimate for different layers and distances.

Figure 3.10: Experimental channel capacity estimate comparison with 2-port circuit model by 93 for $d = 10$ cm and a center frequency of 100 kHz.

3.1.1.5 Discussion of Results

Results indicate that the channel response is relatively flat for the frequency ranges of interest, the noise can be approximated as additive white Gaussian, and the achievable capacity lies in the range of Mbps. The comparison of these experimental results with currently existing analytic channel models and experiments, shows a similar range of achievable capacity, but the channel frequency responses differ in the range of $\pm 5$dB for particular center frequency values. The differences are likely to stem from the lack of literature evidence supporting dielectric properties for porcine tissue, as these characteristics are an important factor within the model. Additionally, the configuration differences between the model and the porcine tissue include: absence of bone
layer, and inner tissue layer exposure to the outside environment. Although these factors are less important when compared to the dielectric properties, it is important to note that likelihood of some contribution to noise and signal interference can be increased. Taking into account all of the comparisons presented in this study, results lead us to believe that the performance of the existing models, though advantageous when estimating channel parameters, may not be able to provide a complete notion of channel behavior for the human body. Thus, for similar methods of modeling, and more complex variants, channel behavior will be validated empirically and modified as best as possible to match the real-world conditions present on the experimental medium.

3.2 GC Analytical Channel Model for Arm-Wrist-Palm

To better mimic the channel that will be used in the application chosen for this study, a new analytical channel model is adopted. This approach relies on a carefully tuned physical layer optimized with respect to the arm-wrist-palm path that will be used to relay the biometric information. For this reason, a 3D analytic model of the tissue pathway is designed in an effort to obtain the channel gain by extending previous formulations reported in [118]. A key assumption made in this work (also validated later through experiments) is that the section of the arm where the biometric signal is picked up and then forwarded through the wrist area is cuboidal in shape (as seen in Figure 3.11). Previous works have considered generalized tissues layered in the form of a perfect cylinder, which does not match well with typical dimensions of the upper forearm. A secondary simplifying assumption is that the thickness and properties of each tissue layer, i.e., skin, fat, tendon, muscle, bone are uniform within the separate partitioned segments that form the arm, wrist and palm, but non-uniform between one another.

Galvanic Coupling requires two transmitting and two receiving electrodes attached on the skin. The three parts of the path: the arm, wrist and palm are studied separately. The first step involved identifying typical tissue thickness and propagation characteristics for each of these three segments. For example, a segment consists of four layers: skin, fat, muscle and bone for the arm and skin, fat, tendon and bone for the wrist (Figure 3.12). Through cubical approximation for the arm and wrist segments, the final representation of the path is simplified to three rectangular shapes with four tissue layers of specific thicknesses. The next step is to analyze the path to estimate its channel gain based on the dielectric properties and dimensions of the tissue layers.
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3.2.1 Tissue equivalent circuit

The formation of a tissue equivalent circuit (TEC) model is used to estimate the channel gain of the path between the arm and the palm that the signal traverses. The impedance between two nodes of the TEC is calculated using an RC-circuit equivalent expression of the given path. The resistance and capacitance values of each path are calculated as $R = \frac{\rho L}{A}$ and $C = \frac{\epsilon A}{D}$.

The resistivity ($\rho$) and permittivity ($\epsilon$) values for each tissue layer are frequency dependent and obtained from a published, standardized database [140]. Here, $L$ represents the length of the path between the two nodes, $A_c$ gives the cross-sectional area of the path, $A$ is the surface area of the capacitance measured and $D$ is the depth of the tissue. After calculating the resistance and capacitance of the path between two nodes, the impedance of the specific path is calculated using $Z = R + \frac{1}{\omega C}$. The impedance, resistivity, permittivity, length, cross-sectional area, surface area and thickness values are inputs to the model from which expressions are derived. The TEC for the entire path requires the calculation of five impedance values: $Z_L$ (longitudinal), $Z_D$ (inter-electrode), $Z_C$ (cross) and $Z_t$ (transverse), trivially obtained by replacing the respective $\rho$, $\epsilon$, $L$ and $A$ values depending on the geometry of the path and using the above standard equations. The topview of the TEC is shown in Figure 3.13. The transverse impedance $Z_t$ is between each layer and its adjacent one.
A 3D representation of the TEC including the $Z_I$ is depicted in Figure 3.14. The nodes are shown via capitalized letters; the arm-wrist and wrist-palm junctions share nodes. Hence, the $Z_D$ and $Z_I$ values are calculated by adding the two inverse impedances of the specific paths, as these impedances are parallel to each other, as also seen in the top-view of Figure 3.13. Nodal analysis is then performed using Kirchoff’s Current Law (KCL) involving the impedances of all paths to obtain an admittance matrix for the entire four-layer, thirty-two node equivalent circuit. For a sample node A, it’s nodal equation is illustrated by equation 3.6. Similarly, the equations for all nodes were derived, but not repeated for space conservation. A simplified representation of the 32 KCL equations is presented within the admittance matrix (A) of the 4-layer 3D circuit and can be examined in equation 3.7.

$$I = \frac{V_A - V_B}{Z_D} + \frac{V_A - V_C}{Z_L} + \frac{V_A - V_D}{Z_C} + \frac{V_A - V_E}{Z_I}$$  \hspace{5mm} (3.6)$$

$$A = \begin{bmatrix}
\sum_{i\in C,L,D,t} Z_{ISA} & -1Z_{Dsa} & -1Z_{Lsa} & \cdots & \cdots & \cdots & 0 \\
-1Z_{Dsa} & \sum_{i\in C,L,D,t} Z_{ISA} & -1Z_{Csa} & -1Z_{Lsa} & 0 & -1Z_{tsa} & \cdots & 0 \\
\cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\
0 & \cdots & \cdots & \cdots & \cdots & -1Z_{tmp} & -1Z_{Cbp} & -1Z_{Lbp} & -1Z_{Dbp}
\end{bmatrix}$$  \hspace{5mm} (3.7)$$

Figure 3.13: Circuit (a) is the TEC of the skin layer in 2D. (b) is the equivalent of (a) used for channel gain calculations.

To obtain the gain of the TEC, $G = \frac{V_{out}}{V_{in}}$, an equivalent T-circuit is derived from figure 3.14 (a) and represented in figure 3.14 (b)). The purpose of this step is to analyze the TEC as a two-port network with z-parameters for the gain calculation. The Z-parameters of the circuit in figure
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Figure 3.14: 3D tissue equivalent circuit with transverse impedances for arm-wrist-palm path. The labeling of transverse impedances is $Z_{t-ij}$, for $i \in \{S, F, M, T, B\}$ corresponding to skin, fat, muscle, tendon, bone and $j \in \{A, W, P\}$ for arm, wrist, palm.

3.14(b) are calculated with equations (3.8) - (3.11) which were derived from a 3-loop mesh current analysis and solved for two-port network parameters.

**Equation 3.8:**

$$z_{11} = R_s + 2Z_{co} + \frac{Z_{in}(Z_{Leq} + Z_{out})}{Z_{Leq} + Z_{in} + Z_{out}}$$

**Equation 3.9:**

$$z_{21} = \frac{Z_{in}Z_{out}}{Z_{Leq} + Z_{in} + Z_{out}}$$

**Equation 3.10:**

$$z_{12} = -\frac{Z_{in}Z_{out}}{Z_{Leq} + Z_{in} + Z_{out}}$$

**Equation 3.11:**

$$z_{22} = 2Z_{co} + \frac{Z_{out}(Z_{Leq} + Z_{in})}{Z_{Leq} + Z_{in} + Z_{out}}$$

In summary, this model takes as input the length of the arm, wrist and palm, the size and distance between transmitting and receiving electrodes as well as the signal frequency. The admittance matrix is then constructed using the equations and inputs of the model. The gain of the entire arm-wrist-palm path is then calculated as the logarithmic ratio of the impedance over the output (between nodes S and T) and that of the input (between nodes A and B). The gain of the channel can be investigated using the model under several configurations and plays an important role in the design parameters of the GC communication system. The 3D arm-wrist-palm model is validated experimentally, in the next subsection, for various configurations. Subsequently, the knowledge of the channel response is utilized in order to test and compare various aspects of the GC communication system design in MATLAB, before an implementation of a real system is demonstrated.
3.2.2 Arm-Wrist-Palm Experimental Model Validation

The proposed model is used to obtain (i) link distances at which the gain of the channel is high (it will be shown later, that the gain shows non-uniform behavior), and (ii) to set the frequency of operation and inter-electrode distance for the design and implementation of a physical system. For these reasons, the performance and predictions of the arm-wrist-palm model must be validated through an experimental setup, shown in Figure 4.6. Here an electrically equivalent synthetic tissue phantom from Syndaver Labs™ is used to represent the human arm, wrist and palm components. For all reported experimental results, the gain between different parts of the phantom is calculated by measuring the channel impulse and frequency response for various communication configurations. This exact process mimics that of the method mentioned in subsection 3.1.1. The Analog Discovery™ is used as both a generator and oscilloscope to transmit the sounding signal and read the signal at different parts of the phantom. Bridging the connection between the phantom and the Analog Discovery, balun circuits (Schaffner IT239) are used to isolate the common ground return paths of the transmitter and receiver. The skin-to-skin gain is of importance to the specific on-skin propagation path to the smart-device. Therefore, the wires of the oscilloscope measuring voltage as well as the function generator are attached to the skin of the phantom.

Observations indicate that the model accurately predicts the gain variation with frequency for both 10 and 15 centimeter distances between transmitting and receiving electrodes (Figure 3.16). Both the experimental setup and the model maximizes the gain between 200 and 500 KHz. The gains experienced over a 10 cm distance are overall higher than those for 15 cm, as expected, and match the experiments with the model predictions. Specifically the focus is on the gain prediction at around 10 cm distance, which is most relevant to the GC-biometric authentication application. Since there is a close agreement with the experimental setup for all frequencies with deviation less than 5 dB, results prove that the proposed arm-wrist-palm model very closely captures the actual channel behavior 

Noting the frequency range where the gain is the highest, a series of experiments were conducted with varying distances to test the reliability of the model as the input parameter of distance changes. The model is validated for varying distances of the palm and the arm, to investigate all possible relative positions of the wristband and the mobile device. Figure 3.17 shows the results of measuring the gain of a sinusoidal input operating at 400 kHz over various distances by altering the palm and arm lengths. Two conclusions are drawn from the results: the model closely predicts the gain with varying distances, though varying the length of the palm portion of the path leads to
Figure 3.15: Experimental setup with phantom for model validation
increasing variation in gain than the arm portion. A possible explanation of this phenomenon is the increased thickness of the palm skin layer. Since the gain is calculated on the skin layer, the thickness of the skin affects the gain to a significant extent.

When varying the arm and palm distances around the wrist, it is observed that the predicted versus observed gain values lie within an acceptable experimental margin of error (Figure 3.18). As is expected, the gain is highest for shorter distances between transmitter and receiver.

Finally, a study of the impact of the inter-electrode separation for the two GC transmitting and receiving electrodes is conducted. Results indicate that greater inter-electrode distance between the individual pairs of the transmitting and receiving electrodes leads to a higher gain between the transmitter-receiver nodes, when all other parameters remain constant. Greater inter-electrode distance in galvanic coupling leads to a higher primary current between the two electrodes, and therefore, a higher secondary current between transmitter and receiver. Similar to the previous experiments, the model predictions match closely with the experimental results, proving once again that the arm-wrist-palm model can predict the channel behavior with high accuracy (Figure 3.19).

Overall, our theoretical arm-wrist-palm model is validated as an accurate predictor of the actual physical channel. This enables various design choices with respect to the communications
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Figure 3.17: Experimental and model data comparison for gain vs. distance, varying palm length (FC = 400 kHz)

Figure 3.18: Experimental and model data comparison for gain vs. distance (FC = 400 kHz)
system design using a combination of the channel model and/or the channel response obtained empirically.

Table 3.2: Tissue thickness values (mm)

<table>
<thead>
<tr>
<th></th>
<th>Skin</th>
<th>Fat</th>
<th>Muscle</th>
<th>Tendon</th>
<th>Bone</th>
</tr>
</thead>
<tbody>
<tr>
<td>Arm</td>
<td>1.00</td>
<td>7.00</td>
<td>15.00</td>
<td>-</td>
<td>20.00</td>
</tr>
<tr>
<td>Wrist</td>
<td>1.00</td>
<td>7.00</td>
<td>-</td>
<td>1.50</td>
<td>15.30</td>
</tr>
<tr>
<td>Palm</td>
<td>1.40</td>
<td>7.00</td>
<td>9.00[138]</td>
<td>-</td>
<td>9.17[139]</td>
</tr>
</tbody>
</table>

3.3 Implantable Sensor Network Design

The next phase of intra-body communication usage will surely extend to scenarios where multiple implantable sensors will be communicating between neighboring nodes. However, a single hop, end-to-end link is not always feasible, due to the attenuation that exists depending on the tissue layer of the body and the link distance. To better understand how one can utilize communication schemes for implantable sensor networks, this section presents a design for a multi-cast and multi-hop communication scheme for intra-body communication. The topology of the sensors is envisioned to be a network of implanted devices that communicate within the human body, between sensors (via
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weak electrical signals), and to the surface of the human body to a relay/gateway node, capable of sending data to remote monitoring personnel using over the air RF. Distinct contributions include: (i) signal reflection and refraction analysis of electromagnetic waves through human tissue boundaries and (ii) the design of a combined multi-hop and multi-cast communication scheme for communication between implants. For these contributions, the human body is modeled as a lossy dielectric medium (tissue conductivity > 0) with four tissue layers (bone, muscle, fat and skin). Using this model, it is determined whether a single transmission that crosses multiple tissue boundaries (e.g., from bone to skin) at a given frequency performs better in comparison to multi-hop forwarding of data and if a single transmission can enable multi-cast communication.

3.3.1 Signal Reflection and Refraction

When electromagnetic (EM) waves interact in a lossy dielectric medium, they are partially reflected, partially transmitted, and even refracted. Reflection, transmission and refraction of a signal within human tissue vary depending on the type of tissue. For communication between implants, the portion of the signal that crosses from one tissue boundary to the next (with the least reflection), has an impact on the signal strength at the receiver, and also, the bit error rate (BER). Thus, studying the reflection and refraction properties used in physics and electrodynamics allow us to analytically verify the reflection levels and how much the signal bends (refracts) as it crosses from one tissue layer to another.

3.3.1.1 Amplitude Reflection Coefficient

The amplitude reflection coefficient (described in Equation 3.12) for normally incident signals, depends heavily upon the angle of incidence of the wave, tissue conductivity, and relative permittivity and permeability of the tissue. The variables $n_1$ and $n_2$ represent the refractive index (the speed of light in vacuum divided by the speed of light in the medium). The refractive index describes how much refraction occurs in a certain medium and relates the speed of light in that particular medium. For example, a medium with a refractive index of 1.5, equates to the speed of light traveling 1.5 times slower in that same medium $[103]$. In a lossy dielectric, the index of refraction relates to the tissue conductivity ($\sigma$), relative permeability ($\mu_r$) and complex relative permittivity ($\varepsilon_r^*$) shown in Equation 3.13 and Equation 3.14, but simplifies to Equation 3.15, since human tissue is considered a non-magnetic medium (unit value of relative permeability).
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Figure 3.20: Example diagram of a multi-cast communication scheme

\[
\Gamma_n = \frac{n_2 - n_1}{n_2 + n_1} \quad (3.12)
\]
\[
n = \sqrt{\epsilon_r \mu_r} \quad (3.13)
\]
\[
\epsilon_r^s = \epsilon_r - \frac{j\sigma}{\omega} \quad (3.14)
\]
\[
n = \sqrt{\epsilon_r^s} \quad (3.15)
\]

Figure 3.20 exemplifies a scenario for calculating the values for the amplitude reflection coefficient for normally incident waves. The mediums of interest are muscle \((n_1)\) and fat \((n_2)\), and the incident wave travels from within the muscle to the fat tissue layer, a higher refractive index to a lower refractive index. Note that for normally incident waves, the angle of reflection \((\Theta_r\) dotted black arrow), refraction \((\Theta_R\) solid black arrow), and incidence \((\Theta_I)\), are all equal to the value of zero degrees.

Figure 3.21 depicts the trend of the absolute value of the amplitude reflection coefficient for various combinations of tissue layer boundaries, as frequency increases within the desired range (100 kHz to 1 MHz) for using low level electrical current for intra-body communication [105]. The frequency dependent nature of the dielectric properties of human tissue has a direct effect on the amount of reflection that occurs at each level of human tissue boundaries. Furthermore, results from this graph aid in the selection of the best operating frequency to conduct communication within
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Figure 3.21: Tissue boundary diagram for (a) normally incident reflection and refraction and (b) obliquely incident reflection and refraction

the human body. It can be observed, that for some situations, it may be more advantageous to use higher/lower frequencies, where reflection is less, for signals propagating from one layer to another.

When electromagnetic (EM) waves interact in a lossy dielectric medium at an oblique angle, polarization occurs, splitting the electric field into two separate components. These components are distinguished and given their names based on their orientation in relation to the plane of incidence. The p-wave has its electric field (E-field) oriented parallel (p) to the plane of incidence (i.e., the incident, refracted and reflected signals lie in the plane of incidence), and the s-wave has its E-field perpendicular (s) to the plane of incidence \[101\]. The result of this behavior of EM waves yields two unique reflection coefficient values for each type of polarization. Equation 3.16, exemplifying the parallel polarization reflection coefficient, differs from Equation 3.12 by showing a dependence on the angle of incidence and angle of refraction. The angle of reflection, as governed by the Law of Reflection, will be equal to the angle of incidence, at the interface normal to the surface where reflection takes place. For the sake of this work, all the analysis done with respect obliquely incident EM wave reflection will assume parallel polarization, and adhere to the laws governing the angle of reflection.

\[
\Gamma_p = \frac{n_1 \cos \theta_R - n_2 \cos \theta_I}{n_1 \cos \theta_R + n_2 \cos \theta_I} \tag{3.16}
\]

There exists a special case for parallel polarization, in which no reflection occurs, and the signal is entirely transmitted from one tissue layer to the other. This phenomenon occurs when the
angle of incidence is equal to what is known as the Brewster Angle. The Brewster angle loses its complete meaning in reference to lossy dielectrics, where the values for relative permittivity and the refractive index are complex, but for angles of incidence that approach the Brewster angle, the reflection coefficient values are significantly less in comparison to other choices for the angle of incidence \[^{101}\]. The Brewster angle for each of the tissue boundary reflection scenarios can be shown by the sudden drop in reflection for a particular value of incidence angle. Figure \[\ref{fig:reflection_coefficient}\] shows an example of the behavior of the reflection coefficient for obliquely incident waves for two values of the frequency range of interest, 100 kHz and 1 MHz, for varying values of the angle of incidence. Here it is observed that the effects of the Critical angle (if \(n_2 < n_1\), incidence angles exceeding the critical angle will be subject to the phenomenon of total internal reflection) and the Brewster angle taking place. For the boundary conditions of skin to air and muscle to fat, observations show a total internal reflection, yielding a reflection coefficient of 1 for significantly smaller angles of incidence, compared to other tissue layer boundary scenarios. This information allows for the selection of the best incidence angle that grants the most acceptable level of reflection, subject to the Critical angle.

### 3.3.1.2 Angle of Refraction

Refraction occurs in a lossy dielectric medium under the influence of EM waves due to the change in speed a wave undergoes when moving from one medium to another. An increase in speed tends to make the signal refract away from the line normal to the surface of the two mediums (Figure \[\ref{fig:refraction}\] where the angle of refraction is greater than the angle of incidence), while a decrease tends to cause bending towards the normal (angle of incidence greater than angle of refraction). The angle...
Figure 3.23: Obliquely incident reflection coefficient vs angle of incidence

of refraction can be solved from Equation 3.17, where $\Theta_R$ and $\Theta_I$ represent the angle of refraction and angle incidence respectively. Figure 3.24 (a-d) plots the angle of refraction versus the operating frequency, for various values of the angle of incidence, subject to the critical angle. The results from these figures help us determine the proper receiver placement for the implant at the next hop tissue layer, while still taking into account possible levels of reflection.

$$n_1 \sin \Theta_I = n_2 \sin \Theta_R$$  \hspace{1cm} (3.17)

### 3.3.2 Multi-cast and multi-hop communication design

After careful study of the behavior of EM wave propagation in human tissues (modeled as lossy dielectrics), the placement of next hop nodes in an intra-body communication scheme and the enabling of a multi-cast communication scheme can be developed. Combining the two approaches can yield a scenario where one can communicate from the inner most tissue layer, to a relay node on the skin surface, with a minimum number of hops, by exploiting the refraction and reflection properties of each human tissue layer. By choosing the appropriate transmission power and frequency, part of the signal can permeate through the tissue boundary to an intended receiver (e.g., from muscle to fat), while the controlled reflected component returns back to a second receiver implant (e.g., also embedded in the muscle). Therefore, two different implants can potentially receive the same transmitted data, though at the loss of signal amplitude. Using the phenomenon of refraction, a sensor can be optimally aligned to receive the dominant component of the signal propagating into the direction of the next hop implant.
Although it is possible to send one signal from the bone layer to propagate to the relay node on the skin surface, several parameters have to be taken into account to ensure successful reception, with a reasonable BER. One must not to choose a particular angle of incidence that could influence unwanted levels of reflection, and at the same time, may yield an angle of refraction that could possibly cause total internal reflection at the subsequent tissue layer. Thus, a good tradeoff must be achieved between the angle of incidence, angle of refraction and the reflection coefficient values in order to design a good multi-cast and multi-hop communication scheme. Many scenarios may exist in which multi-cast and multi-hop communication design parameters can be chosen. These scenarios may change based on the foreseeable application for intra-body communication. One particular design case is presented in Figure 6, where the evaluation of reflection and refraction is provided for a frequency of 300 kHz, and the name designation for each tissue boundary is the following: Skin-Air (SA), Fat-Skin (FS), Muscle-Fat (MF) and Bone-Muscle (BM).

3.3.3 Performance Evaluation

In the setup presented in the previous section, multi-cast communication is deemed possible for a maximum of 1 to 3 nodes, and a maximum of two hops (Bone-Muscle and Skin-Air) is used to facilitate an entire chain of communication with an on-surface relay. The presence of large values
CHAPTER 3. UNDERSTANDING THE BEHAVIOR OF GALVANIC COUPLING SIGNAL PROPAGATION

Figure 3.25: Multi-cast and multi-hop communication scheme

<table>
<thead>
<tr>
<th>Layer</th>
<th>$\Gamma$</th>
<th>$\theta^I_1$</th>
<th>$\theta^R_1$</th>
<th>$\theta^I_2$</th>
<th>$\theta^R_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Skin (Final hop)</td>
<td>0.9410</td>
<td></td>
<td></td>
<td>71.8°</td>
<td>14.3°</td>
</tr>
<tr>
<td>Fat (Multicast)</td>
<td>0.162</td>
<td>$\theta^I_M$</td>
<td>$\theta^R_M$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Muscle (Multicast)</td>
<td>0.5457</td>
<td>5°</td>
<td>71.8°</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Bone (First hop)</td>
<td>0.6787</td>
<td>$\theta^I_B$</td>
<td>$\theta^R_B$</td>
<td>$\theta^I_B$</td>
<td>0°</td>
</tr>
</tbody>
</table>

Of reflection (subject to the other parameters mentioned) limit the use of sending one signal from the Bone layer to traverse through all the tissue layers, and to the relay on the surface. Even though larger angles of incidence in the bone layer can cause fewer reflections, they all lead to total internal reflection taking place within the muscle layer, when approaching the fat tissue layer boundary; thus, validating our approach for the first hop of data forwarding. Taking advantage of the dielectric properties of the Muscle, a multi-cast transmission with 1 node is performed (within the muscle), that enables communication with 3 other nodes (1 also within the muscle, 1 within the fat and 1 within the skin). This pathway serves as the intermediate flow of data between the first and last hop. Lastly, if an attempt is made to use the angle of incidence for the skin-air interface as the value of the angle of refraction from the Fat-Skin Interface, then total internal reflection occurs, and the relay cannot be reached. Therefore, a separate and final hop must exist, forwarding the data along a normally incident path to the on skin relay node. Overall, the studies presented in this chapter can be applied towards work examining beam forming techniques, the use of smart antennas, and state of the art algorithms for antenna switching, to leverage the applicability of such methods in this proposed design.
CHAPTER 3. UNDERSTANDING THE BEHAVIOR OF GALVANIC COUPLING SIGNAL PROPAGATION

Table 3.3: Electrical properties and dimensions of tissue layers

<table>
<thead>
<tr>
<th>Tissue Layer (thickness -mm)</th>
<th>Conductivity (S/m)</th>
<th>Relative Permittivity (F/m)</th>
<th>Refractive Index</th>
</tr>
</thead>
<tbody>
<tr>
<td>Skin (1)</td>
<td>0.001934</td>
<td>1090.5</td>
<td>33.02</td>
</tr>
<tr>
<td>Fat (9)</td>
<td>0.02469</td>
<td>44.075</td>
<td>6.64</td>
</tr>
<tr>
<td>Muscle (25)</td>
<td>0.40693</td>
<td>5226.9</td>
<td>72.3</td>
</tr>
<tr>
<td>Bone (20)</td>
<td>0.021407</td>
<td>191.44</td>
<td>13.84</td>
</tr>
</tbody>
</table>

Table 3.4: Electrical properties and dimensions of tissue layers

<table>
<thead>
<tr>
<th>Implant</th>
<th>Angle of Reflection</th>
<th>Angle of Refraction</th>
<th>Linear Distance from [Source,Relay] (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>S</td>
<td>5°</td>
<td>71.8°</td>
<td>[0, 37.77]</td>
</tr>
<tr>
<td>S₁</td>
<td>5°</td>
<td>N/A</td>
<td>[2.63, 35.84]</td>
</tr>
<tr>
<td>S₂</td>
<td>0°</td>
<td>0°</td>
<td>[37.12, 1]</td>
</tr>
<tr>
<td>S₃</td>
<td>71.8°</td>
<td>N/A</td>
<td>[58.04, 29.14]</td>
</tr>
</tbody>
</table>

3.4 Summary

The results outlined in this chapter show that it is possible to model and validate the understanding of the behavior of the human body medium. Analytical expressions are derived from a 3D tissue equivalent circuit model of the inner layers with various dimensions and frequency dependent dielectric properties. These parameters are used to calculate the impedance values across various paths that current can traverse within the body. Additionally, the model has the capability to illustrate the channel gain behavior under different electrode configurations (terminal separation, link distance and size). To calculate the channel gain and phase, the relationship of Ohm’s Law and Nodal analysis is utilized, where output voltages and currents are calculated. The performance of this model is compared to measurements produced from a wide-band channel sounder, in which the channel impulse and channel frequency response are obtained and stored for reproducible experiments. Results indicate a reasonable agreement between model and experiments from both a porcine channel and a synthetic tissue channel, specific to the GC-application chosen for this study. Knowledge of channel behavior will be used to fuel the design choices for communication systems and modulation schemes explored in the next chapter.
Chapter 4

Galvanic Coupling Narrowband Communication System Design

4.1 Galvanic Coupling System Design

Creating sustainable IBC-based applications requires careful design of the physical layer link. This chapter demonstrates the construction of a communication link through a synthetic human tissue phantom using the GC-concept and the transceiver architecture needed for its implementation. System design and implementation takes place in two phases: (i) simulation within the MATLAB environment with the stored channel impulse response obtained in (ii) a macro-scale sized testbed employing software defined radios for flexible testing of run-time communication. Phase I is conducted to compare selected modulation schemes when propagating known data through the galvanic coupling human body channel. Phase II is designed to validate communication system performance for simulated architectures and link layer protocols, and perform real-time link quality assessment to ensure target BERs are met. Additionally, phase II implementation is designed to introduce more complex algorithms for adaptively selecting the best transmission parameters based on channel observations.

4.1.1 Comparison of GC-Modulation Scheme

Previous channel sounding studies with respect to the GC human body channel, indicate that it can be approximated as AWGN for transmission bandwidths on the order of 50 kHz. Therefore, for the simulations conducted within MATLAB, analysis is focused on narrow-band
CHAPTER 4. GALVANIC COUPLING NARROWBAND COMMUNICATION SYSTEM DESIGN

modulation schemes, for their ability to provide less complex transceivers in the case of a real-world implementation. Specifically, the traditional PSK and QAM modulation schemes are chosen and their performance is compared for target BER/PER metrics, subject to output transmit power levels, as this is a primary design feature when transmitting on or within the human body.

The carrier-based or passband modulation schemes in this chapter are categorized as bi-dimensional modulation with two orthogonal base functions (modulators) given by (4.3). Here, $\sqrt{\frac{2}{T}}$ is a normalization factor and $T$ is the integration period. A set of basis functions of an $N$-dimensional space is called orthogonal if the following relation in (4.1) holds. This operation is known as inner product of two functions and is usually denoted as $\langle \phi_n(t), \phi_m(t) \rangle$ [146]. Also, if the output equation (4.2) holds, the basis functions are also called orthonormal. This concept, although simple in appearance, is a fundamental tool for modulation and demodulation of signals, particularly in correlation-based, coherent reception. With respect to carrier-based modulation, the inner product operation yields the result illustrated in equation (4.4), indicating that they are orthonormal.

$$\int_{t_1}^{t_2} \phi_n(t)\phi_m^*(t)\,dt = \begin{cases} 0 & n \neq m \\ K_n & n = m \end{cases}$$ (4.1)

$$K_n = \int_{t_1}^{t_2} |\phi_n(t)|^2 \,dt = 1 \text{ for all } n$$ (4.2)

$$\phi_1(t) = \sqrt{\frac{2}{T}} \cos(2\pi f_c t)$$

$$\phi_2(t) = \sqrt{\frac{2}{T}} \sin(2\pi f_c t)$$ (4.3)
Figure 4.2: IQ demodulator diagram, with the matched filter $g^*(t)$.

\[ \int_0^T |\phi_n(t)|^2 \, dt = 1 \quad \text{for} \quad n = 1, 2 \tag{4.4} \]

The function $\phi_1(t)$ is called In-phase and $\phi_2(t)$ in Quadrature, therefore the signal $s(t)$ in this bidimensional space is known as IQ modulation \cite{120} and is described in equation 4.5, where $s_I$ and $s_Q$ are the in-phase and the quadrature components of the modulated signal, respectively. More generally, the function $g(t)$ represents the filter of the signal, chosen to adjust the desired spectral properties, now represented in a form presented in equation 4.6, where the components $s_{ij}$ correspond to the $i$-th symbol of one of the $j = 2$ base functions, $\phi_{1,2}(t)$. In this work, the option exists to apply a raised cosine filter. The set $s_i = \{s_1, s_2, s_3, \ldots, s_m\}$ is known as constellation points and their values are given as complex numbers. A simplified, generic block diagram of a the transmitter of an IQ modulation system is shown in Figure 4.1 and the receiver in Figure 4.2. Both QAM and PSK are part of the general group of IQ modulation schemes. Their transmission and reception structural diagram is essentially the same, and the main difference lies within the mapping of the symbols in both amplitude and phase for QAM and strictly phase for PSK.

\[ s(t) = s_I \cos(2\pi f_c t) + s_Q \sin(2\pi f_c t) \tag{4.5} \]

\[ s_i(t) = s_{i1} g(t) \cos(2\pi f_c t) + s_{i2} g(t) \sin(2\pi f_c t) \tag{4.6} \]

4.1.1.1 Experimental Setup and Simulations

Simulations are conducted for up to $10^6$ bits, encompassing PSK and QAM modulation schemes up to a modulation order of ($M = 16$). The signal is attenuated and noise is added according
CHAPTER 4. GALVANIC COUPLING NARROWBAND COMMUNICATION SYSTEM DESIGN

Figure 4.3: Passband Modulation Simulation Block Diagram

Figure 4.4: BER vs Transmit Power for PSK and QAM Modulation Schemes
CHAPTER 4. GALVANIC COUPLING NARROWBAND COMMUNICATION SYSTEM DESIGN

with the results of the channel characterization in Chapter 3. These results are benchmarked based on the target bit error rate of $10^4$ bits. Additionally, the following parameters are selected for all simulated modulation schemes: (i) skin-to-skin (SS) propagation path, (ii) 10 cm link distance, (iii) 5 cm electrode separation, (iv) Symbol rate of 150 kbps (w/ Tx filter) and 200 kbps (w/out Tx filter), and (v) 200 kHz (approximately).

Figure 4.4 presents the measured bit error rate vs output transmit power for GC-IBC. To ensure that the energy per bit is kept constant across all modulation schemes, the output power is scaled based on the number of bits per symbol ($P_{out} = P\log_2(M)$). Results indicate that BPSK modulation outperforms the higher modulation (as expected) in terms of power efficiency. In terms of bandwidth efficiency, the other modulation schemes offer higher data rates, while affording the capability to fit more bits per second in the allotted bandwidth. However, this increase in constellation size increases the system complexity, as well as decreases the distance between each point in the constellation. Both of these factors contribute to a higher power expenditure due to more components in the system and more symbol confusion in the presence of noise, respectively. For the sake of implantable and on-body galvanic coupling communications, a greater constraint is placed on power as opposed to bandwidth. As a result, modulation schemes with a performance profile similar or better to what is achieved by BPSK will be considered. Further evaluation of modulation schemes in proceeding chapters will study the tradeoff in performance between non-coherent and coherent detection schemes to uncover if additional power savings can be met.

4.1.2 GC-IBC Experimental Testbed Architecture

The GC-IBC testbed, implemented via a SDR platform, provides the flexibility needed to optimally design low power communication systems that are suitable for IBC. Towards this end, the hardware support package of the Communications System Toolbox in MATLAB is utilized to demonstrate the effect of various modulation schemes on a prototype testbed. Specifically, the USRP N210 by Ettus Research™ is used as the Software Defined Radio platform to emulate an implanted/on-body sensor. As shown in the GC-IBC testbed block diagram in figure 4.5 and the physical representation of the testbed in figure 4.6, we use USRPs to form a communication link, which are in turn connected to host laptops.

The communication system employs PSK modulation schemes and accommodates data rates up to 200 kbps. Elements of the communication system include, but are not limited to, bit generation, preamble insertion, raised cosine filtering and baseband modulation on the transmitter
side (USRPTx). In the receiver chain, components such as automatic gain control, phase and frequency offset compensation, and baseband demodulation exist. Within the host machine, we utilize the MathWorks® USRP Hardware Driver (UHD) interface, so that data can be transmitted and collected directly from/within MATLAB. Within the USRP, low frequency daughter boards, the LFTX and LFRX, are used for the transmitter and receiver, respectively. The daughter boards operate within the frequency range of DC to 30 MHz, well within the permissible frequency band used for GC-based communication. However, the LFTX and LFRX have almost no internal gain, thus requiring signal amplification. We use external amplifiers (provided by MiniCircuits®) to account for the lossy nature of the channel and the limited hardware capability. Integrated with the Tx is the ZFL-500+, an SMA connector based Power Amplifier (PA) that has a maximum power output of 9 dBm. On the Rx side, a Low Noise Amplifier (LNA), the ZFL-1000LN+ is used, consisting of a noise figure of 2.9 dB.

The human body equivalent channel, takes the form of a custom synthetic tissue plate
model (20 cm x 20 cm), provided by SynDaver Labs\textsuperscript{TM}. The tissue layer thicknesses for skin, fat and muscle are 0.1 cm, 0.5 cm and 25 cm, respectively. The synthetic tissue is composed of salt, water and fiber and is constructed to account for the dielectric properties of actual human tissue to a good approximation. Bridging the connection between the channel and USRP, balun circuits (Schaffner IT239) are used to isolate the common ground return paths of the transmitter and receiver. In order to inject electrical current into the tissue, electrodes provided by TENSPros are used in this study. The electrode separation for both sides of the link is set to a distance of 5 cm, and the distance of the communication link can be varied up to the length of the tissue, but is fixed at 10 cm for comparison purposes.

A MATLAB-based GUI, is used to allow the user to select the physiological data/image for transmission through the channel. Based on the type of physiological information that is chosen for transmission, specific modulation schemes are selected for the link (based on application dependent data rate requirements). The GUI at the Rx displays the BER rate and compares the real-time BER measurement to results from previous channel sounding experiments. An instructional video of the entire system operation is available here \cite{121}.

Results (figure 5.5) are presented in terms of BER for DBPSK modulation. Three distinct transmit power levels are used versus the traditional Signal-to-Noise Ratio method of BER assessment, which is more advantageous when monitoring the transmit power levels used in IBC links (considering tissue exposure limits). The reasonable level of agreement between results obtained from previous stored channel response experiments and real-time channel assessment validates the merit of the GC-IBC testbed for the development of future systems. Although successful run-time transmission
CHAPTER 4. GALVANIC COUPLING NARROWBAND COMMUNICATION SYSTEM DESIGN

and reception of data through the channel has been observed, performance of the link does not depend solely on the channel itself. Data underflows and overflows can occur at the transmitter and receiver, respectively, based on how fast the host machine can process and deliver frames of data from the MATLAB environment to the USRP via Ethernet. This phenomenon causes significant delays at the transmitter and dropped frames at the receiver, the latter of which has a greater affect on error rate measurements. Thus, in some situations, the source of ”error” in reception is difficult to identify. This bottleneck when utilizing the SDR-based tested has motivated a smaller form factor design, for an application specific testbed (presented in a later chapter), in which this drawback can be avoided and greater system portability can be obtained.

4.2 Summary

The results presented in this chapter show that it is possible to simulate and validate various modulation schemes in the MATLAB environment and with Software Defined Radios, respectively. In both cases, a version of the actual human body channel is utilized. The former, exists as a stored channel impulse and frequency response from previous channel sounding experiments on synthetic and porcine tissue. The latter, is implemented on a macro-scale testbed where the actual synthetic tissue phantom is used to propagate known information, from transmitter to receiver, at run-time speeds. The following steps involve the migration to a smaller, portable run-time tested, for an application specific use. For this design, a heavier constraint is placed upon power efficiency and system complexity. Thus, within the following chapter, additional low power modulation schemes will be evaluated and compared for use within an actual custom transceiver solution.
Chapter 5

Wearable System Design for a Galvanic Coupling Application

From body worn sensors to active controlling of smartphones via touch inputs, large amounts of personal data is continuously communicated with the outside world. The prevalent method of authentication is to actively type in the shared password on a hand-held device or transmit secure keys through wireless channels. However, these forms of inputs are highly susceptible to intentional eavesdropping and privacy-manipulation enabled by off-the-shelf but sophisticated software defined radios. Instead of using an over-the-air channel, the main scope of this dissertation proposes a radically different communication pathway to transmit biological signals in the form of biometric information collected (from implantable or on-body sensors) in real-time for authentication. This application-specific approach for utilizing galvanic coupling is chosen for the potential of modulating a non-radiating electrical signal with information propagated along a human-limb path \[106\]. By setting the frequency in the range 100KHz-1MHz, the signal energy is retained within the human tissue. Thus, without any external RF emissions, it mitigates sniffing attacks. The only way to intercept the signal is to physically touch the subject with a pair of receiving electrodes.

5.1 Problem Formulation

The dramatic increase in the every-day usage of personal mobile devices that carry and communicate sensitive data necessitates robust authentication systems. Although research on key generation and sharing for wireless sensors has made rapid strides, increasingly sophisticated RF sniffing attacks coupled with limited computational resources within the sensors, pose practical
Figure 5.1: A biological signal, ECG in this use case, is acquired by a wrist-worn device and its feature set is transmitted through the human skin to the palm where it can be used to authenticate or actuate (unlock devices) by physical contact.

Biometrics have long since played a role in securing such devices, by leveraging individual characteristics such as fingerprints, retina/iris and facial features. Despite the ubiquitous nature of these systems for commercial use, mobile devices still utilize auxiliary passwords for additional forms of security (e.g., two-factor authentication). The use of traditional passwords often yields drawbacks in terms of user recollection, password strength and scale when a user is registered to multiple platforms. Fingerprints, although unique, are transferable and can be left on various surfaces/objects that users have daily interaction with. Such vulnerabilities are exploited via plastic, latex or gelatin based molds, used to forge a copy of the individuals biometric. Additionally, high resolution images and/or videos have found common use in counterfeiting retina and facial recognition software.

Additionally, most authentication techniques in hand-held electronic devices and wearables rely on one-time secret entry. This means that once the device is unlocked, it is accessible to any user. There is an implicit assumption that the devices remains authenticated after the first instance of user identification. However, numerous scenarios can take place in which the security of a device can weaken overtime. Currently, continuous authentication for this target use case is seldom explored. This motivates the need to explore alternative security measures through physiological data, generated by the human body, that have immense potential in biometric applications. Specifically, this chapter performs a case study on the well-known electrocardiogram (ECG) signal, a bioelectrical signal describing the activity of the heart, as a biometric for authentication. ECG is highly personalized and depends on the physical anatomy, structure, function, conductivity of one's own heart, position/orientation of the heart within the torso, the torso geometry, lung size and the conductivity of the surrounding muscle tissue, all of which results in high levels of individual
CHAPTER 5. WEARABLE SYSTEM DESIGN FOR A GALVANIC COUPLING APPLICATION

distinction [110] [111].

• Proposed Approach: ECG with GC transmission The proposed authentication system, shown in Figure 5.1, is composed of a wearable device on the forearm that acquires the ECG signal, processes it to extract the target feature set and then modulates a weak electric current to create the non-radiating GC waveform. This GC signal is transmitted wirelessly through the arm, wrist and palm of the subject to the receiver, be it a data logging entity, actuation interface (e.g., door handle) or a smart-device. The receiver has a GC front-end that captures the signal, and then feeds it to the classifier for pattern matching against a known template of expected ECG features. The ECG classification model leverages a one-vs-rest strategy, where correlation coefficient based template matching is employed between enrollment and test signals.

5.2 Choice of GC Modulation Scheme

The GC channel model, proposed in chapter 3.2.1 and experimentally validated in chapter 3.2.2, are the starting points of the communications systems design. The analytical channel model is employed to select which center frequency, inter-electrode separation and amount of arm and palm channel length needed to achieve best channel gain results. In general, throughout literature, the GC human body channel is approximated as AWGN [141]. However, this chapter leverages the stored channel frequency response, as opposed to static gain values from the model, to test multiple candidate narrow-band modulation schemes and various transmission bandwidths (those that yield flat fading characteristics). Since there are tighter demands on power consumption as opposed to available bandwidth, non-coherent systems and techniques with lower modulation orders are more suited for this application. The goal is to achieve a target BER of $10^{-4}$ or better by selecting one of: Binary Frequency Shift Keying (BFSK), Binary Phase Shift Keying (BPSK) and On-Off Keying (OOK). The best combinatorial approach from these parameters is selected, considering the tradeoffs between performance and system implementation complexity.

5.2.1 Candidate modulation schemes

• Binary frequency shift keying (BFSK): This modulation technique is characterized by it’s high power efficiency and ease of implementation. The number of independent frequency oscillators needed for this modulation is proportional to the number of bits allocated per symbol, and therefore it consumes more power as the number of bits increases. We restrict our current implementation
Figure 5.2: Block diagram of the binary fsk transmitter and receiver

Figure 5.3: Block diagram of ook modulation transmitter and receiver

of FSK to a modulation order of 2 (M = 2), as higher values of M in FSK are not considered due to increased system complexity and bandwidth usage [142]. The transmitter consists of a standard two oscillator setup, where the modulator toggles the carrier frequency based on the data to be sent. The receiver architecture performs its filtering and detection directly at the carrier frequency values, eliminating the need to translate the signal to baseband.

**On-Off Keying Modulation (OOK):** The evaluation of a non-coherent approach for OOK modulation is selected, such that one can eliminate the need for a phased locked loop and the requirement of a quadrature path. The receiver has a bandpass filter, envelope detector and a comparator, where threshold-dependent decision is performed per bit. Unfortunately, adopting non-coherent modulation schemes with simpler architectures have a direct trade-off with performance (e.g., data rate and spectral efficiency). In order to justify the need to perform GC data transfer with non-coherent modulation techniques, a performance comparison analysis is conducted with coherent systems in the proceeding subsection.

**Binary Phase Shift Keying Modulation (BPSK):** While theoretically offering both low power and complexity, the latter may be impacted by the observed channel characteristics where a significant
phase shift occurs at various frequencies. In order to take full advantage of the BPSK performance, a phase locked loop (PLL) is also implemented to track the changes in phase and adapt accordingly. However, integrating a PLL into the architecture introduces additional complexities that can lead to higher power consumption.

### 5.2.2 Power consumption analysis

In order to further investigate the power consumption of the system under the three possible modulation schemes, the power required for all analog front-end components is modeled and in the aforementioned modulation schemes in the MATLAB environment. This model is based on [143], where bandwidth, peak-to-average-power ratio (PAPR), modulation order, and channel gain at various frequencies have an affect on the performance. The overall energy consumption of a system, which depends on its front-end components, is calculated in [143] with equation (5.1).
CHAPTER 5. WEARABLE SYSTEM DESIGN FOR A GALVANIC COUPLING APPLICATION

\[ E_c = ((P_{Tx} + P_{Out}) \cdot T_{on}) + (P_{Rx} \cdot R_{on}) \] (5.1)

The variables \( P_{Tx} \) and \( P_{Rx} \) represent the total power consumption from the transmitter and receiver electronics, respectively. These values are obtained from the summation of the contribution from the individual components of each system design, which consist of a variation of one or more of analog front end devices modeled above. The output power, \( P_{out} \), that is required for the desired level of link reliability is obtained from the data presented in Figure 5.5. The factors of \( T_{on} \) and \( R_{on} \), represent the time that the transmitter and receiver elements are active, respectively and are derived from packet size and data rate.

The front-end components of the design based on the modulation schemes investigated in this section can be found in Table 5.2. The hardware components are simulated with their respective systems for the purpose of calculating the BER of each modulation scheme.

The transmitted and received power of a system designed to perform BFSK modulation would be calculated as follows:

\[ P_{Tx_{BFSK}} = P_{DAC} + 2(P_{mix} + P_{VCO}) + P_{PA} \] (5.2)
Table 5.2: Front-end components for system design with BPSK, OOK and BFSK modulation

<table>
<thead>
<tr>
<th>Modulation</th>
<th>Transmitter</th>
<th>Receiver</th>
</tr>
</thead>
<tbody>
<tr>
<td>BPSK</td>
<td>DAC, mix, VCO, PA</td>
<td>ADC, PLL, mix, VCO, filter</td>
</tr>
<tr>
<td>OOK</td>
<td>DAC, VCO, mix</td>
<td>ADC, filter</td>
</tr>
<tr>
<td>OOK</td>
<td>DAC, 2 mix, 2 VCO, PA</td>
<td>ADC, 2 filters</td>
</tr>
</tbody>
</table>

\[
P_{Rx_{BFSK}} = P_{ADC} + 2P_{filter} \tag{5.3}
\]

The power of each component is calculated using equations from \[143\]. Those are defined as \(P_{PA}\), \(P_{mix}\), \(P_{filter}\), \(P_{DAC}\), \(P_{ADC}\), \(P_{PLL}\) and \(P_{VCO}\) and represent the power consumption for the power amplifier (PA), the mixer, the analog filter, the digital to analog converter (DAC), analog to digital converter (ADC), phase locked loop (PLL) and the voltage controlled oscillator (VCO), respectively.

Similarly, with the components listed in Table 5.2 and equations (5.2) - (5.3), the transmitted and received powers for all modulations schemes are calculated. The energy consumption of the entire system for the three modulation schemes described earlier is calculated in the respective column of Table 5.1 using equation (5.1).

### 5.2.3 Performance Comparison

Constrained by the physical limitations of the synthetic human tissue phantom and the ADC sampling rate of Teensy MCU, the stored channel response generated from the following parameters is used to compare various systems in the MATLAB environment: a center frequency of 100 kHz, an inter-electrode separation distance of 3 cm, and a total channel length of 10 cm. Figure 5.5 plots BER versus transmit power (an important limiting factor when considering GC communication within the body). To fairly compare each technique, the symbol rate is fixed to a value of 50 kbps. Results indicate that the BPSK system offers an improvement in terms of power efficiency and a gradual improvement over bandwidth efficiency, as it is able to achieve the same target BER while occupying less spectrum, and transmitting with less power. However, the performance increase is only approximately 3dB when compared to the modulation techniques of OOK and BFSK. At the same time, it can be seen that the energy consumption of BPSK is approximately 18x of the OOK for the GC channel (Table 5.1). These results indicate that the marginal increase in BER performance does not justify the need to increase system complexity, hardware footprint and energy consumption.
for an application emphasizing low-power operation. Based on these characteristics, it is determined that the OOK modulation is better suited at the PHY layer for integration into an embedded system platform for ECG signal transmission via GC.

5.3 ECG Classification Case Study

5.3.1 Experimental Setup

Proof-of-concept experiments are performed with ECG signals to demonstrate the possibility of recreating similar levels of ECG classification accuracy with those values reported in literature, albeit with lower computational complexity. Within the MATLAB environment, a hybrid method of fiducial and non-fiducial features were extracted from the ECD-ID Database provided by PhysioNet, which consists of measurements sampled at 500 Hz, taken over a 6 month period from 90 subjects in a Lead I configuration. The experimental setup for this portion of the study utilizes 20 subjects with 5 recordings each, evaluating them individually against all other subjects and recordings of each subject. An additional 67 subjects with 1 recording each are strictly used to test known instances of the opposing class.

The accumulation of ECG data is applied to the use case depicted in Figure 5.6. In contrast to authentication methods that utilize a database to compare pre-existing biometric inputs from multiple subjects, this form of pattern recognition is designed to distinguish a particular class from
CHAPTER 5. WEARABLE SYSTEM DESIGN FOR A GALVANIC COUPLING APPLICATION

Figure 5.7: Multiple ECG segments from one subject with varying heart rate

all other possible inputs into the system (one vs. rest or one-class classification). This strategy solely necessitates learning from the class or person of interest and no information about the other classes are present. The basic approach involves producing a result that is a representation of a confidence score used in the decision making. Decision boundaries are formed around the class of interest, with the objective of accepting or rejecting the incoming data based on the measure of certainty. This method is chosen for the sake of resembling a suitable scenario for mobile device authentication, where typically only one user is registering for access.

5.3.2 QRS peak detection and ECG segmentation

The dominant time domain peak of the ECG signal illustrated in Figure 5.8 is known as the QRS complex. Accurate detection of this fiducial descriptor is performed on all recorded data, properly segmented into non-overlapping portions, for which we dub in this work as ”beats”. Evidence from literature supports that the QRS complex remains unaltered by the variation in heart rate [116]. Exemplified in Figure 5.7 each graph of the figure shows the segments of the same subject with varying heart rate in the time (left) and frequency (right) domain. Observations in time indicate heart rate variation causes change in other portions of the signal (the P and T waves) but not on the QRS complex, which remains stable. In frequency, the same magnitude can be recognized in the region that corresponds to the location of the QRS complex. This invariance

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property motivates the use of the QRS complex as an important feature for differentiating subjects by their ECG signals. Figure 5.8 shows an example of an extracted QRS complex for 2 different subjects, visually substantiating the difference in the QRS complex varies among the subjects.

QRS beat detection is performed using a wavelets transform to decompose the signal into different time-varying frequency components in MATLAB using the \texttt{modwt} function, representing an undecimated wavelet transform \cite{134}. The output is truncated to only contain information related to the passband that contains the most QRS complex energy (approx. 5.6 to 22 Hz). To reconstruct the signal, the \texttt{imodwt} function in MATLAB is used in conjunction with the \texttt{sym4} wavelet, chosen due to its resemblance in morphology to a typical QRS complex. The reconstructed waveform is then passed through a non-linear squaring operation to be used within a threshold-based peak detection algorithm. Accuracy within 50 milliseconds is chosen as a tolerance for accurate localization of each peak (based on experimental observations for the worse case detector performance). The QRS detection algorithm is designed to be quick and accurate, supporting the decision to create a low-complexity, fast ECG authentication algorithm for mobile systems. Proceeding QRS Complex Detection, the ECG segmentation process is tasked with producing non-overlapping segments from the entire recording and stores additional localization data representing the extracted QRS complexes.
Figure 5.9: Feature extraction and classification process for one-class classifier
5.3.3 One-class classification performance

The one-class classifier (OCC) implementation is described in Figure 5.9 with a simplified enrollment and a verification/authentication signal. In the enrollment phase, QRS detection and ECG segmentation proceed the averaging off all separated beats, forming a template. During this enrollment phase, the segments of one recording are cross-correlated on a segment-by-segment basis to determine an average value to set as the detection threshold for the subsequent step. In the authentication stage, a similar process is conducted, with only a minor alteration. In order to simulate a faster authentication phase by a mobile system, only beats (designated as one PQRST segment) from the first 2-5 detected segments are averaged to form a testing template. The two templates are time normalized and then cross-correlated to produce a similarity score. The score is compared to the threshold learned in the enrollment phase and any correlation coefficients above the threshold are accepted as the target class, while others are rejected.

The performance of the one-class classifier for all data sets tested solely against subject 1 lies within in Table 5.3, indicating that the subject was correctly identified with an accuracy of 95.89%. For a sample set of 20 subjects, with 1 recording from each subject set as the target class and all other recordings used as test cases, gives a classification accuracy above 97% for all subjects (Figure 5.10), and a false positive rate not exceeding 2.6% (Figure 5.11). The ECG authentication algorithm designed specifically for this application provides high accuracy and low false positive rate,
CHAPTER 5. WEARABLE SYSTEM DESIGN FOR A GALVANIC COUPLING APPLICATION

Figure 5.11: ECG classification false positive rate of 100 recordings from 20 subjects

<table>
<thead>
<tr>
<th>Classified</th>
<th>Actual</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Not Sub. 1</td>
</tr>
<tr>
<td>Sub. 1</td>
<td>88</td>
</tr>
<tr>
<td>Not Sub. 1</td>
<td>3026</td>
</tr>
</tbody>
</table>

Total Accuracy = 95.89%

Table 5.3: Confusion Matrix for One-Class Classification using 4 beats in the authentication
CHAPTER 5. WEARABLE SYSTEM DESIGN FOR A GALVANIC COUPLING APPLICATION

Figure 5.12: Functionality overview of galvanic coupling biometric system

similar to pre-existing algorithms, while offering low-complexity feature selection and classification tailored to a mobile authentication system.

5.4 Galvanic Coupling System Design

The prototype system consists of a Teensy microcontroller unit (MCU) with supporting analog front-end hardware for signal modulation and detection. At the fundamental level, the components within the system have the flexibility to implement either of the modulation schemes discussed in 5.2.

5.4.1 System overview

Figures 5.12 and 6.1 illustrate the design and implementation of a complete system that authenticates a device by using the unique features of the ECG signals. The entire GC wireless transmitter has been implemented, where the signals are confined within the human body. The transmitter MCU is initially configured with a biometric signature unique to the individual (i.e., the stored ECG template created from the algorithm described in Figure 5.9). The ECG data is quantized and converted to its binary representation for interpretation by the Teensy™ device. To prepare for signal transmission, the pulse width modulation (PWM) output of the Teensy-Tx, combined with controllable internal logic, is toggled based on the binary data and pre-selected bit duration. A low pass filter removes the signal harmonics in the frequency domain, while preserving the center tone \( f_c = 100kHz \). The resulting transmitter output from the Teensy MCU, initiated via Bluetooth connection to a smartphone, is a passband OOK modulated signal. The data is sent through the wireless channel within the payload of a frame, which consists of a preamble (13-bit Barker code) for
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Figure 5.13: Physical ECG-GC authentication system testbed

synchronization, data length field, payload (64 bits) and an 8-bit CRC. Once the signal propagates through the human body channel, the analog front-end receiver hardware utilizes a high pass filter (HPF) to remove any low-frequency noise associated with power-line interference and/or baseline drift. An amplifier (MAX4488 by Maxim Integrated™) counteracts the attenuation of the channel and high pass filter, while raising the signal level to meet the turn-on voltage requirements of the schottky diode (used in the subsequent system block). The role of the envelope detector circuit is to convert the signal back to baseband and to remove any possible carrier wave oscillations that could have an adverse affect on the comparator (MAX4488) trigger. The threshold is dynamically controlled through a potentiometer, and is designed to reproduce the original binary bit sequence delivered from the transmitter. The data packets are detected by the Teensy-Rx, decoded and forwarded through the serial port to MATLAB, where processing is offloaded for the ECG authentication algorithm.
5.4.2 Transceiver performance

- **Hardware Validation**: The transmitter and receiver performance is illustrated through the output of an Analog Discovery module. The OOK pulse and its corresponding center tone are generated at the transmitter end. Figure 5.15 portrays the transmitter and receiver performance in both the time (top) and frequency (bottom) domain. The latter provides validation of the comparator output, returning the original bit sequence that is fed into the ADC of the Teensy-Rx. The former depicts the transmitted signal output. The time domain plots indicate the equivalent bit duration of the two signals, and the frequency domain plots indicate the appropriate operating frequency and that the received signal has been translated from passband to baseband.

- **Power Consumption**: We determine the current and subsequently the power consumed by the transmitting and receiving hardware using the shunt resistor method. We leverage high-side insertion,
or placing the 1Ω shunt between the positive supply and the load, in order avoid potential ground loops. For our setup, we used the adafruit PowerBoost 1000C circuit with a 3.7V rechargeable battery to power the Teensy, which then delivers 3.3 V of power to the remaining Tx and Rx circuitry.

We use the Analog Discovery module to measure the voltage drop across the shunt resistor and employ it’s hardware support package with MATLAB to export the data into a host PC for processing. We then calculate the current consumed by using the relationship between resistance and voltage in Ohm’s law \(V = IR\). We subsequently calculate the power consumption by multiplying the current drawn by the supply voltage \(P = VI\). For the transmitting Teensy, we consider the transition between an idle mode and a mode in which the Tx is on and actively sending modulated bits. At the receiver, we account for a transition between an idle mode, an Rx listening mode and an Rx on (receiving) mode. During the idle modes, the MCU is placed in a deep sleep state, via software, where most peripherals are completely shut off. In the Rx listening mode and Rx on modes, we achieve additional energy saving by placing the MCU in a low power state where the peripheral clocks run at a lower speed.

During the operations in which the system is not transmitting or receiving, the operational amplifiers used within the Tx and Rx chain are in shutdown mode and consume approximately .01μA. In the active states of transmission and reception, the operational amplifiers draw 2.2 mA of current. Table 5.4 presents the current and power consumption for all of the aforementioned modes for the transmitter and receiver, with the inclusion of their accompanying front-end hardware. We observe that the transmitting mode consumes the most power, followed by the receiving mode. This behavior is caused by the inability of the Tx-Teensy to initialize the necessary transmitting functions in a low power state and thus, the normal operating mode must be used. Figures 5.17 and 5.4.2 illustrate the total measured current drawn by Teensy Tx and Rx, plotted against time, for different modes of operation. Using the relationship between, power and time, \(E_c = [P_{Tx} \cdot T_{TXon}] + [P_{Rx} \cdot T_{RXon}]\),

<table>
<thead>
<tr>
<th>Mode</th>
<th>Current</th>
<th>Power</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tx Idle</td>
<td>938.87 μA</td>
<td>3.098 mW</td>
</tr>
<tr>
<td>Rx Idle</td>
<td>3.29 mA</td>
<td>10.87 mW</td>
</tr>
<tr>
<td>Rx Listen</td>
<td>14.307 mA</td>
<td>47.20 mW</td>
</tr>
<tr>
<td>Tx On</td>
<td>39.77 mA</td>
<td>131.25 mW</td>
</tr>
<tr>
<td>Rx On</td>
<td>17.32 mA</td>
<td>57.15 mW</td>
</tr>
</tbody>
</table>
we determine that the total average energy consumption of the Tx and Rx are 523.95 mJ and 565.07 mJ, respectively. When comparing the simulated energy consumption results for OOK modulation, presented in Section 5.2, there exists a significant performance difference. In order to more accurately compare these two energy consumption numbers, it is best to implement the system on an application-specific integrated circuit (ASIC) and apply industry caliber research and development for ideal power savings to be achieved [136].

• **Energy Per Bit:** We define the energy per bit as the ratio of the amount of energy delivered into the synthetic tissue medium over the total amount of bits sent through that same medium. We derive its value from the energy consumption values of the transmitter, while in an active state, over the total amount of bits used to transmit one sample ECG recording. Using this description, we measure the energy per bit to be approximately 51 µJ/bit. For an RF-intrabody communication link presented in [50], whose signals still substantially leak into the area surrounding the human body, an energy efficiency of 2.90 nJ/bit is achieved. The difference in performance can be attributed to the RF case employing an ASIC implementation, while our design, which is built from COTS device is not optimized for low power operation.

• **Bit Error Rate:** We tested the end-to-end link performance by using actual ECG signal traces. Figure 5.19 depicts the BER for various transmit powers for a link distance of 10 cm. The transmit power is calculated by measuring the impedance of the synthetic tissue phantom across the input terminals of the transmit electrodes, and using the equation $P_{tx} = V^2/R$. To alter the transmit power levels, we vary the output voltage through the use of a basic attenuator circuit. Observations indicate the experimental capability of achieving a BER of $10^{-6}$ for a transmit power of -2 dBm (0 BER for simulated trials) and no errors for a transmit power of 0 dBm for both experimental and
simulated tests. Longer distances can be supported, at the expense of higher BER and consequently, higher transmit power, or by introducing forward error correction code.

The end-to-end performance is also tested by using actual ECG signal traces to test the classification of subject 1 against the rest of the 19 subjects. The performance metrics (accuracy, true positive rate and false positive rate), depicted in Figure 5.20, display the results of verification experiments tested for link distances of 10 and 15 cm. Observations indicate higher accuracy and true positive values for the 10 cm link distance, a result of the increased gain of the channel which leads to fewer channel induced errors. Similarly, the false positive value is higher for the longer distance. For the second experiment various coding schemes are tested for forward error correction. Performance of the system is evaluated for uncoded transmission, and transmission with Hamming codes for forward error correction. Hamming code rates of 1/3 and 4/7 are notated as FEC1 and FEC2, respectively. As seen in Figure 5.21, introducing the simplistic FEC1 scheme provides negligible improvement for the overall classification performance. By increasing the redundancy,
FEC2 yields an increase in the classification performance, true positive and a reduction in the false positive value. These results lead to the conclusion that the inclusion of a simple FEC scheme to the signal transmitted and received through galvanic coupling can have significant affects on the overall performance of the system, specifically in scenarios where longer propagation distances are desired.

5.5 Summary

Several system components are studied that contribute to the overall system design. First, in chapter 3, the design and experimental validation of an electrical-equivalent circuit model of the human arm-wrist-palm tissue channel and characterization of the GC signal transmitted through these tissues is performed. The results of these channel studies are used to optimize the design of transmission parameters for the galvanic coupling communication system, such as the forward error correction code (FEC) and modulation scheme. Through experimental traces, it is shown how an ECG signal can be used as a reference signature to identify individuals with over 97% classification accuracy in authentication trials performed over 80 random test subjects. Based on
the ECG classification trends exhibited in the machine learning domain, it is highly probable that an improvement in intra-class performance for the true positive rate can be achieved in the presence of additional features and improved signal processing algorithms to account for the acquisition scenarios that do not provide the best signal fidelity. The next step involves the implementation of a proof-of-concept, the end-to-end authentication system using the Teensy\textsuperscript{TM} microcontroller unit with supporting analog front-end hardware for transmission and reception of GC signals. On-Of Keying (OOK) modulation is identified as the preferred technique for this application, and thus a proof-of-concept testbed composed of an embedded system implementation with supporting hardware. The secure method of transmitting biometric information (i.e., ECG signals) via GC intra-body communication technology is also demonstrated with standard ECG trace files and tissue phantoms, confirming over 93\% correctness in authentication tests. The system functions properly to transmit an ECG signal over a 10 cm human tissue path and classifies the subject-of-interest correctly with 93.7\% accuracy. Implementing a Hamming code with a rate of 4/7 on the signal transmitted through the intra-body GC channel increases classification accuracy for longer propagation distances.
Chapter 6

Eavesdropping Susceptibility

To substantiate the claims of a secure biometric transmission system with minimal signal leakage outside the body, we experimentally confirm the level of received signal strength that may be overheard by an over-the-air entity and one that comes in direct contact with the human body medium.

6.1 Over-the-Air Signal Susceptibility

We evaluate the over-the-air signal susceptibility of our system against the work conducted in [148–151], which leverage the use of capacitive coupling (CC) and apply the work in [152] to motivate the use of a hybrid coupling method (GC configuration at the transmitter and CC configuration at the receiver). The purpose of this comparison is to evaluate the susceptibility of our design to over-the-air (OTA) sniffing against the most commonly used non-RF intra-body communication methods in wearables. We perform a series of experiments where we measure the received signal strength (RSS) and BER of an adversarial receiver with the adequate hardware and software means to attempt signal interception.

Thus, the malicious receiver is designed as an exact replica of the receiver presented in Chapter 5 only it does not come in physical contact with the phantom. The adversary also consists of two copper electrodes measuring at 3x3 cm with an electrode separation of 5 cm when configured for GC. The RSS and BER are measured at various arrangements representing distances (as shown in Figure 6.1 and 6.2) displaced horizontally (with the palm facing up), vertically and on the outer part of the phantom arm, opposite the side of the tissue where the true receiver is located. Results from Figure 6.3 illustrate that the RSS at the adversary receiver decreases as we move away from
the transmitter in any direction. For all three displacement orientations, the signal strength at the adversary Rx is lowest for the GC method, followed by the hybrid method and then lastly, the CC method. This behavior can be explained from as the hybrid method forms a return path through the environment (as in the CC case) but its range is confined to a portion of the body, that in which the transmitter is located, as opposed to the entire body. Ultimately, these results show the use of GC to transmit a signal in the body provides a more secure channel where the average RSS at the adversary Rx is at the noise floor at a distance of 13 cm, making decoding very difficult. To confirm the level of decoding at the adversary Rx, we conduct the second set of experiments to measure the adversary BER at different positions around the phantom, identical to those in the RSS studies. The results of Figure 6.4 indicate that the adversarial Rx with GC configuration does not have the capability to decode the signal at any distance or orientation. The CC configuration, however, allows for significantly lower BER values in distances up to 15 centimeters with nearly a \(10^{-4}\) BER at distance of 1 cm away from the tissue, which can lead to a sniffing attack. We thereby conclude that the choice of GC as the method of intra-body communication for our system is advantageous in establishing a secure channel to transmit biometric data.

### 6.2 Contact-based Susceptibility

We evaluate the impact of an adversarial receiver that lies in direct contact with the human body, during data transmission, at various places throughout the channel. We utilize a synthetic tissue slab (also provided by Syndaver Labs) to act as the medium in which the transmitter and intended receiver operate. The adversarial receiver, whose system design is an exact replica of the intended receiver, takes the form of the synthetic tissue hand that we have utilized in all testbed
experiments conducted thus far. The electrode placement options include fingertip, palm and wrist electrode positioning. However, prior experimental work indicated that the fingertip method of electrode placement yielded the highest results in terms of received signal strength. Thus, we connect both the signal and ground electrodes on the fingertips of the index and middle finger, with an electrode separation of 4 cm. We place the adversarial receiver in contact with synthetic tissue slab at positions (shown in Figure 6.7) dubbed inside line-of-sight (iLOS) and reverse line-of-sight (rLOS), and measure the RSS and BER over distances displaced from the location of the Tx. Figure 6.5 and 6.2 depict the RSS and BER plots versus displaced distance from the transmitter, respectively. Here we observe that the adversarial Rx can still obtain a BER of $> 10^{-2}$, up to 7 centimeters away from the transmitter. The values observed in this experiment are higher than those seen in the OTA test, as we expect. However, the adversarial receiver does not obtain a significant advantage in choosing to intercept data using one method versus the other. The results shown in Figure 6.5 and 6.2 also reveal that when there is contact on the channel of interest, the intended receiver also suffers a loss in RSS and increase in BER (as opposed to the OTA case). As the adversarial Rx moves further away from the Tx, the RSS of the intended Rx begins to improve, subsequently causing a decrease in BER.
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Figure 6.3: Average BER at over-the-air adversary Rx at various distances from Tx

Figure 6.4: Average RSS at over-the-air adversary Rx at various distances from Tx

However, this performance when compared to the case when there is no contact with the channel is still significantly degraded. These experiments indicate that the presence of human body contact (whether accidental or intentional), in close proximity to the Tx and intended Rx during any part of the data transmission process, can cause erroneous results. To further understand this behavior, we perform channel sounding experiments in a similar manner to [141], in the subsequent chapter, to gain a better understanding of what phenomenon is occurring at the channel level.

6.3 Summary
CHAPTER 6. EAVESDROPPING SUSCEPTIBILITY

Figure 6.5: Average RSS at contact adversary Rx at various distances from Tx

Figure 6.6: Average BER at contact adversary Rx at various distances from Tx

Figure 6.7: GC Communication testbed for contact-based experiments

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Chapter 7

Detection of Adverse Channel Conditions

In this chapter, we present findings that provide a categorization of the channel behavior observed when direct contact is made during the data transmission and reception process. We also present a means of detecting and mitigating the effects of direct contact by providing alternative transmission parameters for the sending and receiving of biophysical data.

7.1 Impact of Channel Contact

We design and implement a correlative channel sounder in order to measure and record (for post processing) the channel impulse and frequency response of the tissue slab channel in situations where there is no contact and direct contact. This method of channel modeling is based on the observed behavior when applying a white noise signal to the input of a linear system. The output of such a system is cross-correlated with a delayed replica of the input producing a scaled version of the system impulse response. However, in practice the goal is to leverage a deterministic waveform that resembles noise-like characteristics. Therefore, we design a maximum length Pseudorandom Noise (PN) sequence using a linear-feedback shift register (with \(2^m - 1\) cyclic shifts and \(m = 14\)) to modulate and send through the channel. We specifically employ single-carrier BPSK modulation (\(f_c = 550kHz\)) on this PN-sequence with a chip duration of 250 nanoseconds, which equates to approximately 4 MHz of bandwidth. We adhere to the general design criteria of the PN-sequences for channel sounding, which prefers a signal that yield an auto-correlation function with a high correlation peak at the zero-shift point and produce a high peak-to-off-peak ratio. We
CHAPTER 7. DETECTION OF ADVERSE CHANNEL CONDITIONS

Figure 7.1: Channel impulse response for different contact scenarios

Figure 7.2: Channel frequency response for different contact scenarios

selected the signal bandwidth to ensure we could resolve any possible multi-path components while adhering to the frequency range of our GC channel (100 kHz to 1 MHz).

We perform channel sounding measurements for three specific cases that relate to scenarios of no contact and channel contact. The baseline case, or case 1, we take measurements when no external contact is made with the channel, while maintaining the same link distance and electrode separation. For cases 2 and 3 we repeat the same measurement setup but with external contact that causes a visible increase and decrease in the RSS, respectively. Using a similar testbed setup to what is depicted in Figure 6.7 we replace the custom Tx and Rx hardware each with an Analog Discovery module. The PN-sequence was designed in the MATLAB environment, exported to the .csv file format, and loaded into the signal generation software on the Analog Discovery. At the receiver side, we use the oscilloscope feature of the Analog Discovery to export the data into MATLAB for the
CHAPTER 7. DETECTION OF ADVERSE CHANNEL CONDITIONS

Figure 7.3: FEC performance during contact-induced channel experiments

generation of the channel impulse response and channel frequency response.

Figure 7.1 depicts the channel impulse response for all three cases, superimposed for visual purposes. Results for the baseline case present two multi-path components with almost equivalent amplitude. When examining the channel frequency response for the baseline case, the AWGN channel nature that is observed within the GC-research community is validated for our target range of transmission bandwidth (1 Khz to 50 Khz). However, we observe that the presence of touch changes the nature of the impulse response. For example, an external touch in a location that causes an increase in RSS (channel geometry dependent) alters the impulse response such that the first tap is greater in magnitude when compared to the second tap. The opposite case takes place in locations in which external contact decreases the RSS. These effects are also displayed in the channel frequency response presented in Figure 7.2. For our maximum transmission bandwidth range, we still observe a relatively flat fading channel. However, the channel gain can increase or decrease based on location-dependent, external contact with the human body medium. These results validate the behavior observed at both the intended Rx and adversarial Rx that we present in our contact-based experiments within Chapter 6. Regardless of the intention behind eliciting contact upon the human body medium, the intended receiver can still undergo periods of performance degradation. In order to account for such an occurrence, we design a system for detecting periods of channel contact that produce an abrupt change in received data quality. We also provide insight on the appropriate levels of forward error correction (FEC) codes to apply in order mitigate this drop off in performance. Both of these contributions are described in detail in the subsequent subsection.
7.2 FEC Selection for Adverse Channel Conditions

We experimentally induce contact-based events into the synthetic tissue medium, during data transmission, specifically focusing on improving the BER performance of the intended receiver. The preamble sequence, originally used solely for frame synchronization, is modified to encompass the PN-sequence used in the previous subsection. The inclusion of a CRC8 (located in the trailer) for error detection is used to track an instantaneous packet error in between correct frames. If this occurrence continues, the received preamble is cross-correlated with the original data sequence to obtain the channel impulse response and examine if a morphological change has occurred that indicates a touch. In our current system, we experimentally uncover what change in transmission parameters are necessary to combat the channel degradation brought on by touch. Thus, we evaluate the impact of various levels of FEC to determine what code rates are appropriate for this use case. We compare the performance of the system for uncoded transmission, and transmission with Hamming codes for rates of 1/3 and 4/7, notated as FEC1 and FEC2 respectively. As seen in figure 7.3, introducing the simplistic FEC1 scheme provides a large improvement to the average BER from approximately 43% to 18%. The adoption of FEC2 yields very little increase in the performance, resulting in a BER of approximately 17%. Thus, we included trials with a larger, more efficient Hamming code of (15,11), which we call FEC3. These results show a further reduction in BER, validating the need to incorporate adaptive transmission of multiple FEC schemes, for future system designs, where channel state feedback is provided by the receiver.
Chapter 8

Adaptive Modulation with Galvanic Coupling Ultra Wideband

The nature of the human body presents several factors that can attribute to channel attenuation. Tissue hydration, tissue thickness and how these parameters change with respect to the location on the body, all impact the observed impedance levels, subsequently affecting the amount of gain observed at the receiver. The system designed in 5 is designed for a specific set of channel conditions. The implementation of the non-coherent energy-based detection has a threshold statically set for commonly observed voltage levels at a particular distance for one medium. If any of the aforementioned variables were to change, detection performance would drastically decrease. To successfully account for changes that can occur in the human body channel, this chapter will focus on techniques to conduct adaptive modulation and transmission schemes. An important consideration will also be given to the possible hardware constraints that might affect what schemes can be implemented in a GC-transceiver.

8.1 Design Approach

Typical adaptive transmission systems estimate the channel at the receiver and provide feedback to the transmitter to improve performance (e.g., min. BER, max throughput, min. transmit power, etc.) based on an of the estimation of the current channel conditions. Several degrees of freedom exist for augmentation such as constellation size, transmit power, symbol time, coding rate and/or a combination of these methods [144]. The chosen degrees of freedom demonstrated in this chapter consist of a combination of constellation size, coding rate, transmit signal bandwidth
CHAPTER 8. ADAPTIVE MODULATION WITH GALVANIC COUPLING ULTRA WIDEBAND

and center frequency while leveraging Ultra wideband (UWB) modulation techniques. This form of modulation offers the benefit of reduced system complexity (carrier-less), low power (low duty cycled pulses), and high data rates for short range communications at the expense of additional bandwidth [153]. Similar to [3], a pseudo-random noise sequence is employed for channel sounding measurements at the receiving node. Additionally, this sequence is used as the preamble for data transmission, providing the capability of frame synchronization and frequency offset estimation, if necessary. Upon observing the channel frequency response and providing the appropriate feedback, specific UWB pulse properties and/or modulation schemes will be changed to align the transmission bandwidth within the portions of the channel bandwidth that contain the highest gain values.

8.2 Galvanic Coupling Ultra wideband

The -10dB fractional bandwidth of the GC-channel exceeds 20 percent, thus classifying it as ultra wideband as specified by the FCC. Thus, the aforementioned benefits of UWB systems and those afforded by nature of galvanic coupling can be exploited. Outlined in equations 8.1 - 8.3, \( B_f \) is the fractional bandwidth, and \( f_H \) and \( f_L \) are the upper and lower cutoff frequencies of the \(-10\) dB transmission band, respectively. Lastly, \( BW \) represents the bandwidth and \( f_c \) is the center frequency.

\[
B_f \geq 0.2 \quad (8.1)
\]

\[
B_f = \frac{BW}{f_c} = \frac{(f_H - f_L)}{(f_H + f_L)/2} \quad (8.2)
\]

\[
B_f = \frac{BW}{f_c} = \frac{900}{550} = 1.636 \gg 0.2 \quad (8.3)
\]

Ultra wideband modulation consists of the transmission of very short pulses, with low duty cycles with high energy concentrated inside the pulse duration, \( T_p \), yielding very high bandwidths in the frequency domain. Data transmission can be conveyed in the amplitude, phase, position, or shape of the pulses, or a combination of any of these methods to achieve higher order modulation schemes. One example of a typical signal that can be considered for UWB transmission is the Gaussian pulse and its more commonly used higher order derivatives. To produce the desired spectrum allocation, the order of the Gaussian derivative and pulse width of the time domain signal can be altered. This type of transmission also does not require the use of additional carrier modulation [81], and is considered a baseband approach.
CHAPTER 8. ADAPTIVE MODULATION WITH GALVANIC COUPLING ULTRA WIDEBAND

Figure 8.1: First derivative Gaussian pulse with 5 $\mu$s pulse duration

8.2.0.1 Ultra Wideband Pulse Selection

For Galvanic Coupling-UWB, the type of pulse that will be considered plays an important role in the functionality of each candidate modulation scheme (described in the following subsection). Specifically, a pulse becomes highly desired if it is bandlimited and contains high levels of energy concentration in $T_p$. It is also desirable to have an orthogonal set of pulses, that can provide higher orders of modulation and even enable multiple access in the presence of multiple users (attained by assigning a single/group of orthogonal pulses per user). In this scenario, data transmission among the different user signals will not interfere with one another. The two waveforms considered for this work are Gaussian pulses and Prolate Spheroidal Wave Functions (PSWF), along with their corresponding derivatives. Previous work with respect to GC-IBC [53] have used rectangular pulses, which often lead to unwanted harmonics and thus an inefficient usage of the allocated transmission bandwidth. The utilization of PSWF and Gaussian pulses in the context of intra-body communication have seldom been explored.

- Gaussian Derivatives The Gaussian pulses are frequently used in the UWB systems since they can be easily generated by pulse generators (when compared with the rectangular pulses with very short rise and fall time) [145]. Its function is given by equation 8.4 and 1st through 4th derivatives are represented by equation 8.5. These signals are orthogonal between the $n$-th and $(n + 2)$-th order derivatives, therefore a set of no more than two orthogonal waveforms can be obtained. In order to accommodate the signal’s desired spectral mask, the order of the derivative and the pulse width
become adjustable parameters. For example, it has been determined in \[154\] that a 12th derivative Gaussian pulse, with a duration of 1 nanosecond can occupy the spectral mask of the typical UWB RF-based environment (3.6 to 10.1 GHz band). In this study, it has been determined through experimentation that a 5-microsecond pulse of first order Gaussian derivative (depicted in Figure 8.1) is appropriate for the 900 MHz of bandwidth in the 100 kHz to 1 MHz GC-channel.

\[
p(t) = \frac{A}{\sqrt{2\pi \sigma^2}} e\left(-\frac{t^2}{2\sigma^2}\right)
\] (8.4)

\[
p(t) = \frac{e^{-t^2}}{\sqrt{2\pi}}
\]

\[
\frac{d}{dt}p(t) = -\sqrt{\frac{2}{\pi}} \left(e^{-t^2}\right) t
\]

\[
\frac{d^2}{dt^2}p(t) = \sqrt{\frac{2}{\pi}} e^{-t^2} \left(2t^2 - 1\right)
\] (8.5)

\[
\frac{d^3}{dt^3}p(t) = -2\sqrt{\frac{2}{\pi}} e^{-t^2} t \left(2t^2 - 3\right)
\]

\[
\frac{d^4}{dt^4}p(t) = 2\sqrt{\frac{2}{\pi}} e^{-t^2} \left(4t^4 - 12t^2 + 3\right)
\]

- **Prolate Spheroidal Wave Functions** The Prolate Spheroidal Wave Functions (PSWF) were initially in the 1960’s at Bell Labs \[155\]. They are the solution of a nontrivial optimization problem to find a waveform in that is both band-limited signal and time-limited to a certain extent. However, for practical scenarios, a more feasible challenge lies in determining the maximum level of energy
concentration of a signal with finite bandwidth \[156\]. In other words, let \(\alpha^2(T)\) and \(\beta^2(BW)\) (in equation 8.6 and 8.7) be the measure of energy concentration of the signal \(r(t)\) within a time interval \(T_p\) and a bandwidth \(BW\), respectively, with \(R(f) = \mathcal{F}\{r(t)\}\). If equation 8.6 and 8.7 hold, it is important to determine how large can \(\alpha^2(T)\) become for a band-limited signal \(r(t)\). Prolate spheroidal functions, \(\psi_n(t)\), are considered the answer to this problem. This particular set of waveforms have the following properties: (i) \(\psi_n(t)\) has even and odd symmetry with \(n\). (ii) \(\psi_n(t) \sim k_n \frac{\sin ct}{t}\), when \(t \to \infty\), (iii) \(\int_{-1}^{1} e^{j2\pi st}\psi_n(t)t = \alpha_n\psi_n(2\pi s/c)\), i.e., its Fourier transform has the same shape, (iv) Bandwidth is \(c/2\pi\), (v) The pulses are doubly orthogonal (orthogonal in two different intervals), (vi) Bandwidth and pulse width can be simultaneously controlled and are constant for all orders and (vii) Zero DC component and baseband.

\[
\alpha^2(T_p) = \frac{\int_{-T_p/2}^{T_p/2} r^2(t)t}{\int_{-\infty}^{\infty} r^2(t)t}
\]

(8.6)

\[
\beta^2(BW) = \frac{\int_{-BW}^{BW} R^2(f)f}{\int_{-\infty}^{\infty} R^2(f)f}
\]

(8.7)

To represent the signal, a numerical approximation is required, as there is no existing closed form solution. Thus, this work utilizes the equations given by \[157\]. An example of one resulting output of a PSWF pulse generator can be observed in figure 8.2. The flexibility of the PSWF and this algorithm allows an easy adaptation of the pulse to meet the needs in terms of 100 to 1000 kHz of bandwidth for the Galvanic Coupling channel.

8.2.0.2 Ultra wide band modulation schemes

Utilizing either of the pulses described above, this section provides additional background information on the candidate UWB modulation schemes. Specifically, this work focuses on Bi-Phase Modulation (BPM), Pulse Shape Modulation (PSM), Pulse Position Modulation (PPM) and Soft-Spectrum Keying (SSK), a combination of BPM and PSM \[158\] - \[154\]. Let \(s(t)\) be the transmission signal, \(a_i\) the amplitude mapped to the \(i\)-th symbol, \(p(t)\) the pulse, and \(T_s\) the period of the symbol.

**Bi-Phase Modulation** BPM is a binary pulse amplitude modulation (PAM) variation and can be expressed in mathematical form by equation 8.8. This form of modulation uses UWB pulses to represent bits by altering their polarity. Bi-phase modulation (BPM) also allows for a simplistic design and it has the ability to be less prone to channel distortion \[82\]. The polarity changing functionality of BPM also removes the PSD spectral lines as changing the pulse polarity produces a
zero mean. Detection of Bi-phase modulated pulses can be achieved using either an energy detection or template matching demodulation in the receiver. In this work, we implement the latter, using the first derivative of the Gaussian pulse and a 5 µs pulse duration to evaluate its performance against other modulation schemes.

\[
s(t) = \sum_{i=-\infty}^{\infty} a_i p(t - iT_s), \quad \text{with} \quad a_i = \{1, -1\} \quad (8.8)
\]

**Pulse Shape Modulation** Pulse shape modulation (PSM) is an alternative to PAM and PPM modulations. In PSM the information data is encoded by different pulse shapes. This requires a suitable set of pulses for higher order modulations. The orthogonality of signals used in PSM can also be leveraged for applications that require multiple access protocols. The conditions of PSM are presented in equation 8.9, where \( p_{i,m}(t) \) is the \( m \)-th waveform of the \( i \)-th symbol. For an \( M \)-ary modulation scheme, \( M \) pulses are required. Table 8.1 shows the mapping of 4-PSM for a set of \( M = 4 \) orthogonal waveforms \( \{\phi_m(t)\} \).
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Figure 8.5: Example of 4-PSM transmitted signal with PSWF pulse

Figure 8.6: Example of 2-PPM transmitted signal with Gaussian pulse

\[ s(t) = \sum_{i=-\infty}^{\infty} p_{i,m} (t - iT_s) \]  \hspace{1cm} (8.9)

**Pulse Position Modulation** With pulse position modulation (PPM), outlined in equation 8.10, the information of the data bit to be transmitted is encoded by the position of the transmitted impulse with respect to a nominal position. More precisely, while bit 0 is represented by a pulse originating at the time instant 0, bit 1 is shifted in time by the amount of \( \delta_{shift} \) from 0. With PPM, \( M \)-ary modulation can be achieved although it is usually binary.

\[ s(t) = \sum_{i=-\infty}^{\infty} p(t - iT_s + \delta_{shift} T_p), \quad \text{with} \quad 0 \leq \delta_{shift} < 1 \]  \hspace{1cm} (8.10)

**Soft Spectrum Keying** Soft-Spectrum Keying (SSK) employs inner-keying and outer-keying and is based on the M-ary PSM scheme. Bi-phase modulation is applied to the inner-keying while the PSM, with its orthogonal characteristics, are used for the outer-keying. In SSK, described in equation 8.11.
CHAPTER 8. ADAPTIVE MODULATION WITH GALVANIC COUPLING ULTRA WIDEBAND

Table 8.1: 4-PSM Modulation Mapping

<table>
<thead>
<tr>
<th>Waveform $p_{i,m}(t)$</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\phi_{i1}(t)$</td>
<td>00</td>
</tr>
<tr>
<td>$\phi_{i2}(t)$</td>
<td>01</td>
</tr>
<tr>
<td>$\phi_{i3}(t)$</td>
<td>10</td>
</tr>
<tr>
<td>$\phi_{i4}(t)$</td>
<td>11</td>
</tr>
</tbody>
</table>

Figure 8.7: Block diagram of Soft Spectrum Keying modulation transmitter

Figure 8.8: Block diagram of Soft Spectrum Keying modulation receiver

every pulse is considered as a bit itself, where the information bits $\{0, 1\}$ are conveyed in the pulse polarity. Here, $a_{ij}$ is the $j$-th bit of the $i$-th symbol and $p_j(t)$ is the $j$-th orthogonal waveform. For an $M$-ary modulation scheme, $k = \log_2 M$ pulses are required, thus using the pulses more efficiently than PSM, in terms of bits per pulses. Table 8.2 shows the mapping of the pulses for 8-SSK, with
a set of $k = 3$ orthogonal waveforms $\{\phi_k(t)\}$, and Figure 8.7 and 8.8 show the structure of the transmitter and receiver, respectively.

$$s(t) = \sum_{i=-\infty}^{\infty} \sum_{j=1}^{k} a_{ij} p_j(t - iT_s)$$  \hspace{1cm} (8.11)

Table 8.2: 8-SSK Modulation Assignment

<table>
<thead>
<tr>
<th>Sign of Waveform</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\phi_1(t)$</td>
<td>$\phi_2(t)$</td>
</tr>
<tr>
<td>+</td>
<td>+</td>
</tr>
<tr>
<td>+</td>
<td>+</td>
</tr>
<tr>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>-</td>
<td>+</td>
</tr>
<tr>
<td>-</td>
<td>+</td>
</tr>
<tr>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

To better understand the importance of orthogonality in UWB communication schemes, the received signal, $r(t)$, is studied and given by equation 8.12. In this expression, $A_c$ is an attenuation factor and $n(t)$ is the additive white Gaussian noise. In order to detect a particular pulse from a set of other orthogonal pulses, the calculation of the inner product (presented in 4.1) of $\langle \phi_n(t), \phi_m(t) \rangle$ takes place. Assuming perfect synchronization, the correlation with an orthogonal pulse template
Table 8.3: This table shows a summary of the main parameters of the pulse-based modulation schemes evaluated in the simulations. The pulse width for all is $T_p = 5\, \text{s}$, which implies a maximum symbol rate of 200 ksps.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>Order $M$</th>
<th>Bits per Symbol $k$</th>
<th>Required Pulses</th>
<th>Waveform</th>
<th>Bandwidth $BW$ (kHz)</th>
<th>Max Bit Rate $R_{\text{bit,max}}$ (kbps)</th>
</tr>
</thead>
<tbody>
<tr>
<td>BPM</td>
<td>2</td>
<td>1</td>
<td>$k$</td>
<td>Gaussian</td>
<td>587</td>
<td>200</td>
</tr>
<tr>
<td>PPM</td>
<td>2</td>
<td>1</td>
<td>$k$</td>
<td>Gaussian</td>
<td>614</td>
<td>200</td>
</tr>
<tr>
<td>PSM</td>
<td>2</td>
<td>1</td>
<td>$M$</td>
<td>Gaussian</td>
<td>611</td>
<td>200</td>
</tr>
<tr>
<td>PSM</td>
<td>4</td>
<td>2</td>
<td>$M$</td>
<td>PSWF</td>
<td>795</td>
<td>400</td>
</tr>
<tr>
<td>SSK</td>
<td>4</td>
<td>2</td>
<td>$k$</td>
<td>PSWF</td>
<td>620</td>
<td>400</td>
</tr>
<tr>
<td>SSK</td>
<td>8</td>
<td>3</td>
<td>$k$</td>
<td>PSWF</td>
<td>687</td>
<td>600</td>
</tr>
<tr>
<td>SSK</td>
<td>16</td>
<td>4</td>
<td>$k$</td>
<td>PSWF</td>
<td>759</td>
<td>800</td>
</tr>
</tbody>
</table>

will be zero, and any other value (positive or negative) otherwise. Using 8-SSK as an example, let the following bits with antipodal encoding (what in this context is known as BPM inner modulation) \(\{a_1, a_2, a_3\} = \{1, -1, 1\}\) be one symbol of the transmission signal, given by equation 8.13. The symbol decision is simple and evaluates whether $\hat{a}_i(t)$ is greater or less than zero. For the correlation taking place within the first (top) branch in figure 8.8, the results are illustrated in equation 8.14. Lastly, table 8.3 illustrates the waveform type and summary of additional parameters as they relate to each UWB modulation scheme.

$$r(t) = A_c s(t) + n(t)$$ (8.12)

$$s(t) = a_1 \phi_1(t) + a_2 \phi_2(t) + a_3 \phi_3(t) = \phi_1(t) - \phi_2(t) + \phi_3(t)$$ (8.13)

$$\hat{a}_1(t) = \langle r(t), \phi_1(t) \rangle$$
$$= \langle A_c (\phi_1(t) - \phi_2(t) + \phi_3(t)) + n(t), \phi_1(t) \rangle$$
$$= \langle A_c \phi_1(t), \phi_1(t) \rangle + \langle -A_c \phi_2(t), \phi_1(t) \rangle + \langle A_c \phi_3(t), \phi_1(t) \rangle + \langle n(t), \phi_1(t) \rangle$$ (8.14)
$$= A_c + \tilde{n}(t)$$
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Figure 8.10: BER Vs Transmit Power for M=2 Order UWB Modulation

Figure 8.11: BER Vs Transmit Power for BPM and SSK UWB Modulation
8.2.1 Performance Comparison

Similar to what is presented in 5 for passband modulation, each of the above modulation schemes are simulated within the MATLAB environment, using the stored channel impulse method. Performance metrics are illustrated in figure 8.10 and 8.11 for a target BER of $10^{-4}$. To adequately compare each system, the symbol rate was fixed to a value of 200 kbps. Results indicate that the UWB-BPM system, when compared to other order 2 modulation schemes, offers an improvement in terms of power efficiency and bandwidth efficiency. In other words, it is able to achieve the same target BER while occupying less spectrum, and transmitting with less power. By comparing the transceiver diagrams of SSK and BPM, and the BER performance displayed in figure 8.11, that 2-SSK modulation is equivalent to BPM modulation. Intuitively this makes sense, if the same pulse type is considered for each modulation scheme, as one transmitter and receiver correlator pair of SSK has the same structure as the entire BPM transceiver. Additionally, from the same figure, it can be observed that the relationship between modulation order and power efficiency for multi-dimensional signaling techniques holds [146]. As the modulation order for SSK increases, the power efficiency increases, although at the expense of bandwidth. These characteristics reveal that the BPM transceiver system leveraging the benefits of a Gaussian pulse, can be used in combination with SSK to adapt for higher data rates, with only small system design changes to the transmitter and receiver.

8.2.2 Adaptive Transmission Scheme Algorithm

By combining SSK modulation and PBM modulation, the design of an Adaptable Transmission and Modulation scheme is presented. The goal of this algorithm is to reduce the error rate in the channel, while also ensuring that the biometric classification accuracy does not fall below a prescribed level of tolerance. For SSK modulation, higher order schemes of 8 and 16 will be used in conjunction with PSWF pulses for the orthogonality between pulses. Bi-phase modulation will employ Gaussian pulses for their ability to be easily generated and capability of for control of the center frequency and bandwidth by fine tuning the pulse derivative and pulse duration, respectively. Equation 8.15 illustrates the relationship between Gaussian pulse derivative, $k$, the pulse duration $T_p$ and the center frequency, $f_c$. Figure 8.12 depicts the frequency domain representation of a Gaussian pulse with the same pulse duration, but with different order derivatives. As the order is increased, the spectrum is shifted to a different center frequency, while maintaining relatively the same transmission bandwidth. This behavior is an integral piece in the logic used to determine the best ways to adapt the transmission parameters to the present channel conditions. The steps involved in this process
are represented in figure 8.13 in flow chart form. Inputs to this system are the results of channel sounding (CFR and CIR), the initial state of modulation (8 or 16-SSK) and the accuracy of the biometric authentication process.

\[ f_c = \sqrt{k} \cdot \left( \frac{1}{T_p} \cdot \sqrt{\kappa} \right) \]  

(8.15)

•Case 1 If the channel error rate becomes too high, the classification accuracy is evaluated against a pre-set tolerance. Although a significant drop in classification accuracy may not be observed and deemed as unacceptable, channel errors still exist and could potentially have an effect on the classification rate at some point during data transmission. Thus, a Hamming code of (7,4) is used as an FEC on top of the initial UWB modulation scheme (similar to 5). A coding scheme is chosen as a first step, due to its ease of implementation in software, rather than performing costly (in terms of energy) activation of hardware components for the switching of modulation schemes. Next, feedback is sent to inform the GC-Tx to adopt this FEC but keep the current modulation index, and internally at the receiver, an “Attempt Index” is incremented and tracked. This index lets the receiving node know how many iterations of adapting the channel have been attempted. The more attempts that occur, the more conservative the system becomes at transmitting data. Successful attempts to adapt the channel (no increase in error rate above the threshold) will result in the clearing/resetting of the attempt index and the decision to keep most recent modulation scheme chosen. Priority is given to the error rate, as opposed to the classification accuracy, due to that fact that classification performance alone can be attributed to other factors (training, recording, testing, etc.). However, a change in channel error rate does directly have an affect on classification accuracy (based on results from 5), whether drastic changes are noticed or not. If frames are consistently arriving in error, it will be difficult to reconstruct the template signal used for individual identification.

•Case 2 In the case that a second attempt at adapting the transmission and modulation schemes needs to occur, meaning the error rate of the channel once again rises too high, SSK modulation is forgone and the switch to BPM (at the expense of data rate) takes place. Regardless if the classification performance is categorized as too low, the system interprets a second attempt to indicate more adverse channel behavior than previously assumed. Therefore, by using the knowledge of the center frequency and bandwidth obtained from the channel frequency response, the system calculates a new pulse order for the Gaussian wave. This data will be used to inform the transmitter to augment the derivative of the current pulse to the new value that allows for a better alignment between the spectrum of the channel and the transmitted signal for an increased channel gain (i.e., better SNR at
the receiver). Once again, after feedback is initiated, the attempt index/counter is incremented and tracked. Should a third attempt be made by the receiver to improve the system performance, BPM Gaussian pulse center frequency optimization will be maintained and a (7,4) Hamming code will be employed in an attempt to add redundancy for an increase in performance and the expense of data rate.

### 8.2.3 Results and Summary

Preliminary results validate a proof of the claims introduced above. By altering the center frequency of the Gaussian pulse (by increasing/decreasing the pulse order) to match the center frequency of the observed channel, an increase in BER curve performance is obtained. Figure 8.15 and 8.14 illustrate the BER vs Transmit power and corresponding frequency response, respectively, for BPM curves with different center frequencies. The center frequency value of 600 kHz achieves the better performance (lower BER) without having to sacrifice data rate or transmit power, as the channel frequency response (for this iteration) yields its maximum gain around the same center frequency value.
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Figure 8.13: Adaptive Transmission and Modulation Algorithm Flow Chart

Figure 8.14: Channel Frequency Response for BPM with shifted and non-shifted center frequency
Figure 8.15: BER vs Transmit Power for BPM with shifted and non-shifted center frequency
Chapter 9

Conclusion

Analytical and empirical models of the tissue communication channel representing the human body are developed by using equivalent electrical circuits that matched closely with experimental testbed measurements. For the application identified in this thesis, the modulation scheme of OOK is chosen as the preferred modulation technique, and developed on a proof-of-concept testbed composed of an embedded system implementation with supporting hardware and a synthetic tissue phantom. Additionally, a feasibility study using the one class classification method validates the utility of ECG signals as a biometric in literature and obtains classification performance above 97% accuracy for a test pool of over 80 subjects. The biometric communication system functions reliably, transmitting a sample ECG signal over a 10 cm human tissue path with a transmit power of -2 dBm, while maintaining a BER of $10^{-6}$. It is also demonstrated that the GC-secure method of transmitting biometric information by measuring the level of over-the-air signal leakage from the vantage point of an adversarial receiver with satisfactory knowledge of our system design. Future progress towards a more robust end-to-end system is explored in Chapter 10 where the focus is on exploring the utility of ECG and other biological signals for the development of an end-to-end biometric authentication system, developing ways to transmit the GC-signal over longer distances, how to improve the receiver performance and adaptive modulation techniques with channel feedback. Furthermore, we will devise adaptive modulation schemes and reconfigurable front-ends that may adapt the link/physical layer operation based on changing authentication needs or wearable sensor data reporting rates.
Chapter 10

Future Work: Recommendations for System Design Improvements

10.1 Biometric Selection

The dominant source of confusion in the classification process comes from the lack of additional distinguishing features to support our correlation-based method. The dynamic threshold, in almost all cases, is set extremely high which causes a low false positive, but also degrades the true positive performance. Although results still generate a high classification accuracy and indicate strong potential to use ECG signals as a biometric, additional analysis must be undertaken to support the wide adoption of such a method. For this study in particular, an improvement in the false positive and true positive rate are necessary steps in ensuring classification stability. For future experiments, we seek to improve algorithm performance and provide experimental results that emphasize the limitations that exists in a real-time measurement scenarios. Specifically, we aim to conduct a large-scale measurement campaign where the ECG signals from multiple subjects are measured in real-time, an under ambulatory conditions where different physiological states can be observed. Additional possibilities for biometric creation involve invoking the personalized behavior of the human body channel and its changing channel impulse response change during contact based events, to establish the creation of a new biometric key when used in combination of other biological signals.
10.2 Communication System Design

Optimizing the HW and SW system for realizing GC-IBC on a small form factor platform is a key requirement for this biometric authentication system application. This thesis focused on an OOK implementation. However, such a design choice faces drawbacks in performance based on the selection of the detection threshold. Various improvements can be made to threshold selection if OOK modulation is a continued choice for future GC-IBC applications. An averaging circuit is one example of a HW solution that can yield an adaptive threshold based on the received signal strength. The OOK modulation method was also chosen for its ability to yield simplistic non-coherent transceiver designs, resulting in lower power consumption values. Additional non-coherent modulation schemes, such as Binary Frequency Shift Keying, can offer similar power performance, albeit with a slightly larger design footprint, and provide the benefit of removing the need for threshold based detection and the issues that accompany it.

10.3 Adaptive Modulation Scheme

The adoption of FEC techniques to mitigate channel degradation was the first incremental step in creating the foundation for an adaptive transmission scheme. In order to fully realize an end-to-end system, work must be conducted to estimate the channel at the receiver and send feedback of the current channel state, via a bi-directional link, to the transmitter. Most importantly, the ever changing properties of the human body must be studied in detail to quantify characteristics who’s change is shorter than the coherence bandwidth of your channel. It has been previously uncovered that increases in temperature and hydration change slowly over time, but additional states can still be identified. Knowledge of the varying states will allow the transmitter to keep track and adapt subsequent data packets with adaptable parameters such as output power, modulation scheme, code rate, etc. Different modulations allow for more bits per symbol to be sent through the channel, thus achieving a higher throughput or a better spectral efficiency. However, it must also be noted that when using certain modulation techniques (e.g., 64-QAM), better signal-to-noise ratio is needed to overcome any interference and maintain the desired BER. The use of adaptive modulation can be seen in examples where the conductivity levels of the human body may present channel gain values that resemble path loss for a shorter link distance. Therefore, higher modulation orders can be utilized for increased data rate.
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• Conference:


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