Modulation Techniques and Channel Assessment for Galvanic Coupled Intrabody Communications

A Thesis Presented

by

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To my grandfathers, Joaquín and Juan Felipe.
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Abstract of the Thesis

Modulation Techniques and Channel Assessment for Galvanic Coupled Intrabody Communications

by

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Intrabody Communications has emerged as a topic of interest for research due to its great potential to enable a new generation of healthcare devices. As a part of a whole ecosystem of biotelemetry, it remains as a “missing link” between the sensors that collect the data from within our body and the connected applications that may help us better monitor our health.

This work focuses on the physical layer of an approach known as Galvanic Coupling. This technique applies a differential alternating field directly to the biological tissue with the help of a pair of electrodes, creating a current that propagates through and across tissues and is detected by another pair of electrodes. This method offers advantages regarding energy consumption, bit rate and hardware complexity compared to other methods.

The focus is, first, on the experimental assessment of the channel, resulting in its characterization in terms of noise, path gain and frequency response. Experimentation is made with porcine tissue, which presents similar dielectric properties compared to the human tissue. This method makes it possible for us to study inner tissues that otherwise would be difficult to access. The results provide us with more accurate channel parameters for simulation and design.

Secondly, the analysis and proposal of several $M$-ary Pulse-Based Modulation (PBM) schemes is made, using the Prolate Spheroidal Wave Functions (PSWF). Their implementation and performance characteristics are evaluated, along with the commonly used Continuous Wave Modulation (CWM) schemes.

Finally, the synchronization and multiple access issues are addressed, as important components of most practical implementations. A scenario is studied with a central receiver and several single-hop satellite transmitters. Alternatives for a protocol are then proposed.
Introduction

A new frontier in the development of wireless technologies is within our physical selves. Ongoing research in Intrabody Communications (IBC) in academy and industry is making way for a new form of communication with promising applications in healthcare. This approach uses the human body itself as medium for the transmission of signals among a network of superficial or embedded devices, with the potential to enable a new generation of systems particularly well suited for biotelemetry.

Medical devices development trends for 2015 [1] show that the growth in implants and wireless communication is further boosted by an aging population, and that “the device industry will continue to lay the groundwork for a future in which there is an implant to restore an acceptable level of functioning to virtually every compromised joint and organ.” The question, still unsolved, is how to link these implants reliably and efficiently.

One key element to outline the importance of this technology is its capability to keep constant monitoring of human physiological functions, thus allowing real-time or near real-time portable systems that barely exist today. Furthermore, the breaking commercial success of smartphones and wearable devices brings on a huge computing power and connectivity that can be leveraged to create a seamless integration with the Internet and its associated services, like mobile apps, remote diagnosis and more.

Intrabody Communications is a relatively new field. The seminal work of Zimmerman [2] in 1996 demonstrated the transmission of information through the human body using the Capacitive Coupling (CC) approach, with a rather modest bit rate. In 2002 Oberle [3] introduced the concept of Galvanic Coupling (GC.) Since then, research groups around the world have made strides in various directions, including the theoretical development of suitable models for the human body channel and experimental assessment of various modulation techniques. As recent as 2012, the IEEE approved the 802.15.6 standard for Wireless Body Area Networks, for “wireless communications
in the vicinity of, or inside, a human body” [4], a step that encourages both the academy and the industry to develop new practical applications.

We follow the method of Galvanic Coupling. It is a technique that employs weak alternating electrical current generated by a pair of electrodes attached to human tissue that creates a propagating field that is detected by another pair of electrodes.

This work explores and tests the characteristics of the biological tissues as a channel and goes further to design suitable modulation schemes. The channel, namely the human tissue, is modeled both as a lossy dielectric medium and as a 2-port lumped element electrical circuit, based on previous works, particularly by Swamanithan [5]. The experimental assessment of the properties of different tissue layers—skin, fat and muscle— is made using porcine tissue, which presents very similar dielectric properties, compared to the human body. This alternative allows us to perform skin-to-skin (SS), muscle-to-muscle (MM) and cross-layer (MS, SM) measurements that are important for applications with implanted devices. Regarding digital modulation, we detail strategies and schemes for optimizing data rate and power consumption, specifically by proposing various $M$-ary Pulse-Based Modulation (PBM) schemes using the Prolate Spheroidal Wave Functions (PSWF.)

The thesis is outlined as follows: Chapter 1 reviews the basic concepts of IBC, its applications and regulations, and explains galvanic coupling and other methods. Chapter 2 deals with the problem of channel modeling, presenting a review of theoretical models and the results of experimental assessments of the tissues. Chapter 3 explores the modulation techniques available for galvanic coupling, presenting both continuous-wave and pulse-based modulation schemes and analyzing the results of simulations and experiments. Finally, Chapter 4 develops the components of a proposed multiple access communication system, before reaching the conclusions and recommendations in Chapter 5.
Chapter 1

Intrabody Communications

Intrabody Communication (IBC), also known as Human Body Communication (HBC), is a wireless data communication technique that uses the body as transmission medium for digitally encoded information. As defined by IEEE 802.15.6, it is a non-RF method that uses the Electric Field Communication (EFC) technology, in which “data transmission from one device to another is performed through the body of a user, and devices can thereby communicate without a wire or wireless technology” [4]. There are different physical means to convey the information, including ultrasound waves, as recently proposed by Santagati and Melodia [6]; electric field or capacitive coupling, originally proposed by Zimmerman [2], Hachisuka [7], and others; and waveguide with weak electric currents, also known as galvanic coupling, investigated by [3][8][9][10][5] and others, including this work, with emphasis in channel modeling, digital modulation and transceiver design.

Recent research [9] shows that IBC is a promising short-range link alternative with low transmission power below 1 mW, with achievable data rates from a few kbps to up to 10 Mbps, depending on the implementation. IBC offers advantages with respect to over-the-air radiofrequency (OTA-RF) in various important aspects, namely power consumption, tissue heating, attenuation and leakage of the signals outside the body (that raises concerns about data security.)

This chapter gives an overview of the foreseen applications for IBC, explains the main techniques currently studied in IBC—giving further details about the galvanic coupling approach—, and presents the characteristics of the system developed for this work.
CHAPTER 1. INTRABODY COMMUNICATIONS

There are important and widespread medical conditions that could be treated more effectively if constant monitoring and immediate action were readily available. This scenario could yield personalized drug administration, faster reaction to emergency situations, and other benefits for both patients and caregivers.

Biotelemetry for healthcare and fitness is the immediate and more natural field of development for Intrabody Communications [11]. As a whole, the healthcare “smart sensor” market is expected to grow sharply and reach US$ 117 billion by 2020 [12], but this emergence comes along with an entire ecosystem of concurrent technologies to make it possible, ranging from MEMS sensors and biocompatible circuitry, to wireless intrabody links, wearables and cloud applications. As part of this environment, intrabody communications is one of the enablers of an envisioned “predictive, preventive, personalized and participatory” medicine [13], a scenario in which systems biology, big data, social networks and the Internet of Things (IoT) will help revolutionize healthcare. Figure 1.1 illustrates some of the components of a connected on-body system.

This ecosystem of sensors and actuators and biomedical applications claim for a well suited short range communication protocol, capable of delivering reliable, secure, and low-power transmission among implanted and external devices. The flow of information of such system is shown in Figure 1.2.
CHAPTER 1. INTRABODY COMMUNICATIONS

Figure 1.2: The information flow in a generic automated healthcare system with implanted devices. Source: own made.

Applications in Chronic Diseases Treatment  A specific area of application with a big market and potential benefit is chronic diseases, or Non Communicable Diseases (NCD, as defined by the World Health Organization, WHO [14].) There are four categories of such diseases, namely cardiovascular diseases (e.g. heart attacks and stroke), cancers, chronic respiratory diseases (like asthma) and diabetes. They are among the most common, costly, and preventable of all health problems. In United States, half of adult population in 2012 was reported with “one or more chronic health conditions. One of four adults had two or more chronic health conditions” [15]. Their features of persistence and dependence on diet, health habits and medications make them well suited for a real-time monitoring system.

As an example let us briefly analyze High Blood Pressure (HBP) treatment nowadays and possible alternatives in the future. Most people who have HBP will need lifelong treatment [16], in order to keep blood pressure below 140/90 mmHg or less (for patients with diabetes or chronic kidney disease.) Recommendations include following a healthy lifestyle and a healthy diet, like limiting the salty food and alcoholic drinks. The appropriate sensors may warn the patient when inadequate levels of detrimental substances are in the body. As far as medicines are concerned, many blood pressure drugs “can safely help most people control their blood pressure” and implanted drug delivery devices can automatically and more efficiently take care of the daily doses, based on current metrics and preventing symptoms before they even appear. This represents an automatic “closed loop” that facilitates also remote monitoring by specialized professionals.
1.2 Existing IBC Techniques

The main technologies currently studied and applied to IBC are briefly described in the following sections. Table 1.1 summarizes important characteristics of each.

1.2.1 Radio Frequency

Short range radio frequency systems face some drawbacks for IBC, namely: rapid attenuation within the human tissue, heating, high power consumption in comparison with other methods, and the fact that it is not confined in the human body but to an area around it (making it possible to be detected by external agents.) Nevertheless, they are popular and widespread among many applications, and recently, with the rise of smartphones and wearables, they are well positioned and getting an increased share of the market, particularly Bluetooth.

Within the envisioned system of a complete networked healthcare system (as the one depicted in Figure 1.1), radiofrequency technologies play mostly the role of external communication links, between devices (wearables, smartphones) and between them and local area or cellular networks. The main technologies in WBAN are briefly described in the following paragraphs.

**Bluetooth** Is the dominant technology of wireless body area networks, boosted by the wide adoption in wearable technologies and mobile telephony, audio, consumer electronics, health and wellness, sports and fitness, automotive, and smart home [17]. With the release of Bluetooth 4.0 Smart, or Low Energy, more networked devices are expected to appear.

**ZigBee** This is a competitor mostly in the industrial and smart home fields, which claims low power and versatility. Wireless body area networks is not its strength.
CHAPTER 1. INTRABODY COMMUNICATIONS

ANT+  Specially designed for WBAN and wireless sensor networks, it offers advantages compared to Bluetooth Low Energy in the ultra low power segment, including multiple topologies with multiple channels, for example, or one sensor to multiple displays, while BLE only allows star networks. ANT+ promises smooth interoperability among manufacturers [18] but its market share is still small compared to Bluetooth in this segment.

NFC  Although nowadays it is mostly used for electronic payment systems and access control, it can be used for other applications, particularly when integrated with a smartphone, which are increasingly being equipped with this technology (one of the latest being the iPhone.)

UWB  Due to its high data rate, Ultrawideband is thought of as cable replacement and for other similar applications. To some extent, the modulation schemes and the system explained in Section 3.2 qualify as ultrawideband, given its spectral characteristics. UWB possess important ranging capabilities that could also be used in medical applications.

1.2.2 Ultrasound

Santagati, Melodia et al. [6] have proposed a new approach to IBC leveraging on the fact that the body is mostly water (65 %), and therefore they have implemented an acoustic communications system with ultrasound waves propagating through the tissue. This novel technique employs theory and instruments that have been previously studied in other fields, like piezoelectric transducers, which are further developed for underwater communications at low frequencies, indoor localization in sensor networks, and in medical ultrasonic imaging, specially. Nevertheless, for IBC it is a new alternative.

The underlying channel modeling relies on the acoustic wave propagation in a medium, described by the Helmholtz equation, as:

$$\nabla^2 P - \frac{1}{c^2} \frac{\partial^2 P}{\partial t^2} = 0$$ (1.1)

where $P(x, y, z, t)$ is the acoustic pressure scalar field (the evolution of the pressure in time and at all spatial locations), and $c$ is the acoustic wave propagation speed in the medium.

Three important aspects are noticed in this technique $a.$ it presents low propagation speed (compared to electromagnetic waves), $b.$ it shows strong multipath propagation thus requiring special attention in the signal processing, and $c.$ there is a high attenuation of power, with an exponential
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decay with respect to distance, as given by (1.2). This attenuation is in fact equivalent to the behavior of a lossy dielectric medium described in Section 2.1.1, which applies for galvanic coupling. The acoustic pressure path loss is

\[ P(d) = P_0 e^{-2\alpha d} \]

where \( P_0 = P(0) \) is the initial pressure and \( \alpha \) (Np/m) is the amplitude attenuation coefficient, a function of the carrier frequency \( f_c \) in the form \( \alpha = a f^b \), where \( a \) (Np/(m MHz)) and \( b \) are tissue attenuation parameters.

Depending on the operating frequency, distances ranging from \( \mu \)m to cm can be achieved for an acceptable attenuation tolerance. As a general rule, the higher the frequency, the smaller the emitting elements but higher the attenuation.

Santagati’s paper goes further to describe a modulation technique and multiple access control for ultrasound. The transmission is referred to as Ultrasound WideBand (UsWB) because it employs short pulses that span a wider bandwidth, not unlike the modulation schemes presented in Section 3.2. Ultrasound intrabody communication is a promising technique that can actually coexist with other systems.

1.2.3 Galvanic and Capacitive Coupling

Both Galvanic Coupling (GC) and Capacitive Coupling (CC) are related methods employing electrodes (instead of an antenna or other transceiver), a factor that facilitates their deployment. The difference between them lies in the propagation phenomena: one is the electric field between the body and the environment and the other is the body acting as a waveguide of an ionic current. In comparison, capacitive coupling achieves longer transmission distances (less attenuation) than galvanic coupling but at lower data rates and under heavier influence of external factors, that is, it is more susceptible to environmental noise. Regarding implementation, only electrode configuration determines whether it is capacitive or galvanic.

“Since IBC is not a radiation methodology, low frequency carrier (less than 1 MHz) is a possible and common selection. The advantages of using low frequency carrier, in general, can minimize the local heating, and allow one to simplify the design of the transceiver, thus reducing the overall power consumption (system clock) and the risk of eavesdropping at the expenses of data rate.” [19]
CHAPTER 1. INTRABODY COMMUNICATIONS

Figure 1.3: Diagram of the capacitive coupling on the human body. Source: [9].

**Capacitive Coupling**  Capacitive coupling uses the human body as transmission medium between two electrodes, and the signal goes through a capacitive return path. There is a high dependence on the position of the electrodes and its adherence to skin and also the environmental noise, as it is part of the system. Above 100 MHz the body might attenuate the signal more than air because of the antenna effect. Electric field coupling and galvanic coupling are used interchangeably in this work.

**Galvanic Coupling**  Galvanic Coupling (GC) is a transmission system that uses pairs of electrodes to *couple* an alternating electric signal with the tissue. This signal induces a current\(^1\) with a principal flow between the two electrodes and a secondary flow that propagates through and across the tissue layer. A second pair of electrodes is capable to pick up the difference of potential and the receiver decodes the information contained therein. This is illustrated in Figure 1.3.

A main characteristic of GC is that the propagation occurs not only in the layer where the electrodes are coupled, but also across other adjacent layers. The propagation of the electric current in the tissue is better described in Chapter 2.

1.3 System Overview

For this work, the generation of the data bits, the filtering, pulse generation, modulation and reception and decoding is performed in Matlab, whereas the actual coupling of the signal with the tissue for the experimental stage is done using the Analog Discovery by Digilent.

\(^1\)“Galvanic” relates to electric currents.
Figure 1.4: An overview of the main components of the experimental setup. On the reception side, the recording of the data is performed by the software Waveforms by Digilent and the post-processing and decoding is made by Matlab.
Chapter 2

Human Body Channel Characterization

One of the first steps towards the establishment of a new wireless communication technology is the appropriate understanding and the convenient modeling of the channel characteristics, in order to provide the tools for an effective system design and implementation.

For galvanic coupling in intrabody communications, the uniqueness of this medium stems from the multi-layered and heterogeneous tissue composition of the body, each layer with its own propagation characteristics. Also, hydration levels or body mass index make a difference on the channel parameters, and there is a heavy dependence on the spatial arrangement and position of the electrodes in the body, as it has been found analytically and experimentally in this and other works [20]. Experimentally, the channel shows additive white Gaussian noise (AWGN) behavior and presents no phase inversion or multipath components, making it possible to simplify the signal processing and allowing certain kind of phase modulation schemes.

This chapter provides an overview of the basic elements for the understanding of the modeling of the human body as a communication channel. First, the two main approaches are explained: namely the behavior of the propagation of a wave in a lossy dielectric medium and the lumped-element circuit analysis. Secondly, the experimental results of the channel characterization using porcine tissue are presented, that include the frequency response, the noise analysis and the derivation of the channel capacity. Finally, the heat transfer mechanism in biological tissue is explored.
CHAPTER 2. HUMAN BODY CHANNEL CHARACTERIZATION

2.1 Channel Models

The human body is not the common subject of modeling as communication channel. Just recently, though, new models have been proposed for ultrasound waves and electric field and galvanic coupling. The body is a complex structure, made mostly of water plus other biological material including cells, blood and electrolytes (ionized constituents of organic matter) in different concentrations depending on the section. This leads to some interesting effects like capacitance effect in galvanic coupling due to cell membranes, or multipath in ultrasound due to tissue layer boundaries.

Further simplifications of the human body have to be made with the compromise of flexibility and accuracy. For example, a common geometrical modeling of the human forearm is an array of concentric cylinder layers containing the skin, fat, muscle and bone tissues, as in Figure 2.1.

![Figure 2.1: The forearm concentric layered model includes bone at its core, muscle, fat and skin in the outer layers. Source: [21].](image)

Signal Transmission Mechanism in Biological Tissue  The galvanic coupling technique (see Section 1.2.3) modulates ionic currents over the biological tissue [19]. This ionic current is conducted via the movable charges and free dipoles in extra-cellular fluids at lower frequencies, and through intra-cellular fluids at higher frequencies, creating a capacitive effect.

This mechanism shows “no obvious local body heating” [19] and is capable of successfully conveying the signal over distances of a couple of tens of centimeters and across layers, as demonstrated in the present work.
CHAPTER 2. HUMAN BODY CHANNEL CHARACTERIZATION

Figure 2.2: How the current propagates in cells. Source [9].

Regarding the coupling to low-frequency electric fields, the International Commission on Non-Ionizing Radiation Protection (ICNIRP) specifies that

“The interaction of time-varying electric fields with the human body results in the flow of electric charges (electric current), the polarization of bound charge (formation of electric dipoles), and the reorientation of electric dipoles already present in tissue. The relative magnitudes of these different effects depend on the electrical properties of the body that is, electrical conductivity (governing the flow of electric current) and permittivity (governing the magnitude of polarization effects). Electrical conductivity and permittivity vary with the type of body tissue and also depend on the frequency of the applied field. Electric fields external to the body induce a surface charge on the body; this results in induced currents in the body, the distribution of which depends on exposure conditions, on the size and shape of the body, and on the body’s position in the field.” [22]

In this mechanism, a current density\(^1\) \(J\) is induced in the medium when an oscillatory field\(^2\) \(E\) is applied to the material [23], as

\[
J = \sigma E + j\omega \varepsilon_0 \hat{\varepsilon} E = \sigma E + j\omega \varepsilon_0 (\varepsilon' - j\varepsilon'') E
\]  \hspace{1cm} (2.1)

where \(\sigma, \omega, \varepsilon_0, \varepsilon', \varepsilon''\) are tissue parameters. When a differential signal is applied with a pair of electrodes to biological tissue, another pair of electrodes acting as receivers may detect the signal across their difference of potential. An illustration of this mechanism is shown in Figure 2.3.

\(^1\)The electric current density is given in ampere per square meter A/m\(^2\) in SI units.
\(^2\)The electric field is given in newton per coulomb (N/C) or volt per meter (V/m) in SI units.
Figure 2.3: This illustration shows the galvanic coupling propagation mechanism as both a propagating field and ionic currents between the two pairs of electrodes.

The Dielectric Properties of the Biological Tissue  The dielectric properties of biological tissue are conductivity, \( \sigma \), and permittivity, \( \epsilon \), and are always frequency dependent [23]. An extensive set of measurements for different body parts based on experimentation performed with both human and animal samples can be found in [24].

These parameters are obtained from the complex relative permittivity, \( \hat{\epsilon} \), expressed as

\[
\hat{\epsilon} = \epsilon' - j \epsilon''
\]  

(2.2)

where \( \epsilon' \) is the relative permittivity of the material and \( \epsilon'' \) the out-of-phase associated loss factor [25], so that

\[
\epsilon'' = \sigma / (\epsilon_0 \omega)
\]  

(2.3)

where \( \omega \) is the conductivity of the material, \( \epsilon_0 \) is the permittivity of free space and \( \omega \) the angular frequency of the field. The SI unit of conductivity is siemens per meter (S/m) given that \( \epsilon_0 \) is in farads per meter (F/m) and \( \omega \) in radians per second (rad/s).

Furthermore, it is possible to utilize a lumped element modeling, as the frequencies involved yield a wavelength much bigger compared to human body and its organs, therefore biological materials can be modeled with a resistance \( R_m \) (due to dissipation loss) and a capacitance \( C_m \) (due to charge holding) and its basic circuits are represented in Figure 2.4.
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![Basic circuits](image)

Figure 2.4: Basic circuits representing the equivalent impedance based on the dielectric properties of biological tissue.

2.1.1 Wave Propagation on Lossy Dielectric Medium

The human tissue can be characterized as a lossy dielectric propagation medium [26]. As such, the energy of the wave attenuates by a factor \( e^{2\alpha d} \) [27] and, specifically, the power per unit area flowing at point \( d \) is given by

\[
P(d) = P(0)e^{-2\alpha d}
\] (2.4)

where \( d \) is the linear distance between transmitter and receiver, in our case. The magnitude \( P(0) \) is the power per unit area (W/m\(^2\)) flowing at \( d = 0 \).

If both the transmitter and the receiver have the same effective area, then the gain can be computed as

\[
A_{dB}(d) = -10 \log_{10} \left( \frac{P(d)}{P(0)} \right)
\] (2.5)

\[
= 20 \log_{10}(e) \alpha d = 8.686 \alpha d
\]

Path Loss Model Fitting  Based on several measurements in the porcine tissue, we were able to determine the parameters for an exponential fit as in (2.4). This model provides a simple approximation that is suited for most system design problems. Figure 2.5 presents the set of measurements and their fitting model, that have been constrained to the range of 3 cm to 15 cm, where it shows a better adherence and is also the scope of our system.
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Table 2.1: Model parameters

<table>
<thead>
<tr>
<th>Layer</th>
<th>$P(0)$ (W/m$^2$)</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>

Figure 2.5: The measurements and the fitted exponential model for the gain on different layers.

The difference in measurements for the same distance and medium of up to 10 dB in the worst case, is a result of the great sensibility of the reception to the position of the electrodes, coupling, heterogeneity of the tissue and humidity. Our recommendation is to use this model estimates only as a reference for design, but every actual system should be able to fine tune its own parameters once put in place (and be aware that these conditions may change over time, anyway.) There is no mention in the literature of the negative exponential behavior of the channel (probably because in vivo experiments do not allow to modify distances very easily)

The parameters $P(0)$ and the attenuation coefficient $\alpha$ obtained for different layers are summarized in Table 2.1.
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2.1.2 Lumped Element Electric Circuit Model

A lumped element electric circuit model provides a good approximation of gain and frequency response of the channel, taking into account (in more or less detail) the geometric characteristics of the medium and its dielectric properties. Due to its simplicity, this is a preferred model.

One of the first proposals of a lumped circuit was made by Wegmüller in [28], where a simple 2-port, single-layered discrete body model was presented that considering longitudinal transmit impedance (from transmitter to receiver), input impedance (at the transmitter pair of electrodes), output impedance (at the receiver pair of electrodes), cross impedances (between the two pairs of electrodes) and finally the coupling impedance (between the electrodes and the tissue.) This model, though, ignores the paths across adjacent layers, that do have an effect on the overall gain and frequency response.

More recently, the work by Swaminathan et al. [5] proposed a more accurate spatial representation of the propagation for the human forearm by taking into account the flow across layers, thus providing a three dimensional multi-layered human forearm Tissue Equivalent Circuit (TEC) model, with a large set of configurable parameters in order to provide a closed-form estimate of the channel gain and frequency response, for different values of input frequency, transmitter and receiver location, and distance and separation between the electrodes. The reader is referred to [5] for a detailed explanation. For brevity, we are only including here the results of the analysis.

The proposed model is a network of impedances in a three-dimensional array with $T$ layers, four nodes in each layer (two for each transmitter and receiver), and four terminals, therefore $N = 4 \cdot T + 4$ nodes in total, connected by:

**Impedances in the same layer** $Z^X_L$ between the pair of electrodes, $Z^X_L$ between the electrodes of the same polarity in transmitter and receiver, and $Z^X_C$ between between the electrodes of opposite polarity in transmitter and receiver, where $X = \{S, F, M, B\}$ represents the layer.

**Impedances across layers** $Z^{X,Y}_T$ between electrodes in different layers, where $X, Y = \{S, F, M, B\}$ represents the combination of adjacent layers.

**Coupling impedances** $Z_{C_{\cdot}}$ between the tissue where the transitter and receiver are connected and the electrodes.

Figure 2.6 shows a detailed illustration of the configuration described above.
In order to obtain an analytic expression for the gain and phase of the system, the tissue admittance is first derived using the circuit in Figure 2.4b, that is

\[ Y = \frac{1}{Z} = \frac{1}{Z_{\text{ext}}} + \frac{1}{Z_{\text{int}}} = G_{\text{ext}} + \frac{1}{R_{\text{int}} + jX_{C_m}} = F_W \left( \sigma M_1 + \frac{1}{\sigma \kappa M_1 + j\omega \kappa M_2} \right) \] (2.6)

where \( Z \) is the total impedance, \( G_{\text{ext}} \) is the conductance of the internal branch, \( M_1 \) is the ratio of cross sectional area \( (A) \) and length of the channel \( (L) \) with respect to the direction of the impedance measurement, \( M_2 \) is the ratio of \( A \) and thickness of the tissue layer, \( F_W \in [1, 10] \) is a correction factor based on the variability of dielectric properties dependent on tissue water content, and \( \kappa = R_{\text{ext}}/R_{\text{int}} \) is the ratio of external to internal cell resistance.

To solve the system, the Kirchhoff Current Law is used, with the admittance matrix given by
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\[ M_G = \begin{pmatrix} 
\sum_{i=1}^{n} \frac{1}{z_{1i}} & -\frac{1}{z_{12}} & \cdots & -\frac{1}{z_{1n}} \\
-\frac{1}{z_{21}} & \sum_{i=1}^{n} \frac{1}{z_{2i}} & \cdots & -\frac{1}{z_{2n}} \\
\vdots & \vdots & \ddots & \vdots \\
-\frac{1}{z_{n1}} & -\frac{1}{z_{n2}} & \cdots & \sum_{i=1}^{n} \frac{1}{z_{ni}} 
\end{pmatrix} \]  \tag{2.7}

where \( Z_{nm} \) is the impedance between node \( n \) and node \( m \). If \( \mathbf{V} \) and \( \mathbf{I} \) are vectors with the voltages of the nodes of interest and the currents, respectively

\[ \mathbf{V} = \begin{pmatrix} V_1 \\ V_2 \\ \vdots \\ V_n \end{pmatrix} \quad \text{and} \quad \mathbf{I} = \begin{pmatrix} I \\ 0 \\ \vdots \\ 0 \end{pmatrix} \]  \tag{2.8}

then the system is solved as

\[ M_G \cdot \mathbf{V} = \mathbf{I} \]  \tag{2.9}

yielding

\[ |G(\omega, E_L, d; E_S, \mathbf{T})| = 20 \log \left| \frac{V_o}{V_i} \right| \]  \tag{2.10}

and

\[ \angle G(\omega, E_L, d; E_S, \mathbf{T}) = \arctan \left( \frac{\Im(V_o)}{\Re(V_i)} \right) \]  \tag{2.11}

where \( \mathbf{T} = [T_s, T_f, T_m, T_b]^T \) is the vector of tissue thicknesses for skin, fat, muscle and bone, respectively, \( V_o \) is the potential difference across the two nodes where the receiver electrodes are located and \( V_i \) is the source voltage. The flexibility of this model allows to place the transmitter and receiver in any location.

Equations (2.10) and (2.11) are handy and flexible expressions for design and understanding of the GC-IBC channel.
2.2 Experimental Assessment of the Channel Impulse and Frequency Response

It is the purpose of this work to assess experimentally the biological tissue as a communication channel for galvanic coupling. In order to do so, several sessions of measurements were performed, with porcine tissue as test medium, and the setup shown in Figure 2.7.

Experimental Setup  According to an extensive study by Gabriel [24], it is possible and commonplace to perform studies on electrical properties on animal biological tissue that are equivalent to human tissues, particularly ovine and porcine, but also rat, frog and rabbit [25]. “The differences in the dielectric properties between animal and human species are not systematic. (…) Data for samples of animal origin are not significantly different except at the low-frequency end, where the conductivity is higher for a longitudinal section.” As well as our own studies, this work employs “excised animal tissue, mostly ovine, some porcine, from freshly killed animals,” except that we use exclusively porcine tissue.

The characteristics of the experiments are the following:

- It was performed at room temperature.

- The moisture of the tissue diminishes fast, so it was kept relatively constant with water, as much as possible.
• The experiments were performed within the first three days of the excision.
• More than one tissue was used, and the dimensions varied from one to the other.
• Instead of electrodes, alligator clips were used, as it was determined before that the change has little effect in the results.
• Transmitter and receiver were placed longitudinally and transversely with respect to the muscle fiber direction, depending on the experiment.

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.

2.2.1 Channel Probing

To check if the channel is non-frequency selective the channel impulse and frequency response were studied experimentally through a channel sounding procedure.

A Note on Correlative Sounders  A white noise signal \( n(t) \) satisfies

\[
E[n(t)n^*(t - \tau)] = R_n(\tau) = N_0\delta(\tau)
\]  (2.12)

where \( R_n(\tau) \) is the autocorrelation function of the noise, and \( N_0 \) is the single-sided noise-power spectral density. Let it be applied to the input of a linear system, so that the output is

\[
w(t) = h(t) * n(t) = \int h(\zeta)n(t - \zeta) \, d\zeta
\]  (2.13)

where \( h(t) \) is the impulse response of the system. If the output \( w(t) \) is then cross-correlated with a delayed replica of the input \( n(t - T) \), the resulting coefficient is proportional to \( h(t) \), evaluated at the delay time \( T \), that is

\[
E[w(t)n^*(t - \tau)] = E \left[ \int h(\zeta)n(t - \zeta)n^*(t - \tau) \, d\zeta \right]
= \int h(\zeta)R_n(t - \zeta) \, d\zeta
= N_0 h(\tau)
\]  (2.14)
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Figure 2.8: The measured channel impulse response (CIR) for the Muscle to Muscle tissue communication scenario.

This result allows us to determine the impulse response of a linear system using white noise as input. Nevertheless, it is not possible in practical terms to generate white noise, so experimental systems must employ deterministic waveforms with a noise-like character. For this purpose maximal-length pseudo-random binary sequences (\(m\)-sequences), also known as pseudonoise (PN) sequences are commonly used and also will be adopted in this work.

The measured Channel Impulse Response (CIR) for one tissue communication scenarios (MM) can be seen in Figure 2.8. It is noticed that there is a high peak-to-off-peak ratio, providing good correlation results from the experiments. All of the CIR for the other communication scenarios obtained from the experiments show a very similar impulse response, indicating the presence of no multipath in the channel.

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 2.9, 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.
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Figure 2.9: Channel Frequency Response (CFR) for all tissue communication scenarios for \( d = 10 \) cm.

2.2.1.1 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 is a good approximation of 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 is summarized in Table 2.2.

<table>
<thead>
<tr>
<th>Medium</th>
<th>( N_0 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>MM, SM</td>
<td>(-107.0)dBm</td>
</tr>
<tr>
<td>MS, MS</td>
<td>(-105.5)dBm</td>
</tr>
</tbody>
</table>

Based on these results, the channel is considered 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 (2.15) 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.
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\[ C = BW \cdot \log_2 \left( 1 + \frac{P_{RX}}{N_0 \cdot BW} \right) \] \hspace{1cm} (2.15)

Figure 2.10 shows the results for different Tx–Rx combinations whereas Figure 2.11 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 [5].

Figure 2.10: Channel capacity estimate for different layers and distances, under the assumption of an AWGN channel.

Figure 2.11: Experimental channel capacity estimate comparison with 2-port circuit model by [5] for \( d = 10 \) cm and a center frequency of 100 kHz.
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2.3 Thermal Analysis

One of the main concerns in using any wireless communication method on or within the body is the inherent heating of the tissue due to the applied energy. Galvanic coupling has to be necessarily constrained in terms of transmission power to stay under safe conditions, as it generates significant currents in the tissue. Nevertheless, the maximum limit of 1 mW is well below the international regulation on electromagnetic fields on the human body.

Our purpose in this section is two-fold: a) evaluate the change in temperature with a transmitter coupled on skin, and b) compare the effect of the two modulation paradigms (continuous wave and pulse-based) to assess which one prevents heating the most.

The underlying assumptions are the following: a one-dimensional description of the temperature in a semi-infinite tissue layer is enough to provide an estimate of the temperature distribution and a comparison of the different power inputs (the modulation schemes.) For this analysis we follow the approach of [29].

The most popular approach to heat propagation in biological tissue is based on the study of Pennes [30]. He provided the so called “bioheat transfer equation,” a three-dimensional description of the temperature depending on physical parameters of the tissue, blood, core temperature, and surrounding temperature. The generalized one-dimensional bioheat transfer equation is given by:

\[
\rho c \frac{\partial T}{\partial t} = k \frac{\partial^2 T}{\partial x^2} + \omega_b \rho_b c_b (T_a - T) + Q_m + Q_r(x, t)
\]  

(2.16)

All variables are detailed in Table 2.3. The key difference here with respect to the standard heat transfer is that the factor \( \omega_b \rho_b c_b (T_a - T) \) takes into account the “perfusion” of the blood, that is, the effect of the blood flow on the temperature distribution, modeled as “volumetrically distributed heat sinks or sources” [29]. The blood perfusion rate term \( \omega_b \) is a frequency with units 1/s or equivalently mL/(s mL).

The derivation of the solution for this problem is developed in more detail in [29]. We are presenting here the results. The solution to equation (2.16) is given as

\[
T(x, t) = T_0(x) + W(x, t) \exp \left( -\frac{\omega_b \rho_b c_b t}{\rho c} \right)
\]  

(2.17)

Overall, equation (2.17) provides the value of the temperature at a linear distance \( x \) and at time \( t \), as a function of \( T_0(x) \), the initial temperature at \( t = 0 \), \( W(x, t) \), which is a term that groups the influence of a spatial heat source, tissue parameters and boundary conditions, and
### Table 2.3: Tissue and Blood Thermal Parameters

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Name</th>
<th>Units</th>
<th>Value in Simulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>$A_e$</td>
<td>Electrodes contact area</td>
<td>m$^2$</td>
<td>$3 \times 10^{-4}$</td>
</tr>
<tr>
<td>$c$</td>
<td>Specific heat of tissue</td>
<td>J/(kg °C)</td>
<td>4200</td>
</tr>
<tr>
<td>$c_b$</td>
<td>Specific heat of blood</td>
<td>J/(kg °C)</td>
<td>4200</td>
</tr>
<tr>
<td>$h_0$</td>
<td>Heat convection coefficient</td>
<td>W/(m$^2$ °C)</td>
<td>10</td>
</tr>
<tr>
<td>$h_f$</td>
<td>Heat convection coefficient</td>
<td>W/(m$^2$ °C)</td>
<td>100</td>
</tr>
<tr>
<td>$k$</td>
<td>Thermal conductivity of tissue</td>
<td>W/(m °C)</td>
<td>4200</td>
</tr>
<tr>
<td>$L$</td>
<td>Distance between skin surface and body core</td>
<td>m</td>
<td>$3 \times 10^{-2}$</td>
</tr>
<tr>
<td>$P_0(t)$</td>
<td>Spatial heating power flux at skin surface</td>
<td>W/m$^2$</td>
<td>Eq. (2.26)</td>
</tr>
<tr>
<td>$Q_m$</td>
<td>Metabolic rate of tissue</td>
<td>W/m$^3$</td>
<td>33 800</td>
</tr>
<tr>
<td>$Q_r(x, t)$</td>
<td>Spatial heating</td>
<td>W/m$^3$</td>
<td>Eq. (2.25)</td>
</tr>
<tr>
<td>$t, \tau$</td>
<td>Time</td>
<td>s</td>
<td>—</td>
</tr>
<tr>
<td>$T(t)$</td>
<td>Tissue temperature</td>
<td>°C</td>
<td>Eq. (2.17)</td>
</tr>
<tr>
<td>$T_a$</td>
<td>Artery temperature</td>
<td>°C</td>
<td>37</td>
</tr>
<tr>
<td>$T_c$</td>
<td>Body core temperature</td>
<td>°C</td>
<td>37</td>
</tr>
<tr>
<td>$T_f$</td>
<td>Fluid temperature</td>
<td>°C</td>
<td>25</td>
</tr>
<tr>
<td>$W(x, t)$</td>
<td>Transformed temperature</td>
<td>°C</td>
<td>Eq. (2.22)</td>
</tr>
<tr>
<td>$x, \xi$</td>
<td>Spatial coordinate</td>
<td>m</td>
<td>—</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Thermal diffusivity of tissue</td>
<td>m$^2$/s</td>
<td>$1.1905 \times 10^{-7}$</td>
</tr>
<tr>
<td>$\eta$</td>
<td>Scattering coefficient</td>
<td>1/m</td>
<td>200</td>
</tr>
<tr>
<td>$\omega_b$</td>
<td>Blood perfusion</td>
<td>1/s</td>
<td>$0.5 \times 10^{-3}$</td>
</tr>
<tr>
<td>$\rho$</td>
<td>Density of tissue</td>
<td>kg/m$^3$</td>
<td>1000</td>
</tr>
<tr>
<td>$\rho_b$</td>
<td>Density of blood</td>
<td>kg/m$^3$</td>
<td>1000</td>
</tr>
</tbody>
</table>
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\[ \exp \left( -\frac{\omega_b \rho_b c_b}{\rho c} t \right), \]

an exponential factor that vanishes to reach thermal equilibrium as \( t \to \infty \) and depends on the physical properties of tissue and blood.

As in (2.17), the higher the blood perfusion rate, the faster the temperature falls off as \( t \to \infty \) (notice that for the simulation \( \rho_b = \rho \) and \( c_b = c \) therefore they cancel out and the exponential term depends solely on \( \omega_b \).

The initial basal temperature is given by

\[
T_0(x) = T_a + \frac{Q_m}{\omega_b \rho_b c_b} + \left( T_c - T_a - \frac{Q_m}{\omega_b \rho_b c_b} \right) \cdot \frac{\sqrt{A} \cosh \left( \sqrt{A} x \right) + \frac{h_0}{k} \sinh \left( \sqrt{A} x \right)}{\sqrt{A} \cosh \left( \sqrt{A} L \right) + \frac{h_0}{k} \sinh \left( \sqrt{A} L \right)}
\]

\[ (2.18) \]

\[
\frac{dT_0(x)}{dx} = \frac{\left( T_c - T_a - \frac{Q_m}{\omega_b \rho_b c_b} \right) \cdot \left[ \frac{\sqrt{A} \sinh \left( \sqrt{A} x \right) + \sqrt{A} \frac{h_0}{k} \cosh \left( \sqrt{A} x \right)}{\sqrt{A} \cosh \left( \sqrt{A} L \right) + \frac{h_0}{k} \sinh \left( \sqrt{A} L \right)} + \frac{h_0}{k} \left( T_f - T_a - \frac{Q_m}{\omega_b \rho_b c_b} \right) \cdot \left. \frac{\sqrt{A} \cosh \left( \sqrt{A} (L - x) \right)}{\sqrt{A} \cosh \left( \sqrt{A} L \right) + \frac{h_0}{k} \sinh \left( \sqrt{A} L \right)} \right] }{\sqrt{A} \cosh \left( \sqrt{A} L \right) + \frac{h_0}{k} \sinh \left( \sqrt{A} L \right)}
\]

\[ (2.19) \]

The Green equation is used to solve the differential equation, and its expression is

\[
G_1(x, t; \xi, \tau) = \frac{2}{L} \sum_{m=1}^{\infty} e^{-\alpha \beta_m^2 (t-\tau)} \cos(\beta_m x) \cos(\beta_m \xi) H(t - \tau)
\]

\[ (2.20) \]

where,

\[
\beta_m = \frac{2m - 1}{2L} \pi, \text{ with } m = 1, 2, 3, \ldots
\]

(2.21)

The solution for the transformed temperature \( W(x, t) \) is

\[
W(x, t) = \frac{\alpha}{k} \int_0^t g_1(\tau) \, d\tau
+ \int_0^t d\tau \int_0^L G_1(x, t; \xi, \tau) \frac{Q_r(\xi, \tau)}{\rho c} \exp \left( \frac{\omega_b \rho_b c_b}{\rho c} \tau \right) \, d\xi
\]

(2.22)

in which
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\[ g_1(t) = \left[ k \frac{dT_0(x)}{dx} \right]_{x=0} + f_1(t) \exp \left( \frac{\omega t \rho_b \Theta_b}{\rho_c} \right) H(t) \]  \hspace{1cm} (2.23)

We will adopt a surface adiabatic condition and spatial heating, where the heat flux is given by

\[ q_r(x, t) = P_0(t) \exp(-\eta x) \]  \hspace{1cm} (2.24)

and the spatial heating can be obtained as

\[ Q_r(x, t) = -\frac{\partial q_r}{\partial x} = \eta P_0(t) \exp(-\eta x) \]  \hspace{1cm} (2.25)

for a power source given by

\[ P_0(t) = \frac{s^2(t)/R_{TX}}{A_c} \]  \hspace{1cm} (2.26)

In this last term is where our transmission signals reside. \( P_0(t) \) is the time-dependent heating power on skin surface.
Chapter 3

Modulation Techniques for Galvanic Coupling

Due to its novelty, there are not clearly defined modulation schemes for the physical layer of Intrabody Communications in any of its forms (galvanic or capacitive coupling, ultrasound.) Some works have focused in quadrature phase-shift keying (QPSK) [26][31], differential binary phase-shift-keying (DBPSK) [8], on-off-keying (OOK) and direct sequence spread spectrum (DSSS) [2][32], frequency-shift keying (FSK) [33], continuous phase frequency shift keying (CPFSK) [3], and pulse position modulation (PPM) [34][6].

For the present work we will evaluate six different modulation schemes, divided in two categories: Continuous Wave Modulation (CWM) schemes, as explained in Section 3.1 and Pulse-Based Modulation (PBM) schemes, detailed in Section 3.2. The purpose is to compare them as two different paradigms to assess the performance of each one and examine their advantages and disadvantages.

In this chapter, first a description of the modulation schemes is made, including the selection of the waveform for PBM.

A Note On Orthogonality  A set of base functions of an $N$-dimensional space is called orthogonal if the following relation holds:

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

(3.1)
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This operation is known as *inner product* of two functions and is usually denoted as $\langle \phi_n(t), \phi_m(t) \rangle$ [35]. Also, if

$$K_n = \int_{t_1}^{t_2} |\phi_n(t)|^2 \, dt = 1 \text{ for all } n$$ (3.2)

then they are called *orthonormal*. This concept, although simple in appearance, is a fundamental tool for modulation and demodulation of signals, particularly in correlation-based coherent reception for pulse-based modulation.

### 3.1 Continuous Wave Modulation

*Carrier*, or *continuous* wave, modulation is a bidimensional modulation with two orthogonal base functions (*modulators*) given by (3.3),

$$\phi_1(t) = \sqrt{\frac{2}{T}} \cos(2\pi f_c t)$$
$$\phi_2(t) = \sqrt{\frac{2}{T}} \sin(2\pi f_c t)$$ (3.3)

where $\sqrt{\frac{2}{T}}$ is a normalization factor and $T$ is the integration period. According to the inner product operation given in (3.2) they are orthonormal, so that

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

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* [36] and is written as

$$s(t) = s_I \cos(2\pi f_c t) + s_Q \sin(2\pi f_c t)$$ (3.4)

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, therefore

$$s_i(t) = s_{i1} g(t) \cos(2\pi f_c t) + s_{i2} g(t) \sin(2\pi f_c t)$$ (3.5)
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![Diagram of an IQ modulator](image)

Figure 3.1: IQ modulator diagram, showing the optional (but desirable or required) transmission filter.

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, a raised cosine filter is applied.

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 3.1 and the receiver in Figure 3.2.

**Quadrature Amplitude Modulation and Phase Shift Keying** 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, with corresponding constellation diagrams shown in Figures 3.3a and 3.3b.

**Simulation and Experimental Diagram** The components of the system that is simulated and experimentally tested is shown in Figure 3.4. Both the convolutional encoding and the filtering are optional steps.

### 3.2 Pulse-Based Modulation

Pulse-based modulation or simply pulse modulation (also known as Impulse Radio in the context of ultrawideband) consists in the transmission of very short pulses with high energy.
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Figure 3.2: IQ demodulator diagram, with the matched filter \( g^*(t) \).

Figure 3.3: Constellation diagrams with Gray coding: each adjacent symbol differs by only one bit.
concentration in the interval $T_p$, yielding very high bandwidths. The data can be conveyed in the amplitude, phase, position, or shape of the pulses, plus the combination of them for higher order schemes.

Intrabody communications use low frequencies (below or about 1 MHz), even though, the pulse modulation hereby described fits part of the definition for Ultrawideband (UWB), as given by the FCC Part 15 Rule. An ultrawideband pulse is any signal for which [37]:

$$B_f \geq 0.2$$ (3.6)

or

$$BW \geq 500 \text{ MHz}$$ (3.7)

where $B_f$ is the “fractional bandwidth,” defined as:

$$B_f = \frac{BW}{f_c} = \frac{(f_H - f_L)}{(f_H + f_L)/2}$$ (3.8)

where $f_H$ and $f_L$ are the upper and lower cutoff frequencies of the transmission band of $-10$ dB, $BW$ is the bandwidth and $f_c$ the central frequency.

Our system has a bandwidth of 900 kHz and a center frequency of 550 kHz, then using 3.8
Features  Borrowing from the analysis already performed in the context of radiofrequency ultrawideband, we list here some of the main characteristics and advantages of pulse modulation.

- An important feature is that when $N$ orthogonal pulses are used, it is possible to create an $N$-dimensional space, useful for higher order modulation schemes.

- It is carrier-less (baseband) and therefore a mixer is not required, yielding simpler and cheaper transmitters.

- No additional filtering required to fit a spectral mask (once the pulses have been “fine tuned” to the desired frequencies.)

- For some PBM schemes, unlike QAM, the amplitude of the received signal is not important therefore no adjustment is necessary.

- In a multipath environment (not in the case of IBC) it is possible to apply techniques like Rake receivers thanks to its fine time resolution.
For the same reason above, it is possible to perform ranging strategies.

New multiple access techniques have been applied, including Time-Hopping (TH).

The “hard limit” of the symbol rate of a PBM scheme is the inverse of the pulse duration, because it is impossible to “shrink it” anymore. Even though, it is possible to achieve higher bit rates if the scheme is $M$-ary. For example, for BPM and pulses of $T_p = 2\mu s$ the maximum bit rate is $R = 1/T_p = 500$ kbps, whereas for 4-PSM the maximum bit rate is twice that value.

3.2.1 Pulse-Based Modulation Schemes

There are both binary and $M$-ary modulation schemes. In this work we will introduce new schemes with promising application in intrabody communications.

The most common modulation schemes are explained below \[37][38]. 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.

**Pulse Amplitude Modulation**  PAM is described by

$$s(t) = \sum_{i=-\infty}^{\infty} a_i p(t - iT_s)$$  \hspace{1cm} (3.9)

**On-Off Keying**  OOK is the simplest PAM modulation scheme, with one bit per symbol, for which

$$s(t) = \sum_{i=-\infty}^{\infty} a_i p(t - iT_s) , \hspace{0.5cm} \text{with} \hspace{0.5cm} a_i = \{1, 0\}$$  \hspace{1cm} (3.10)

**Bi-Phase Modulation**  BPM is another binary PAM variation and can be expressed as

$$s(t) = \sum_{i=-\infty}^{\infty} a_i p(t - iT_s) , \hspace{0.5cm} \text{with} \hspace{0.5cm} a_i = \{1, -1\}$$  \hspace{1cm} (3.11)

![Figure 3.6: BPM](image)
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Pulse Position Modulation  PPM encodes the information in $\delta_{\text{shift}} T_p$, and $\delta_{\text{shift}}$ is known as “modulation index.” $M$-ary modulation can be achieved although it is usually binary. It smooths the “spectral lines,” that are a consequence of the periodic pulse transmission of other schemes.

$$s(t) = \sum_{i=-\infty}^{\infty} p\left(t - iT_s + \delta_{\text{shift}} T_p\right) , \quad \text{with} \quad 0 \leq \delta_{\text{shift}} < 1 \quad (3.12)$$

Figure 3.7: Binary PPM

Pulse Shape Modulation  PSM employs different orthogonal pulse waveforms. If

$$s(t) = \sum_{i=-\infty}^{\infty} p_{i,m}\left(t - iT_s\right) \quad (3.13)$$

then $p_{i,m}(t)$ is the $m$-th waveform of the $i$-th symbol.

Figure 3.8: Binary PSM

For an $M$-ary modulation scheme, $M$ pulses are required. Table 3.1 shows the mapping of 4-PSM for a set of $M = 4$ orthogonal waveforms $\{\phi_m(t)\}$.

Soft Spectrum Keying  In SSK, every pulse is considered as a bit itself, where the information bits $\{0, 1\}$ are conveyed in its sign or amplitude. For an $M$-ary modulation scheme, $k = \log_2 M$ pulses are required\(^1\).

\(^1\)Thus using the pulses more efficiently than PSM, in terms of bits per pulses.
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Table 3.1: 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>

$s(t) = \sum_{i=-\infty}^{\infty} \sum_{j=1}^{k} a_{ij} p_j(t - iT_s)$  \hspace{1cm} (3.14)

where $a_{ij}$ is the $j$-th bit of the $i$-th symbol and $p_j(t)$ is the $j$-th orthogonal waveform.

Figure 3.9: SSK with OOK inner modulation

Table 3.2 shows the mapping of the pulses for 8-SSK, with a set of $k = 3$ orthogonal waveforms $\{\phi_k(t)\}$, and Figure 3.10 and 3.11 show the structure of the transmitter and receiver, respectively.
Table 3.2: 8-SSK Modulation Assignment

<table>
<thead>
<tr>
<th>Sign of Waveform</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\phi_1(t)$  $\phi_2(t)$  $\phi_3(t)$</td>
<td></td>
</tr>
<tr>
<td>+  +  +</td>
<td>000</td>
</tr>
<tr>
<td>+  +  -</td>
<td>001</td>
</tr>
<tr>
<td>+  -  +</td>
<td>010</td>
</tr>
<tr>
<td>+  -  -</td>
<td>011</td>
</tr>
<tr>
<td>-  +  +</td>
<td>100</td>
</tr>
<tr>
<td>-  +  -</td>
<td>101</td>
</tr>
<tr>
<td>-  -  +</td>
<td>110</td>
</tr>
<tr>
<td>-  -  -</td>
<td>111</td>
</tr>
</tbody>
</table>

Figure 3.10: Transmitter structure of the 8-SSK modulation

Reception of PBM Signals  The received signal $r(t)$ is given by

$$r(t) = A_c s(t) + n(t)$$  \hspace{1cm} (3.15)  

where $A_c$ is an attenuation factor and $n(t)$ is white Gaussian noise. To “detect” a pulse
CHAPTER 3. MODULATION TECHNIQUES FOR GALVANIC COUPLING

from among a set of orthogonal waveforms we should perform the inner product presented in (3.1), \( \langle \phi_n(t), \phi_m(t) \rangle \). Assuming perfect synchronization, the correlation with an orthogonal pulse template will be zero, and any other value (positive or negative) otherwise.

Let us use 8-SSK as an example. Let the following bits with antipodal encoding\(^2\) \( \{a_1, a_2, a_3\} = \{1, -1, 1\} \) be one symbol of the transmission signal, as

\[
s(t) = a_1 \phi_1(t) + a_2 \phi_2(t) + a_3 \phi_3(t) = \phi_1(t) - \phi_2(t) + \phi_3(t)
\]

In the correlation in the first branch in Figure 3.11 the result is

\[
\hat{a}_1(t) = \langle r(t), \phi_1(t) \rangle \\
= \langle A_c(\phi_1(t) - \phi_2(t) + \phi_3(t)) + n(t), \phi_1(t) \rangle \\
= \langle A_c \phi_1(t), \phi_1(t) \rangle + \langle -A_c \phi_2(t), \phi_1(t) \rangle + \langle A_c \phi_3(t), \phi_1(t) \rangle + \langle n(t), \phi_1(t) \rangle \\
= A_c + \tilde{n}(t)
\]

The symbol decision is simple and just evaluates whether \( \hat{a}_i(t) \) is greater or less than zero.

---

\(^2\)What in this context is known as BPM inner modulation.
CHAPTER 3. MODULATION TECHNIQUES FOR GALVANIC COUPLING

Figure 3.12: These are the first five symbols of two pulse-based modulation schemes.

3.2.2 Waveform Selection

The selection of the waveform for PBM will be considered here due to its importance for the modulation schemes. In particular, we are interested in a bandlimited pulse with high energy concentration in a given time interval, but also it is desirable to have an orthogonal set of pulses that lead to new modulation schemes such as PSM and SSK, described above. Two waveforms are considered for this work: Gaussian derivatives and prolate spheroidal wave functions (PSWF.) Previous works [34] have used rectangular pulses, regardless of the bandwidth occupancy and the
modulation capabilities of other sets. The utilization of PSWF pulses for IBC has not been reported.

**Gaussian Derivatives**  The Gaussian function is given by

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

(3.16)

The derivatives are the following

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

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

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

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

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

These signals can be used as pulses and 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 spectrum of the signals in a desired spectral mask, there are two adjustable parameters: the order of the derivative and the pulse width. For example, it has been determined in [38] that a 1 ns pulse of the 12th derivative fits the spectral mask of the UWB in the 3.6 GHz to 10.1 GHz band, whereas in this work it is found that a 2 \(\mu\)s pulse of the first derivative is well suited for the 900 MHz available bandwidth, and it is the pulse chosen for simulation and testing.

**Prolate Spheroidal Wave Functions**  The PSWF were studied by Slepian, Landau, and Pollak in the 1960’s at Bell Labs [39]. They are the solution of an optimization problem, in particular, that “a nontrivial bandlimited signal cannot be so timelimited, however, and a very natural question is to determine how large” the energy concentration of a signal can be for a signal with a finite bandwidth [40]. In other words, let \(\alpha^2(T)\) and \(\beta^2(W)\) be the measure of energy concentration of the signal \(r(t)\) within a time interval \(T\) and a bandwidth \(W\), respectively, with \(R(f) = \mathcal{F}\{r(t)\}\), if
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\[ \alpha^2(T) = \frac{\int_{-T/2}^{T/2} r^2(t) \, dt}{\int_{-\infty}^{\infty} r^2(t) \, dt} \quad (3.17) \]

and

\[ \beta^2(W) = \frac{\int_{-W}^{W} R^2(f) \, df}{\int_{-\infty}^{\infty} R^2(f) \, df} \quad (3.18) \]

then the question is how big can \( \alpha^2(T) \) be for a band-limited signal \( r(t) \). The answer to this optimization problem is the set of prolate spheroidal functions, \( \psi_n(t) \), having the following properties:

- \( \psi_n(t) \) has even and odd symmetry with \( n \).
- \( \psi_n(t) \sim k_n \sin c t \), when \( t \to \infty \).
- \( \int_{-1}^{1} e^{j2\pi st} \psi_n(t) \, dt = \alpha_n \psi_n(2\pi s/c) \), i.e., its Fourier transform has the same shape.
- Bandwidth is \( c/2\pi \).
- The pulses are doubly orthogonal.
- Bandwidth and pulse width can be simultaneously controlled and are constant for all orders.
- Zero DC component and baseband.

There is no closed form of the signals, so it is necessary then to adopt a numerical approximation. The method used in this work is the one given by [41]. With such procedure we can obtain a pulse like the one in Figure 3.13a with a spectral mask depicted in Figure 3.13b, fitted for UWB. The flexibility of the PSWF and this algorithm allows us to easily adapt the pulse to our needs in the bandwidth of 100 kHz to 1000 kHz. In fact, the pulse width, and the lower and upper limit of the frequency band can be explicitly indicated.

3.3 Simulations and Experimental Testing

A platform for testing the modulation schemes was created in Matlab for both continuous wave and pulse-based modulation. Regarding PBM, the function pulsegen generates both Gaussian and PSWF pulses as described in Section 3.2, fitting in the allowed bandwidth. The
Figure 3.13: PSWF pulse for the UWB spectral mask.
Table 3.3: 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 $T_p = 5\, \mu s$, which implies a maximum symbol rate of 200 kbps.

<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>900</td>
</tr>
</tbody>
</table>

duration is fixed and set to 5 $\mu s$. Then pulsemmod and pulsedemod perform the modulation and demodulation of the signals. Optionally, convolutional encoding and Viterbi decoding can be enabled for both.

Figure 3.14 and 3.15 illustrate the structure of the programs. In the channel, for a simulation the signal is attenuated and noise is added according with the results of the channel characterization in Chapter 2.

A Note on the Simulations Several simulations are made, for up to $10^6$ bits and encompassing all modulation schemes. The set of simulations is made with an AWGN channel based on the analysis in Chapter 2, in which the parameters for attenuation and noise power spectral density are obtained. These results are a benchmark of the achievable bit error rates.

In the following sections, an analysis of the characteristics of each modulation is made, and the results of the simulations are compared to finally provide a recommendation for this system.

Unless otherwise noted, all simulations were made for the following parameters:

- **Medium**: skin-to-skin (SS)
- **Distance between transmitter and receiver**: 15 cm
- **Symbol rate**: 150 kbps
CHAPTER 3. MODULATION TECHNIQUES FOR GALVANIC COUPLING

Parameter Selection

Data Bits

PSK or QAM
Modulation Object

Raised Cosine
Tx Filter Object

Passband
Modulation

BER
Analysis

Received Bits

PSK or QAM
Demodulation Object

Raised Cosine
Rx Filter Object

Passband
Demodulation

Figure 3.14: Structure of the program in Matlab for CWM

Parameter Selection

Data Bits

pulsegen

pulsemob

CHANNEL
Simulation or Experiment

pulsegen

pulsedemod

Received Bits

BER
Analysis

Figure 3.15: Structure of the program in Matlab for PBM
CHAPTER 3. MODULATION TECHNIQUES FOR GALVANIC COUPLING

- CWM bandwidth: 200 kHz (approximately)
- Number of bits of simulation: $10^6$

Regarding the comparison of the two classes of modulation schemes:

**Schemes for** $M = 2$  The only binary modulation in CWM performs better than (or equal to) the PBM schemes.

![Figure 3.16: Simulations with $M = 2$.](image)

**Schemes for** $M = 4$  For $M = 4$, CWM still outperforms PBM.
CHAPTER 3. MODULATION TECHNIQUES FOR GALVANIC COUPLING

Figure 3.17: Simulations with $M = 4$.

Schemes for $M = 8$  There is an interesting change in $M = 8$, for which PBM shows better performance. This is analyzed in more detail in the Power Efficiency section below.

Figure 3.18: Simulations with $M = 8$. 

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Schemes for $M = 16$ Here again PBM presents better BER. As in $M = 8$, this performance comes at a price, especially at the receiver where more correlators are required to demodulate the signal.

Figure 3.19: Simulations with $M = 16$.

Varying Distances As expected, with increased distance, worse performance, in both cases. For $d = 5$ cm, no errors were registered.
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Figure 3.20: Two modulation examples for different distances.

Different Channels  The channel plays a heavy role in performance. Whether the signal is going through or across tissue will change the gain and thus the power at the receiver.

Figure 3.21: QPSK for different channels.
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Figure 3.22: 4-SSK for different channels.

Different Bit Rates  The higher the bit rate, the higher the bit error rate.

Figure 3.23: Different bit rates.

All Pulse-Based Schemes at Same Symbol Rate  The ranking of all PBM schemes is shown here.
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Decrease in Performance as $M$ Increases  As noted before, there is a steeper degradation of the performance of CWM as the order of the modulation increases.

Figure 3.24: All PBM schemes.

Figure 3.25: Performance with respect to $M$. 

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3.4 Criteria for Selection of the Modulation Scheme

System Complexity  The absence of a mixer and a filter in the transmitter of the pulse modulator makes here an important difference. In Chapter 4 we analyze a set of sensors with only transmission capabilities, that reduce the power consumption and hardware considerably.

Bandwidth Efficiency  In this concern, CWM performs better, as for this set of simulations the bandwidth was of 200 kHz versus the 700 kHz of average of the PBM, for the same symbol rate in all cases.

Power Efficiency  For power efficiency, there is an interesting finding: as the order of the modulation, $M$, increases, the CWM schemes see a serious degradation of its performance, whereas for PBM it is milder. This can be seen clearly in Figure 3.25.

Conclusions

- Although not completely fair comparison (as PBM has wider bandwidth), an initial assessment of the two modulation categories show that, if only hardware is considered, PBM provides better value.
- BPSK and BPM, as well as QPSK and 4-SSK\(^3\) have exactly the same performance.

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\(^3\)Notice that BPM and 2-SSK are completely equivalent.
Chapter 4

Elements of Intrabody Network Design

The practical implementation of a Galvanic Coupled Intrabody Communications (GC-IBC) network requires solving important challenges regarding synchronization and multiple access schemes. The former is critical if an impulse modulation method is applied, whereas the latter is a very natural challenge that arises in a myriad of possible applications, for example, when a population of sensors reach a single data sink.

We will focus our efforts in the description of a synchronization method and a Medium Access Control (MAC) protocol for the Pulse-Based Modulation (PBM) schemes described in Chapter 3. Similar issues have been addressed previously for Impulse Radio Ultrawideband (IR-UWB) and we adopt some of its techniques to our own needs.

Scenario of a Sensor Network In the following sections we will describe synchronization and multiple access strategies for an envisioned application with multiple satellite transmitters and a single receiver, as in Figure 4.1, and with an operation regime outlined in Table 4.1. The description is given as follows:

Satellite Transmitters They are low power, low complexity, sensor-transmitter units, possibly many and distributed over a region of the body. There is a single hop between each one and the central receiver. They have no reception capabilities in order to save energy and hardware. The sensors might be different, although there also might be some redundancy in the information they send.

Central Receiver It is in charge of the synchronization of the received signals and capable to receive data from multiple sensors simultaneously. It presents higher complexity, with transmission/re-
CHAPTER 4. ELEMENTS OF INTRABODY NETWORK DESIGN

Figure 4.1: The scenario of a network of sensors communicating to a single receiver in an uncoordinated, unidirectional star topology.

ception capabilities.

4.1 Synchronization

In a PBM regime, it is critical to have good synchronization. As in IR-UWB, “even a slight misalignment can severely degrade the system performance (. . .) especially if correlation based coherent receiver is opted” [42], as is the case in our system.

One category of methods to attain synchrony in an impulsive signal uses the data pulses themselves to detect the timing offset, \( \tau \). The most well known are: Timing with Dirty Template (TDT), Timing with Code Matching (TCM), and Timing with Energy Detection (TED.) The reader is referred to the work by Akbar [42] and references therein.

A second class of synchronization methods uses dedicated synchronization headers in the packet along with (possibly) special circuitry for the detection of the offset. The preferred strategy in this work will be to introduce a special header in every sent packet. This is a reasonable alternative because of the low data rate and bursty nature of the transmission, therefore it is not crucial to “save time” by doing synchronization “on the fly.” We will follow closely the design proposed by Di Benedetto [43], where a synchronization branch is attached to the receiver, as shown in Figure 4.4.

Synchronization Process A series of pulses with appropriate auto-correlation, noise-like properties is sent to activate the Synchronization Branch of the receiver’s front end. The packet format is given in Figure 4.2.
### Chapter 4. Elements of Intrabody Network Design

Table 4.1: Features of a proposed network for analysis.

<table>
<thead>
<tr>
<th>Central Unit</th>
<th>Sensors</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Communication</strong></td>
<td>Reception from sensors and transmission to other neighbor receivers or an external data sink.</td>
</tr>
<tr>
<td><strong>Behavior</strong></td>
<td>Continuously senses for new transmissions from the network and synchronizes with new incoming packets, but does not send any acknowledgement. Once a “connection” with a new packet is established, it can detect other incoming packets simultaneously.</td>
</tr>
<tr>
<td><strong>Regimes</strong></td>
<td>A. Continuous detection.</td>
</tr>
<tr>
<td><strong>Synchronization</strong></td>
<td>Synchronizes with every incoming packet from different transmitters.</td>
</tr>
<tr>
<td><strong>Power</strong></td>
<td>Higher demand.</td>
</tr>
<tr>
<td><strong>MAC</strong></td>
<td>The receiver knows the multiple access code of the associated devices and after synchronization it matches the code of the transmitter for receiving a packet.</td>
</tr>
</tbody>
</table>
CHAPTER 4. ELEMENTS OF INTRABODY NETWORK DESIGN

<table>
<thead>
<tr>
<th>Synchronization Header</th>
<th>Preamble</th>
<th>Payload</th>
</tr>
</thead>
</table>

Figure 4.2: Packet format.

The segment of the packet with the synchronization header from the transmitter is called here $s_{sync}^T(t)$. This signal will be defined as a bi-phase-modulated $m$-sequence of length $N_{m-seq}$. The reason to choose this signal is due to its noise-like properties, namely the adequate auto-correlation properties (see equation 2.12.) Figure 4.3 presents the auto-correlation function of an example of such signals.

![Pulse Bi-Phase Modulated PN Sequence](image1)

Figure 4.3: The auto-correlation of the synchronization header shows a high peak to side lobes ratio.

In the implementation, though, the receiver has a copy of it, $s_{sync}^R(t)$, and a cross-correlation

$\hat{R}_{\tilde{s}s}(\tau) \triangleq \tilde{s}_{sync}^T(t) \ast s_{sync}^R(t)$

is applied once the incoming packet is detected. When the two signals are “aligned,” a peak is generated and detected, so that it triggers the synchronized timing for this packet.

The peak detector of the correlation function in the Synchronization Branch will start the timing mechanism, now with a known offset timing, $\tau$.

**Identification of the Transmitter** Each transmitter follows a certain multiple access scheme to send the data while sharing the medium. For the receiver to demodulate the signal, it has to necessarily know what transmitter is processing, therefore an identification procedure has to be established before the actual transmission of the data.
CHAPTER 4. ELEMENTS OF INTRABODY NETWORK DESIGN

Figure 4.4: The structure of the receiver with special circuitry for synchronization purposes.

In [43] a three way handshake is proposed, but that procedure cannot be adopted here if we want to keep the sensors without the reception capability (thus saving in hardware.) Therefore, two proposals are made with a PHY/MAC cross-layer design philosophy.

**Identification Information in the Preamble** In this approach, after synchronization is made, \( N_{ID} \) bits of identification are transmitted in a continuous manner, that is, before the multiple access scheme is entered, thus allowing the receiver to learn the identity and adopt the correspondent sequence afterwards.

**Simultaneous Synchronization and Identification Header** Another possibility is to encode the identification information within the synchronization header, so that it has the signature of each transmitter. This way, the data decoding can start right after the synchronization, avoiding the information in the preamble as above.

In any case, [43] reports that 40 to 100 pulses might be necessary to achieve a high probability of detection. A limitation on this scheme is that during the header there is a higher chance of collision, but as long as they do not collide, the receiver is capable of listening to several transmitters. Also, available codes might be limited. If they are hardcoded in the transmitters, that could be a problem for scalability.
CHAPTER 4. ELEMENTS OF INTRABODY NETWORK DESIGN

4.2 Multiple User Scheme

Among the intended applications of galvanic coupling for intrabody communications, many scenarios will include the deployment of several devices distributed within a region of the body. These devices are interested in transmit collected data from their sensors and send it out to other units or to a central collecting device. A typically adopted method for impulse modulation is Time Hopping.

4.2.1 Time Hopping

Time Hopping (TH) is a method that allows multiple access of uncoordinated devices by subdividing the time of the frame into several time slots and assigning a sequence of positions in which the transmitters will transmit every frame. The signal of a TH system that takes into accounts a number $N_u$ of users is described by

$$s_{TH}(t) = \sum_{k=1}^{N_u} \sum_{j=-\infty}^{\infty} p_j(t - jT_s - c^k_jT_c)$$  \hspace{1cm} (4.1)

where $p_j(\cdot)$ is the pulse of the $j$-th symbol and its form depends on the modulation scheme, $c^k_jT_c$ is the time shift on the $j$-th pulse given by the time hopping code of the $k$-th user.

A multiple user scheme requires to divide the frame time $T_f$ into $N_c$ time slots, called chips, of duration $T_c$.

A problem that arises from multiple access is the interference between users, described in the following section.

4.2.2 Multi-User Interference (MUI)

If the effective bit rate of a user is given by

$$R_{bit} = \frac{1}{N_e \cdot T_f} = \frac{1}{N_a \cdot N_c \cdot T_c}$$  \hspace{1cm} (4.2)

It is possible to estimate easily an error probability under the following assumptions: the pulse width is given by $T_p$ and a frame time $T_f$. It is common assumption that for an asynchronous network, the probability that one or more pulses collide when $P_u$ packets are transmitted in the same medium by $N_u$ active users follows a Poisson process, as
CHAPTER 4. ELEMENTS OF INTRABODY NETWORK DESIGN

Pr\{Pulse Collision\} = 1 - e^{-2(P_u-1)T_i/T_f} \tag{4.3}

The number of packets on the medium varies on average according to the number of active users $N_u$, the packet length $L$, the data rate $B_{bit}$ and the packet generation rate $G$,

$$P_u = \frac{N_u \cdot L \cdot G}{R_{bit}} \tag{4.4}$$

We will assume a scenario for our application with the values in Table 4.2.

<table>
<thead>
<tr>
<th>Table 4.2: Time Hopping Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of active users $N_u$</td>
</tr>
<tr>
<td>Packet length $L$</td>
</tr>
<tr>
<td>Data rate $B_{bit}$</td>
</tr>
<tr>
<td>Packet generation rate $G$</td>
</tr>
<tr>
<td>Number of encoded bits $N_e$</td>
</tr>
</tbody>
</table>

Knowing the probability of collision, we can estimate the probability of an erroneous pulse selection. Under the assumption that a random selection is made due to the interference, any of the $M$ symbols (pulses) can be selected, therefore

$$\text{Pr}\{\text{Pulse Error}\} = \left(1 - \frac{1}{M}\right) \text{Pr}\{\text{Pulse Collision}\} \tag{4.5}$$

In this proposal, the data is encoded to overcome certain amount of errors introduced by channel or interference. If one bit is represented with $N_e$ bits and at least $\lceil N_e/2 \rceil$ are needed to correct the errors, then the probability of error is

$$\text{Pr}\{\text{Bit Error}\} = \sum_{i=\lceil N_e/2 \rceil}^{N_e} \binom{N_e}{i} \text{Pr}\{\text{Pulse Error}\}^{(i)} \cdot (1 - \text{Pr}\{\text{Pulse Error}\})^{N_e-i} \tag{4.6}$$

So that the probability of successful transmission of the whole packet is

$$\text{Pr}\{\text{Success}\} = (1 - \text{Pr}\{\text{Bit Error}\})^L \tag{4.7}$$
4.3 Operation Regimes

An application measuring the glucose levels plus other biomarkers in the digestive system can provide with a clear picture of the status of a high risk chronic patient. In order to do so, let us suppose that a “swarm” of sensor-transmitter units are deployed in the thorax and/or belly of the patient, within reach of a central, possibly external receiver.

This is a quasi-static network with certain flexibility, but most of the times there is a fixed, known number of sensors. As described in Table 4.1, data comes both at regular intervals for routine control (A), and also at increased rates when it is worth noticing. For example: if the level of glucose is stable, transmissions are made in regime A, but when any new situation happens (like a fast change or after crossing a threshold), then the rate increases and enters regime B, with predefined rates. This behavior is desirable in order to, first, save battery, and then to make available a “better resolution” of the data when an emergency occurs.

These regimes can be easily implemented with analog or digital circuitry, in the processing of the data from the sensor.

A Note on the Reliability of the System  The described configuration of the network lacks typical methods to ensure reliable transmission of data, for example, acknowledgments (ACK), request-to-send/clear-to-send (RTS/CTS) and adaptive rate or modulation. Still, this is done with the premise of keeping the complexity of the transmitter as low as possible, in order to be able to have more of them with a longer period of life. Also, it is assumed that losing some data points is not a significant problem and that some redundant sensors can be used (thus giving more resiliency to the system.)
Chapter 5

Conclusions

1. A platform was developed for the simulation of a Galvanic Coupled Intrabody Communications (GC-IBC) system, in order to test several modulation schemes and the experimental parameters of the channel.

2. The experimental data from measurements in porcine tissue was collected to determine the parameters to describe noise behavior and channel gain, showing that: the noise is additive white Gaussian and the gain decays exponentially with distance.

3. For the first time, a set of $M$-ary pulse-based modulation schemes is proposed and tested for GC-IBC. This technique is well suited as it fits the spectral mask, requires low energy and presents satisfactory performance in terms of bit error rate. Furthermore, the hardware complexity is in general lower than that of the continuous wave schemes.

4. Both Gaussian derivatives and prolate spheroidal wave functions were set to fit the available bandwidth for GC-IBC, therefore enabling the aforementioned modulation scheme.

5. The outline of a networked system with a myriad of single hop, unidirectional transmitters is presented, while two important issues are discussed: synchronization and multiple access.

5.1 Future Work

- The experimental validation of the simulation with a on-the-loop system and the appropriated interface with the biological tissue (namely electrodes and signal processing.)

- A comparison of multiple access schemes in both PBM and CWM.
CHAPTER 5. CONCLUSIONS

- A multihop network: to account for the possibility of large deployments (possibly in the entire body) with a few sinks or receivers.

- An in-depth analysis of the differences in hardware complexity of continuous wave modulation and pulse-based modulation.
Bibliography


BIBLIOGRAPHY


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