Modulated Scatterer Technique Probe for Intra–Body RF Channel Measurements

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Summary

For sensing human body parameters, several sensors (implantable or digestible) can be located inside the body. Since in-body locations are not physically accessible (or at least difficult and with some risks), wireless communication with the sensing modules is required. This in-body wireless communication is a very challenging area, mainly because of the unknown and very attenuated channel and its variation with different human subjects and the environment they reside in (temperature, atmospheric conditions, etc.). The topic of this thesis is enabling in-body channel measurements for accurate channel modeling in order to allow for precise in-body localization and efficient data transmission. A measurement probe should provide accurate data measurements, is very small in size and is very power efficient in order to meet the highly demanding requirements imposed on medical electronics devices for intra-body communications. A probe employing the MST (Modulated Scattering Technique) method has been chosen to fulfill these demands.

Measuring the channel using MST has several advantages over conventional techniques where a probe in receiver mode is employed, such as low field perturbation, simple design, small size due to simpler circuitry operation, low power consumption allowing for self-powering (energy harvesting). This makes a MST probe perfectly suited for this type of intra-body communication research since it meets all the requirement for medical equipment.

In this thesis the design and validation of a MST probe is presented. Measurements have been carried out, using the proposed device, over several frequencies of operation in order to evaluate the probe performance. After initial testing in air and further testing in several tissue environments the probe performance has been evaluated. The results of the measurements provide a good match to the theory. For the MICS (Medical Implant Communications Service) and ISM (Industrial Scientific and Medical) bands the proposed MST probe provides an efficient method for channel measurements. It is a technique suited for investigating the RF (Radio Frequency) channel for intra-body communications. However it is not without issues and there are drawbacks that need to be worked in order to improve accuracy, stability and flexibility of the system. Working out these issues will allow the move to later, more complex stages where a more precise measurement data set needs to be applied under stricter conditions and a more realistic environment.
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1 Introduction

1.1 Introduction to the field of Wireless Body Area Networks (WBAN) and Intra Body Communications (IBC)

The world is changing, it is in a constant state of flux. Even though the human race is an insignificant spec in the universe the same applies to it as well. We as a species obey the same laws that govern the rest of the cosmic order. Thus we are also changing, adapting, evolving, progressing, branching out in different directions of development and constantly trying to grow and evolve. Even more so, we are perhaps the single organism on the planet, that has as a main purpose to develop in new and grandiose ways, not only as a means to adapt and survive but as a purpose of existence itself. One might say that it has become our main task, the goal we will be forever chasing for as long as we exist. Thus with our growth as a species comes growth as a society and in the field of science as a means to achieve improvement.

This being said it is clear that technology will never stop evolving, furthermore with even greater and greater pace. Thus what once was an innovation is now a thing of the past. This applies for many technological field and we are going to take a closer look in one of them in order to give reasoning for why is the research we are about to present so interesting and innovative.

Wearable devices have been a part of our daily lives for long times. Humans have proven ingenious in inventing contraptions to improve their quality of life and the physical and mental capabilities of their bodies and minds over the years. From simple gadgets as the wristwatch, the glasses to more complex devices as personal radio and audio players all the way to nowadays smart phones and watches having computational capabilities rivaling those of a desktop computer of 5 years ago. However, one might argue that even those relatively small devices, that are able to store our entire personal information and connect us to the rest of the digital world, are at the zenith of their existence. A new trend is developing in the field of wearable devices. As advances in technology allow them to be smaller and smaller, with even higher amounts of storage, computing power, and less power consumption, the time is coming for us to think of the next step.

The idea of embedding electronics on and in the human body is not new. An even older concept is that of using on-body sensors in order to obtain important information about the human body. It has been the practice for a long time in medical engineering to use small, un-obstructive sensors for Electromyography (EMG), Electrocardiography (ECG), Pulse Oximetry readings and so on. Those sensors are now small and cheap enough to be finally integrated in not only medical grade equipment, but also in personal wearable devices. Thus the need arises for those numerous devices to communicate with each other, to connect and exchange their data in order for it to be processed in the most efficient way possible using a centralized system. So the issue of developing a Wireless Body Area Network (WBAN) in order for devices encompassing the human body to communicate with each other in an efficient manner, becomes a major task for the near future.
New methods for wireless communications over, around and in the human body will be a hot research topic in the upcoming years. The human tissue needs to be investigated as a medium for in-body communication. Up till now in the field of RF wireless communications the human tissue has only been seen as an object of obstruction, attenuating and degrading signal level. Now it needs to be seen as a possible radio channel for transmitting data. The need of deeper research in how radio waves propagate around and through human tissue and the human body as a whole is evident and crucial in designing better ways for WBAN communication. Significant work in the field has already been done [1…9], however due to the complexity of human tissue it is limited in many aspects. The scope of the research needs to be expanded including new scenarios, more complex body models, as well as different environment and the uncertainties and influence on relevant parameters they might impose.

Communication using the human body as a medium has been an interesting area of research and development in the field of Electrical Engineering [10] for a long time. The idea of using the human body as a transmission medium for Intra-Body Communication Systems (IBC) for example might not be a new one [11], [12], however with recent advancements in medical technology and the ever growing needs of wearable computing and the advancements in medical sensor and monitoring systems the needs of the field have expanded.

Furthermore with the rise of the number of sensor types and the increase of possible metrics that can be gathered, the quantity of the devices operating over, on and inside the human body has drastically increased, so the need for interconnecting all the devices in a suitable WBAN makes it even more important to investigate how human tissue affects RF wave propagation. After all, RF waves in most of the conventional telecommunication systems nowadays propagate through air. Not much is currently known about propagation in more exotic media as human tissue. It is evident that more knowledge about the human body as a propagation medium is needed, as the demand for it increases with the new, more sophisticated wearable technology coming to the market.

This open up the research field of IBC and WBAN for a whole new set of questions to be answered, goals to be met, phenomena to be investigated, demands on both hardware and software modeling to be satisfied. In order to do so, gathering of a lot of measurement data is necessary to be used as a basis upon which models are to be constructed and systems developed. That means that conventional measurement techniques need to be reviewed and refitted for the purpose of efficient measurement in this new human body medium.
1.2 Research Goal/Question

The goal of this research project is to design a method to measure RF signal propagation in an intra-body radio channel for real-time communication and localization with a mobile intra-body medical sensor. The results are to serve for development of new and improvement on existing medical systems for intra-body monitoring and medical examination. Furthermore, there are several novelities to consider, which have, to our knowledge not been investigated in detail before. The main reason for this is that nowadays more and more medical devices with functionalities that are completely new to the field are being introduced. This gives a new requirement, that new types of data need to be accumulated and transmitted wirelessly between devices in real time, forming a WBAN of devices. Furthermore, those devices are located inside the human body on places that are inaccessible. This will introduce new and relevant propagation medium parameters that affect RF wave transmission and needs to be accounted for.

Another factor to consider is the fact that unlike most of the medical devices and personal wearable computing systems nowadays are static. They are affixed to the human body (smart watches, pedometers, heart rate monitors, etc.) or embedded in it (pacemakers, oximeters, etc.). We are however investigating the broader case where the actual sensor is moving inside the human body. This introduces even more uncertainty in the channel modeling and adds to potential error and loss of data.

An example of a very innovative research project in the field of IBC is the University of Twente’s real time cancer investigation system, the Nano Pill. It is currently being researched in the BIOS group. The system is a pill sized device, which upon swallowing from the patient passes through his system through the course of the day and investigates cells for the presence of cancer. The system employs a novel method for cancer detection via nano machinery, which we are not going to discuss since it is not the topic of this thesis. What is important to note is that it should be able to relay in real time any relevant data to an outside receiver (possibly one or more medical patch receivers on the human skin). This on its own is a novelty in the field from wireless communication system perspective since IBC systems research so far has been mainly concerned with static devices. Furthermore the device is to perform its task real-time so that an app on the patient’s phone could pick it up and send relevant data to the doctor via e-mail or a short message, notifying him of a possible cancer presence. Also the system should be able to provide the doctor with accurate localization data at the time of cancer detection to aid future surgery. Thus it is of utmost importance to be able to accurately predict the wireless channel behavior in the environment of intra-body wireless networking, where one device is embedded in the human body and the other is on the outside of it. Inaccurate location information could lead to false diagnosis and endanger the health of the patient in many ways.

The Nano Pill is just one example, however it shares a lot of similarities with other intra-body communication devices, the closest of which would be the generic capsule endoscopy device. In capsule endoscopy a small capsule examines the human digestive track, however in most cases data is just stored for later use. There is no real time data transmission and localization
measurement. Traditionally such devices are equipped with 360 degree cameras that record a video which is later reviewed by a skilled medical specialist. Up till now there was no need to investigate how such a device might communicate with an outside receiver, and do it in the most efficient way due to the many constraints that come from the size of the sensor and the human body itself. However, if location and real-time communication is possible this will open new possibilities. An example is the Vector Project, which is currently being developed by a consortium of companies and universities. More information on it is available at [13].

The main goal of this thesis is to develop and test a setup which is sufficiently accurate, making it possible to measure the RF channel in the case of bidirectional communication with an intra-body device. This is to provide valuable information, upon which later on a channel model can be build. The aforementioned setups is to be used to simulate the environment for a device, which is to perform intra-body measurements, reporting its findings and location wirelessly to a station outside the body in real time and in situations where the device is moving inside the body. It is very important to be able to predict the wireless radio channel behavior as accurately as possible in order to obtain reliable location information.

Knowing the channel properties and therefore allowing for a more efficient data transmission (using Adaptive Modulation and Coding for instance), also gives a possibility for reduced power consumption. This is a critical issue for such small devices, which is even further constraint by being inside a life subject (conventional chemical power cells are not an option). Channel knowledge will aid in minimizing device sizes, making it less intrusive for the patient and allow for more efficient energy consumption and thus longer measurement cycle lifespan.

1.3 Literature

There is a significant amount of literature on the topic of RF measurements via the use of scattering probes. Some examples are [14…17], which date 50 years back. It is evident that the type of measurements in this thesis are nothing new in the field as far as the measurement method itself goes. It is, however important that the particular setup as a measurement tool for IBC is a new and interesting one. The Intra-Body communication channel has not been investigated in detail, because of its complexity and the many factors influencing any measurement.

The bulk of the research work done in the field is mostly concerning WBAN and IBC systems. It is also mainly for systems that operate outside of the human body, intra body channel modeling has mostly been done via numerical simulations and software modeling tools. Examples are [18…21].

It would appear that relatively low amount of research has been published regarding communication between devices inside the human body. Furthermore, the few publications that are available, use numerical modeling and simulation, no real measurements were performed. The reason for this is the complex nature of the human body as a RF wave propagation medium and the difficulties it presents for designing an accurate, power efficient, small and inexpensive measurement system.
1.4 Thesis overview

The rest of the thesis is structured as follows.

Chapter 2 discusses the wireless channel properties of wearable device systems, which the WBAN and IBC are an example of. Also a comparison between the well-known case for propagation in air, and the scenario of this thesis: human tissue as a medium is done. The expected influence on RF wave propagation of the human body and the challenges it presents when it comes to system design, will be discussed in the chapter. Also a detailed explanation of the electromagnetic properties of human tissue is presented as well as a comprehensive set of data about possible tissue phantom materials.

In Chapter 3 options will be presented for channel measurement techniques and their advantages and disadvantages will be compared. It also presents the choice of measurement technique and a detailed theoretical explanation.

In Chapter 4 the probe design, measurement setup specifics and the challenges those bring are presented.

Chapter 5 presents the measurement results of testing the MST probe in air.

In Chapter 6 the measurement results for testing the MST probe in a human tissue phantom will be presented and compared with the results of Chapter 5.

In Chapter 7 a summary of the work, conclusions and future work and improvements are given.
2 Wireless channel properties and the human tissue as a propagation medium

Since the main goal of our research is to find a way for accurate, efficient and flexible field measurements for IBC systems it is of utmost importance to have a clear view of what the influencing channel parameters are. Potential issues that might impose constraints over the design and measurement procedure will be identified and will lead to a number of choices to be made in order to decide on the optimum system parameters for the use case.

2.1 Wireless channel parameters for free space medium (air)

From the point of view of physical channel representation there are several parameters that are common for any medium. Even though their values might differ, they could be viewed as the basis from which every wireless system design starts and builds upon. Let’s first look at the case of free space propagation, which is one that has been investigated in detail since the dawn of wireless communication. There were two main parameters to take into account before starting the design of the system: the operational frequency band and the limitations to the transmit power. In the next two sections these two will be discussed in more detail.

2.1.1 Operational frequency band

An important parameter that influences the measurement system (or any RF system) is the frequency band of operation. It is impossible to design a system that functions in the entire radio spectrum. Each frequency band has special properties, like the attenuation as a function of the distance to the transmitter, reflections, dispersion, etc. Also in human tissues (or any medium) propagation is frequency depended.

Another concern is that the radio spectrum is already crowded with various systems transmitting and receiving over multiple bands and is also limited because of regulations. The radio spectrum in each country is monitored and controlled by authorities. Bands are strictly allocated depending on the type of service the device is performing. There are exceptions, bands that are license exempt and free use is allowed. However, those frequency bands are even more crowded and the chaotic nature of transmission leads to noise pollution of a high magnitude. The possibility for interference in such bands is even worse, since they are heavily used by others, making it a challenge to achieve highly reliable transmission/reception in those bands.

There are several option to be considered, which could prove to be usable for the scope of this thesis project. Below is a short summary of the frequency bands used for WBAN systems [22]:

- Med Radio: 401-402 MHz and 405-406 MHz for narrow band (less than 100kHz) and max. EIRP (Effective Isotropic Radiated Power) of -16dBm.
- Medical Implant Communications Service (MICS): 402-405 MHz for bandwidth less than 300kHz and max. EIRP of -16dBm. The communication is of type: Listen before talk (LBT)
• ISM (Industrial Scientific and Medical): 433.050 – 434.790 MHz, 902-928 MHz, 2.4 to 2.4835 GHz, 5.725 to 5.875 GHz, for bandwidth more than 500 kHz and max. EIRP of +30dBm.
• Wireless Medical Telemetry System (WMTS) frequencies: 608–614 MHz (bandwidth more than 1500 kHz and max. EIRP of +10.8 dBm), 1395–1400 MHz (max. EIRP of +22.2 dBm), 1427–1429.5 MHz except at the locations listed in CFR Part § 90.259(b)(4) where WMTS may operate in the 1429–1431.5 MHz band (max. EIRP of +22.2 dBm).
• UWB: 3.1 to 4.9 GHz or 6 to 10.6 GHz.

![Figure 1: Frequency bands of WBAN operation](image)

However since WBAN devices are just now becoming widespread at the commercial market, standardization for such equipment is still in process. Therefore most of the devices, including sensors and measurement probes work either in the MICS bands or in the ISM bands (its sub GHz part). Also there is no worldwide unified standard to be used, besides the ISM as can be seen from Figure 1 [23]. It is however not the best suited, from channel properties point of view, because of its high frequency of operation leading to increased RF wave attenuation. This does not apply to the sub GHz part of the ISM band, however most commercial equipment uses the 2.4 GHz spectrum, which is a license exempt band and is highly attenuated and polluted by various interferers.

In this project also the MICS and sub GHz ISM bands are used to perform measurements. This frequency range is optimal for in-body communications, since lower frequencies are not suited because of bandwidth limitations on channel capacities and higher frequencies introduce too much attenuation in the human body. Therefore it is clear that the system needs to be able to perform its operation over as large a spectrum as possible in order to function with more than one standard.

Extensive research has been carried for all of these radio bands for various scenarios employing numerous techniques [24…28]. What is interesting is that most of the research is done for wireless body networks operating exclusively outside the human body. The intra-body research is much more limited and it is mostly based on numerical simulations rather than actual measurements. Even though the behavior of RF waves has been investigated to some extent it is mainly for the purpose of transmitting around the subjects tissue (in free space), not through it as main means of communication. Mostly modeling has been used since direct measurement is problematic, since it is difficult to use a human person as a test subject.
An interesting exception is the research presented in [29], which investigate propagation at 418 MHz for a device embedded in the subjects body. However, it is clear that there is the need of an effective setup that represents the human tissue environment in a manner that is both detailed and accurate enough so that measurements could be done in order to compare with simulated data.

2.1.2 Limitations to transmitting power. Specific Absorption Rate (SAR) safety guidelines

Another important factor to consider is radiated power. In order to obtain good measurement results we need to transmit sufficient energy in order to receive a signal level at the measurement site that is distinguishable from noise and interference. That conclusion leads to use power as high as possible. However, there are again regulations to what is permissible. Since the device is to be positioned inside or in very close proximity to living human subjects the effects of irradiating the human body are also to be considered. The human body as a medium will be discussed in detail later in this chapter, however. there are several standards that propose guidelines and regulations for the maximum rating when it comes to exposure of human beings to electromagnetic fields. Those include regulations from the FCC, IEEE, and IEC. One of the most used standards is the IEEE STD C95.1 and the IEC 62209, which are, unfortunately, both incomplete and are still the scope of further development especially in the light of new wearable and medical devices coming into existence nowadays. Therefore the information on the subject is still not complete, however, there are some general, albeit outdated guidelines that mostly apply for the case of smartphone exposure of the human body.

In order to define a threshold for the power limit a metric is required, in this case the Specific Absorption Rate (SAR) has been used. It has been defined by the IEEE STD C95.1 as:

The time derivative of the incremental energy \( dW \) absorbed by (dissipated in) an incremental mass \( dm \) contained in a volume element \( dV \) of given density \( \rho \).

\[
SAR = \frac{d}{dt} \left( \frac{dW}{dm} \right) = \frac{d}{dt} \left( \frac{dW}{\rho dV} \right)
\]

The SI unit of SAR is the watt per kilogram (W/kg).

There are several ways for practical calculation of SAR:

SAR can be related to the electric field at a point by [30]:

\[
SAR = \frac{\sigma |E|^2}{\rho}
\]

where

\( \sigma \) is the conductivity of the tissue (S/m)

\( \rho \) is mass density of the tissue (kg/m³)

\( E \) is ems electric field strength in tissue (V/m)
SAR can be related to the increase in temperature at the point by [30]:

\[
SAR = \frac{c \Delta T}{\Delta t} \bigg|_{t=0}
\]

where

- \( \Delta T \) is the change in temperature (°C)
- \( \Delta t \) is the duration of exposure (s)
- \( c \) is specific heat capacity (J/kg °C)

The following values for SAR threshold are in accordance with the aforementioned standards [30], [31] and are also the values adopted by mobile equipment manufacturers (mainly cell phones):

*For controlled environments (for devices emitting RF energy under the control of an aware user):*

**General exposure**

For frequencies between 100 kHz and 6 GHz:

- \( SAR \leq 0.4 \text{ W/kg} \) as averaged over the whole-body
- \( SAR \leq 8 \text{ W/kg} \) as averaged over any 1 g of tissue (defined in a tissue volume in the shape of a cube)
- \( SAR \leq 20 \text{ W/kg} \) as averaged over any 10 g of tissue (defined in a tissue volume in the shape of a cube) for the case of the hands, wrists, feet and ankles

**Low-power devices**

For frequencies between 100 kHz and 450 MHz:

\( P_{rad} \leq 7 \text{ W} \)

where \( P_{rad} \) is the radiated power

For frequencies between 450 MHz and 1500 MHz:

\( P_{rad} \leq 7(450/f) \text{ W} \)

where \( f \) is the frequency in MHz

*For uncontrolled environments (for devices emitting RF energy without control or knowledge of the user):*

**General exposure**

For frequencies between 100 kHz and 6 GHz:

\( SAR \leq 0.08 \text{ W/kg} \) as averaged over the whole-body
\[ \text{SAR} \leq 1.6 \, \text{W/kg} \quad \text{as averaged over any 1 g of tissue (defined in a tissue volume in the shape of a cube)} \]

\[ \text{SAR} \leq 4 \, \text{W/kg} \quad \text{as averaged over any 10 g of tissue (defined in a tissue volume in the shape of a cube) for the case of the hands, wrists, feet and ankles} \]

**Low-power devices**

For frequencies between 100 kHz and 450 MHz:

\[ P_{\text{rad}} \leq 1.4 \, \text{W} \quad \text{where } P_{\text{rad}} \text{ is the radiated power} \]

For frequencies between 450 MHz and 1500 MHz:

\[ P_{\text{rad}} \leq 1.4(450/f) \, \text{W} \quad \text{where } f \text{ is the frequency in MHz} \]

The system developed in this thesis should be able to operate with radiated power values within the threshold stated above. However the recommendation has been designed for the case of cell phones, therefore this is more a rough estimate than precise values.

Even though the standards for WBAN devices are still in development and no exact values for the power limit are known, it is clear that a system is needed, that is as efficient and sensitive as possible in order to be able to operate within the power constrained environment defined for WBAN devices and still produce viable and measurable results. This will also have the benefit of the system being able to operate for a longer period, with a smaller power source. This is going to be one of the major challenges to overcome in the design process.

### 2.2 Radio wave propagation in the human body

A number of issues can be expected in case of the human body as transmission medium. RF wave propagations due to difference in human tissue and its inhomogeneous structure could lead to large variations of field strength, signal phase, etc. In the next section more detailed information on the electromagnetic properties of human tissue and how this influences radio wave propagation will be presented.

#### 2.2.1 Electromagnetic properties of human tissue

Depending on the tissue type radio, waves propagate differently through the human body.

As in any medium there are three main factors influencing propagation: the permittivity (\( \varepsilon_r \text{ relative to that of air} \)), permeability (\( \mu_r \text{ relative to that of air} \)) and conductivity (\( \sigma \)) of the medium. For a dielectric (for human tissue the conductivity is mainly influenced by the imaginary part of the permittivity) the parameters are reduced to only the permittivity. Most of the research done in the field (e.g. [32]) presents the data in the form of two parameters, \( \varepsilon_r \) and \( \sigma \), most of the times the real part of the permittivity and the imaginary part of the conductivity.

Since human tissue is a nonmagnetic medium only \( \varepsilon_r \) and \( \sigma \) are investigated.
The main factor influencing $\varepsilon_r$ and $\sigma$ in a given tissue type is its water content. Depending on how much water comprises a given type of tissue, parameters can vary significantly. Some tissue types like fat for instance are poor conductors at all frequencies and have low permittivity ($\varepsilon_r = 5.5787$, $\sigma = 0.041159$), others like muscle (containing a much higher percentage of water than fat) display order of magnitude higher values for permittivity and conductivity ($\varepsilon_r = 57.1080$, $\sigma = 0.796950$). For a human propagation environment it is much harder to predict radio wave behavior compared to a free space system. The reason is that the human body comprises of more than one tissue type, it is inhomogeneous in structure and furthermore there is a huge margin for variation between individuals.

It is important to explain why permittivity and conductivity are the main parameters that model the system.

For any given medium the group velocity (the speed with which the RF wave front propagates) obeys the following equation:

$$v_g = \frac{c}{\sqrt{\varepsilon_r \mu_r}}$$

where

- $v_g$  group velocity [m/s]
- $c$ speed of light in vacuum [m/s]
- $\varepsilon_r$  relative permittivity of medium (real part)
- $\mu_r$  relative permeability of medium (1 for nonmagnetic medium)

For air where it is assumed $\varepsilon_r \approx 1$ the RF wave propagates with the speed of light. However this is not the case for other media like human tissue for example where values of $\varepsilon_r$ around 50 are not uncommon. This will result in large phase variations.

There will also be an antenna mismatch since we still have a receiver residing on the outside of the body which in most cases is in an air volume. Also there is the reflection in the boundaries, not only between the body and the outside air volume but also between different tissues, which is probably even more severe. In addition, this is very hard to model since tissue distribution and tissue type is highly dependent from person to person, even if the human body structure is well known in general. Significant signal attenuation come from all those boundary mismatches that severely degrades the signal.

Another issue with using the human body tissue as a propagation medium is that it contains water, making it practically a weak conductor (the water in tissue is not pure water but saline and so it conducts). For a weak conductor there are losses due to heat dissipation when the wave travels through. Those depend on how conductive the medium is (how high the imaginary permittivity is). The more conductive the tissue (e.g. $\sigma = 57.1080$), the more attenuation and the smaller the penetration depth of the RF wave.
So, it is clear that permittivity/conductivity will play a huge role. Furthermore, the human body is multilayered and inhomogeneous, so it is not only a more attenuating medium, but also one harder to accurately model, when compared to air. Those are the main reasons as to why IBC is such a hard field to do channel measurements for.

In Table 1 values of conductivity, relative permittivity, losses and penetration depths are given as an example [1]. The values have only been reported for a single frequency, however the source provides data over a range of frequencies which include all the MICS and ISM bands. It is possible to do measurements in all the bands suited for WBAN using the tissue properties data, since it has already been thoroughly researched and is available.

<table>
<thead>
<tr>
<th>Tissue name</th>
<th>Frequency $f$ [MHz]</th>
<th>Conductivity $\sigma$ [S/m]</th>
<th>Relative Permittivity $\varepsilon_r$</th>
<th>Loss tangent $\tan \delta$</th>
<th>Wavelength $\lambda$ [mm]</th>
<th>Penetration depth $d$ [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Skin Dry</td>
<td>402.5</td>
<td>0.689130</td>
<td>46.7290</td>
<td>0.65861</td>
<td>103.950</td>
<td>055.198</td>
</tr>
<tr>
<td>Skin Wet</td>
<td>402.5</td>
<td>0.669850</td>
<td>49.8580</td>
<td>0.60002</td>
<td>101.360</td>
<td>058.238</td>
</tr>
<tr>
<td>Fat</td>
<td>402.5</td>
<td>0.041159</td>
<td>05.5787</td>
<td>0.32949</td>
<td>311.260</td>
<td>308.650</td>
</tr>
<tr>
<td>Muscle</td>
<td>402.5</td>
<td>0.796950</td>
<td>57.1080</td>
<td>0.62323</td>
<td>094.441</td>
<td>052.535</td>
</tr>
<tr>
<td>Esophagus</td>
<td>402.5</td>
<td>1.003800</td>
<td>67.4630</td>
<td>0.66449</td>
<td>086.449</td>
<td>045.566</td>
</tr>
<tr>
<td>Stomach</td>
<td>402.5</td>
<td>1.003800</td>
<td>67.4630</td>
<td>0.66449</td>
<td>086.449</td>
<td>045.566</td>
</tr>
<tr>
<td>Colon</td>
<td>402.5</td>
<td>0.858960</td>
<td>62.5470</td>
<td>0.61331</td>
<td>090.350</td>
<td>050.950</td>
</tr>
<tr>
<td>Small Intestine</td>
<td>402.5</td>
<td>1.903800</td>
<td>66.0720</td>
<td>1.28680</td>
<td>079.911</td>
<td>025.990</td>
</tr>
</tbody>
</table>

Table 1: Electromagnetic parameters for some human tissue types

It should be noted that ‘fat’ has a relative permittivity of 10 times less on average compared to other tissues as well as a drastically lower conductivity. This can be attributed to the fact that all the other tissue types contain much larger water percentages compared to fat. Also the ‘small intestinal tissue’ is an exception due to the fact that it has twice the conductivity of other tissues (except for fat). Those large variations are very important since they can introduce large swings in signal level depending on where in the human body the device is currently residing. Furthermore tissue types variation is not uniform over the whole human population. While we are relatively sure where the small intestine is and what its dimensions are for an average individual this is not the case for body fat. There are different types of fat (white, brown, subcutaneous, etc.), also there are zones with more fat cell than others for some individuals than others. Also body fat composition is time dependent (even though it varies over a large period of time – weeks). It is almost impossible to obtain an accurate enough description of what the fat tissue distribution would be for an average person.
2.3 Human tissue phantom properties. Measurement phantom design

It is clear that radio propagation in human tissue is nothing like that in free space. That makes channel measurements in such an environment more challenging. In order to do channel measurements a medium is required that represents human tissue from electromagnetic perspective. The other option is to use real tissue, however it tends to degrade extremely fast with time and such research is difficult to realize since it is hard to obtain viable tissue samples, since those are not generally available in electronics stores. The purpose of this research is to find a material that is suitable in a way that RF wave propagation would be sufficiently close to what it would be in real human tissue.

There are in general 3 types of such material depending on their structure. There are liquid phantoms, solid phantoms and gel phantoms which are something in between liquid and solid. Gels are made from liquids and a thickening agent in order to improve their rigidity. As one might guess liquid phantoms, together with gel phantoms are much easier to produce, much cheaper and also production time is much shorter. Solid phantoms are much more expensive since they are made from custom materials. In most cases they are made by a specialized manufacturer and are design for a very narrow and specific measurement scenario. Their fabrication and design involves a lot more time and once made their properties can’t be changed. Liquid/gel phantoms on the other hand can have their properties adjusted easily by changing material proportions.

Since the many advantages of liquid/gel phantoms when it comes to flexibility of measurement and wider application of the design we have chosen to use a liquid/gel phantom type in order to be able to adjust its electromagnetic properties dynamically. This will provide more measurement scenarios to investigate.

While making a complicated and accurate phantom, one would make a liquid/gel phantom from several materials in order to obtain precise values of the desired electromagnetic characteristics, for this research, the development of a simple and efficient channel measuring device, a simple, yet flexible setup is more suited.

While in general, solutions used in such type of research are comprised of many components this is not necessary at this moment. Normally one would make a liquid/gel phantom with many different layers to model the human body. In the research described in this thesis, only the concepts are tested and a simple solution is sufficient at this moment.

The simplest solution is using water (deionized water is preferred since it contain no trace salts) as a basis. Pure water is nonconductive therefore some saline agent is added to it. NaCl is used in the form of table salt. In this way a two component phantom is build.
2.3.1 **Salinity as a parameter**

Salinity of a liquid solution basically refers to how much salt is in it. In chemical disciplines salinity is measured in mg/L (how many mg of salt is dissolved in a liter of solution). It is also referred to as ppm (parts per million). Using a weighting scale a precise material can be made with the option to easily adjust its properties.

The next step is to see what is the influence of salinity on permittivity.

In the literature [33], [34], [35] the permittivity of a saline solution is calculated using available analytical models like the Debye model, the Cole-Cole, or the double Debye, [36]. The Debye model is chosen because of its simplicity and good results.

The complex permittivity of a saline solution can be represented in Debye form as [36]:

\[ \varepsilon = \varepsilon_\infty + \frac{\varepsilon_0 - \varepsilon_\infty}{1 - i2\pi\tau f} + \frac{i\sigma}{2\pi\varepsilon_0^* f} \]

where

- \( \varepsilon_0 \) is the static dielectric constant of the solvent
- \( \varepsilon_\infty \) is the optical dielectric constant of the solvent
- \( \tau \) is the relaxation time
- \( \varepsilon_0^* \) is the permittivity of free space (8.854x10\(^{-12}\) F/m)
- \( \sigma \) is the ionic conductivity of the dissolved salt in [mho/m]
- \( f \) is the electromagnetic frequency of measurement

The above equation can be expanded an the permittivity calculated which we have done in Appendix A, which presents the Strogyn [37] method in detail and also provides the script and results of a short Matlab simulation. It is to be noted that there are several ways to calculate the permittivity using the Debye equation. For example there are two methods that are generally used in order to account for effects due to temperature and salinity change – The Stogryn model and the Klein and Swift model. The Stogryn model is used, since it is more accurate for frequencies below 1 GHz.

In Appendix A some example values are given, showing that salinity is a major factor in the phantom permittivity value. Furthermore it is demonstrated that changes in salinity influence mainly the imaginary part of the permittivity (increasing salinity increase permittivity). There is little variation in the real part, however it is 3 degrees of magnitude smaller than compared to the change in the imaginary one and can be neglected.

We can conclude that changing the salinity of our solution will change the RF wave attenuation due to the change in energy dissipated into heat in the medium.
2.3.2 Temperature as a parameter

By looking at the Strogyn model [37] one can see that the temperature of the solution is another relevant parameter influencing the Debye permittivity. From the results it is evident that changes in temperature greatly influence the real value of the permittivity as calculated by the Strogyn model. Furthermore increasing temperature decreases the permittivity.

In conclusion, modeling the solution temperature can directly influence the speed of propagation of the RF wave in the tissue phantom medium as well as affect boundary reflection and antenna mismatch.

2.3.3 Physical properties as a parameter

The permittivity of the phantom solution and its temperature have been defined as the two parameters of the tissue phantom that are most relevant. Those are the variables to adjust in the course of this research in order to simulate the effect of propagation in different tissue types, from electromagnetic perspective.

For the purpose of familiarizing the reader with the measurement setup in even more detail we will also provide information about the physical structure of the particular phantom used.

Starting with the simplest solution, which is a liquid phantom, a container is needed to put it in. Typically containers are made of some kind of plastic (PET, PVC). For PTE $\varepsilon_r \approx 3.4$ and for frequencies above 10 kHz it can be assumed to have a constant value as seen from [38]. PVC is more or less the same with $\varepsilon_r \approx 3.4$, again taking a near constant value for the microwave frequency range [39].

Thus it is clear that the liquid has permittivity much lower than any body tissue. This means that there will be a significant reflection between the container and the tissue phantom, which as stated before will influence the measurements. Additionally, there will be another reflection between the surrounding air and the container, which will further attenuate the signal, since the antenna will not make direct contact with the container as seen in Appendix C.

As for the phantom liquid itself, a simple solution of water and NaCl is used. Water volume/weight is 200 ml as this has been found to be the optimum amount for filling the container. The temperature of the water will be varied and measured via a multi-meter with a thermo coupler attached to it. Salinity will be regulated via changing the amount of salt dissolved in the water. The salinity concentration on measurements of the salt is weighted on a digital scale with 0.1 g resolution.

The exact values for salinity and temperature will depend on the particularities of the measurements at the time of their doing.

It is important to mention that this type of phantom is very simplified. It represents only one tissue type, has homogeneous structure and is not constructed with the precision a standardized measurement setup would require. In reality the human body is inhomogeneous
and comprised of more than one tissue type giving it a kind of layered structure. It is clear that the setup is not nearly as close a representation of the human body environment as one would like it to be, however for the purpose of validating the measurement technique it is sufficient.

In future research more complex phantom will be used. Such a complex phantom will be layered, comprised of several tissue type equivalent liquids/gels. Furthermore, the exact values of the permittivity of the layers will be measured using a probe in order to obtain accurate and calibrated measurement data which is to be used for the actual channel modeling process.
3 Measurement Techniques. The best choice for human tissue environment

3.1 Probe requirements and environmental factors

For the purpose of investigating the radio wave propagation a number of measurements need to be done using the designed body tissue phantom. Thus it is necessary to have a transmitter residing in the phantom, which is to radiate at the desired frequency, making it possible to measure the signal level on the outside of the phantom.

Also most of the IBC systems use tiny capsule shaped probes, which have severe limitations with respect to wireless communication due to their size (e.g. radiated power constraints and small ineffective antennas).

3.1.1 Small size

First the intra-body device has to physically resemble the IBC device. These type of devices are very small since they are supposed to be as little intrusive as possible to the human subject. That means that the measurement probe has to have both a very small footprint and a very small volume, making it crucial to design, with making the device as small as possible, in mind.

3.1.2 Minimum field perturbation/number of electrical components

The second thing to consider is that it is important to design such a system, that the intra-body RF radiator circuitry does not affect the measured field. It is after all an electronic device made of metal components that disturb the field locally, which affects received signal on the outside. The number of electrical components used in the probe should be limited in order to perturb the field as little as possible.

3.1.3 Energy efficient operation by a way of wireless/self-powering

Since IBC devices reside in the human body it is not possible to power them in a wired way. Power has to be provided either wirelessly, via energy harvesting or via a battery. Bottom line is that the probe needs to operate in the most power efficient way, with minimal energy consumption since the aforementioned powering limitations. This coincides in a way as a requirement with the requirement of having a minimum number of components since the less electronics the devices utilizes, the less the power consumption of the system as a whole will be.

3.1.4 Fluid tight enclosure

Finally the device should be fluid tight since it will reside in a body tissue phantom liquid/gel, which for the most part is made of water. Furthermore the water will be salty, making it even better at corroding metal, thus parts of the devices made of metal which are not conducting any currents should still be insulated from the phantom liquid/gel. Even though this is more of a mechanical requirement, rather than an electric issue, it should be taken into account.
Taking all of the above requirements in mind a choice needs to be made on the type of system which is best suited for the measurement setup conditions. A more detailed look in the possible measurement techniques is provided. A choice will be presented and justification will be given to the reasoning behind it, based on weighing possible advantages/advantages of the available methods.

### 3.2 Available choices

There are many available techniques to consider when making a choice for the purpose of this research. However due to certain particularities of our setup which we have already discussed in enough detail some are less suited than others.

It is to be note that we are going to perform Near-Field (NF) measurements, in fact Reactive-Near-Field (RNF) is the region the system operates in. This is practically the part of the Near-Field region that is closest to the location of measurement.

As stated there are many possibilities as far as measurement methods for electric field go. One can use for example one of the following techniques [47]:

- Thermal RF Power Sensors
- Detector (diode) RF Power Sensors
- Receiver-based Amplitude Measurements
- Monolithic Amplitude Measurement
- Direct RF Sampling Amplitude Measurement

Those, however are all based on the same premise of measuring the desired metric at the location of investigation. Thus in a way those are all direct methods falling in the same category.

In essence the purpose of any measurement as stated above could be realized by one of two ways – by measuring the value of the desired characteristic quantity (current, voltage, field, power, etc.) either directly or indirectly (by measuring an auxiliary parameter related to the quantity of interest).

The direct method is the simplest, however not the ultimate one. For example if the area where the measurement needs to be performed is hard to access or the measurement equipment introduces significant disturbance to the measurement itself. In such cases the indirect method seems like the better option.

#### 3.2.1 Direct measurement technique – receiver mode probe

For the purpose of doing NF measurements a good example is the use of a network analyzer connected to a small receiver antenna positioned at the point of interest. Such an antenna for direct measurement is called a probe and it directly measures the value of interest and displays it on the analyzer display or stores it for post processing. The probe could be stationary or move around if a scan of an area is required. Such probes are small in general, since they should be much smaller than the RF wavelength in order to obtain a valid measurement.
A good example is presented on Fig. 2, which represents an antenna measurement setup for obtaining its NF radiation pattern [40]:

By observing the example setup two disadvantages of the direct measurement method are noticeable.

- The probe is connected via a metal conductor to the measurement equipment (analyzer), thus it locally perturbs the field. This compromises the measurement since it affects the measured field value.

- One might try to limit this negative effect by using a very thin, highly resistive wire. However such a line does not provide sufficient amplitude and phase stability to carry out accurate high frequency measurements in the microwave region we are interested in.

Those are the main concerns making the direct measurement not particularly suited for NF measurements.

![Figure 2: NF measurement setup](image)

### 3.2.2 Indirect measurement techniques – Perturbation techniques

In order to determine the field for a setup like the one from Figure 2 one could also use an indirect method. The field could be locally perturbed by the use of a small object that is to perform a controlled field disturbance. Such an object is generally referred as a scatterer.

There are however two requirements stated by Bolomey [40], which need to be fulfilled in order for the object to function as a scatterer:

- The object must be small enough, so that it does not affect the field too much. Thus it needs to affect the field, but do so in a limited and predictive manner and not make it unrecognizable.
- However there is the danger of the probe being too small, so that it still affects the field, but not sufficiently. There is an effect that comes due to the presence of the
probe, but it is too small to be detected, for example if it is below the noise floor of the measurement equipment.

Obviously the above two requirements are opposite in nature. A balance is to be found between the two in order to have a detectable signal and at the same time not perturb the field too much and yield erroneous measurements. Make the setup as sensitive as possible is important in order for the system to work by perturbing the field as little as possible.

Indirect measurement systems seem to have a number of advantages over direct measurement systems. The most intriguing aspect of such a system is the fact that perturbation of the measured field is actually a desired effect, whereas with direct measurement techniques it is something one would like to avoid. This sort of gives indirect measurement systems an edge, especially for NF measurements.

Those systems also have negatives as well, but at least it would seem worth it to further investigate what specific indirect measurement techniques are available and what advantage/disadvantages each possesses.

3.2.2.1 Types of Indirect/Perturbation Measurement Techniques

There are several types of Perturbation Measurement Techniques that the author has knowledge of. Those mainly depend on the specifics of the measurement setup and the quantity to be investigated, as well as the method by which the field is being perturbed.

3.2.2.1.1 Cavity Perturbation

This method is used for measuring the amplitude of the electric field in a resonant microwave cavity. The measurement is done by introducing a small, moving, dielectric sphere to the chamber and observing the change in the resonant frequency as affected by the sphere’s movement. According to [41] other uses include permittivity/permeability measurement of materials.

3.2.2.2 Perturbation Measurement of SAR in Phantoms

This method is a lot similar in its principles to the previously described one, however the quantity of interest here is SAR. By moving a small lossy sphere inside the desired medium phantom (like for our case a body tissue type of phantom made of liquid) one can determine the SAR for the desired medium. This is done by simply irradiating the space with an antenna and recording the changes in received signal introduced by the sphere’s movement at the antenna site.

3.2.2.3 Perturbation by a Scatterer

This again is similar to the previously described techniques as a principle, however it has a major difference in it functional realization. Instead of perturbing the field by moving a lossy object a static object is used. The aforementioned object, referred to as the probe is scattering the incident field coming from the irradiating antenna. This can be realized in a number of ways – electrically, electromechanically, using a piezoelectric effect, etc..
Such a setup is a lot simpler to construct from mechanical perspective and being physically static is a huge advantage. An example is the case where the probe needs to be embedded in a solid material, which is normally not accessible (concrete structure for example).

An additional advantage is the possibility to either measure with the same antenna that is irradiating the scatterer (analogous to a monostatic radar setup), or at another location by a separate receiver antenna (analogous to a bistatic radar setup). Having two options could be seen as the system being more flexible and adjustable depending on the particularities of the measurement environment.

It is, however, not without its disadvantages. One disadvantage is the fact that there might be field components at the receiver scattered no solely by the probe itself. There might be contributions to the total received signal at the receiver due to other surrounding objects, due to the receiver itself (improper matching of the antenna) and so on. Thus it might be hard to distinguish the desired reflected signal by the probe in such a polluted RF environments since it is already inherently low level.

From the described techniques it would seem that the case when one uses a scatterer to perturb the field has the most to offer. It has the advantage of both being easy to implement, since the probe does not have to move to scan the field. Furthermore it is particularly suited for setups where the measurement area is hard to access or probe retrieval is a problem, which is exactly the case for IBC communications. Investigation in further detail and elaborating on several possible ways to implement the system, depending on how the scatterer itself functions is done. Those different scatterer types will bring a corresponding set of benefits as well as negatives, which we shall comment on.

3.2.3 Comparison – Direct vs. Indirect Measurement Techniques

In direct measurements systems the received signal is transmitted via a physical line to the receiver, allowing for a low-loss line to be used in order to have a high signal level and low interference. In contrast, in indirect systems the connection to the receiver is wireless, leading to lower received signal level and more interference. This is a disadvantage to indirect measurement systems.

Wireless transmission of the measurement data also brings the most significant benefit of indirect measurement systems compared to direct ones. Since there are no physical connections to transmit the aforementioned signal, there is no field disruption in addition to the probe itself, which in a sense leads to high accuracy when the measured quantity is considered. It is true that highly resistive wires could be used for direct measurement systems in order to minimize perturbation, but those are bulky, bending is a problem and any movement affects the transmission parameters.

One could argue that if there was a way to boost the received signal level indirect measurement systems’ main disadvantage over indirect systems would be eliminated. Possible ways to deal with the issue are imply increasing the RF power fed into the system in order to boosts received signal level. This is not a very good solution since there are SAR limitations
that vary for different systems and are strictly monitored by the proper authorities. Another way is to use coherent phase-locked detection, which is very sensitive and can detect signals in a very low SNR environment. However such receivers are more complex, costly and require more signal post processing.

A comparison of all the advantages/disadvantages of each technique is presented in Table 2. The conclusion is that the indirect method is the better choice, especially if a way is found to mitigate its disadvantages.

<table>
<thead>
<tr>
<th>Advantages</th>
<th>Direct Measurement Technique</th>
<th>Indirect Measurement Technique</th>
</tr>
</thead>
<tbody>
<tr>
<td>Prone to local perturbation due to wiring</td>
<td>Low RF signal level makes the vulnerable to interference</td>
<td>The need of greater sensitivity requires expensive hardware</td>
</tr>
<tr>
<td>Local perturbation is not a limiting issue to measurement error</td>
<td>More precise measurement</td>
<td></td>
</tr>
</tbody>
</table>

Table 2: Direct vs. Indirect measurement technique comparison

Ways for dealing with the main issues of indirect measurement techniques – the possibility of the received signal to be overshadowed by other parasitic fields (reflections from boundary mismatches, thermal noise, variations due to mechanical instabilities, interference from other systems), have been investigated.

One possibility to improve the method is to give the scattered signal a distinguishable attribute in order for the receiver to easily separate it from the rest and be able to receive it even in a very noise environment without the need to boost RF power. This is type of technique is referred to as the Modulated Scatterer Technique (MST) which will be discussed in the next section.

### 3.3 Modulated Scatter Technique (MST) in detail

A way to boost sensitivity that is not limited would be the use the scattering. It has all the advantages of indirect measurement techniques and deals with the low received signal issue in an elegant way. It seems like the best choice to consider and a deeper look into it is presented below.

#### 3.3.1 Mechanically Modulated Scatterer - Spinning Dipole

Depending on the way the modulation is realized one can identify different types of MST. For example one could produce the desired effect by mechanically affecting the probe. Such a method either spins or vibrates the scatterer probe with a low frequency. This can easily be realized by a small motor/vibrator attached to the probe support mechanism. The reflected signal is modulated and the receiver can extract it from the rest of the received RF signal by means of a homodyne phase-locked amplifier or an I/Q demodulator.
This method however would appear impractical since it negates one of the advantages of the scattering method – the fact that the probe could remain static.

3.3.2 Electrically Modulated Scatterer

Another way to modulate the reflected signal is by means of electrically modulating the scatterer probe. This can be done in a number of ways, some of which are better than others, but the main idea is to transmit a low frequency signal to the probe (by means of highly resistive wires, optical fiber in order not to perturb the field). The aforementioned signal is to act as modulating signal to a nonlinear semiconductor device (switch, photodiode, phototransistor, etc.), which is connected to the part of the probe’s structure that acts as the antenna. An example of such a setup presented in [40] can be seen in Figure 3. The scattered signal is modulated by the low frequency signal and can easily be detected by an I/Q demodulator at the receiver site. One might argue that this is not much different than a direct measurement method. However since the signal carried to the probe is not the high frequency signal we want to measure, but a low frequency signal, low interference lines can be used for its transmission. Thus the field is not perturbed additionally by the signal lines, like it would have been have we used a direct measurement setup. In this way a measurement setup is implemented that at the same time limits unwanted field perturbation and is able to separate the low level received signal from the composite unmodulated RF wave that is reflected in the direction of the receiver site.

Figure 3: Electrically modulated scatterer setup

This type of MST is particularly suited for the case of IBC measurement since physical movement of the probe is neither necessary nor an issue when it comes to unwanted field perturbation. One disadvantage is that MST probes are not so commonly used, making it hard to obtain a complete off the shelf system. However, the system has the advantage of low complexity and the required electrical components in order to run the scatterer circuitry is minimized thanks to the principles of MST itself. For this reason it was easy to build the MST measurement probe from off the shelf components, that are cheap and generally available. This can be seen as a major advantage of such a system, since it also allows for in depth customization for the particular measurement setup.
It should be mentioned that there are two possible configuration for the MST Setup: monostatic and bistatic [40], which both have their advantages and disadvantages. For this reason they are explained in further detail and one is chosen as the optimum for the setup.

3.3.3 Monostatic Configuration for MST measurements

A basic setup with monostatic arrangement is presented on Figure 4. A generator is connected to the radiating antenna via a circulator, so no signal is reflected to the source and no signal from the source is coupled into the receiver. The probe is irradiated via a suitable antenna (horn, patch, conical, etc.) and the scattered signal is received by the same antenna used for transmitting the initial signal (in bistatic configuration mode there is an auxiliary receiver antenna). The received, reflected signal is separated from the source by the circulator and detected by the receiver. It is to be noted that in such a monostatic configuration the following relation exists between the investigated field and the measurement at the receiver site [40]:

\[ U \propto |E_p|^2 \]

where

- \( U \) voltage detected by the receiver due to field reflected by the probe in [V]
- \( E_p \) field at the probe location due to energy radiated by the antenna in [V/m]

The actual measured field at the MST probe location is proportional to the square root of the voltage at the antenna terminals:

\[ E_p \propto \sqrt{U} \]

The requires conversion of the measurement data in order to obtain the true field value.

Figure 4: Monostatic MST setup

The technique is especially suited for Near-field measurements, which is what our investigation is aimed at. The way the probe side of the system functions, how the reflected
signal is modulated and received via the probe and its specific design will be explained in
detail in the following section.

3.3.4 Bistatic Configuration for MST measurements

The bistatic configuration setups is presented on Fig. 5. In this case there are two separate
antennas at different physical locations. One antenna is connected to the RF generator as was
the case with the monostatic setup. The scattering probe is irradiated however the reflected
and modulated signal is not received by the same antenna, thus there is no need of a circulator
circuit. Instead the second antenna receives the reflected RF wave and feeds it to the input of
the receiver. For this setup the relation between the voltage at the antenna terminals and the
field is as follows [40]:

\[ U \propto E_p \]

Clearly there is a one to one relation between the voltage induced on the receiving antenna
terminals and the measured field. The field equation is as follows:

\[ E_p \propto U \]

From the two equations above it is clear that no conversion of the measured data is needed as
was the case for the monostatic setup.

![Figure 5: Bistatic MST setup](image)

3.3.5 Comparison between Monostatic and Bistatic configuration

The important differences between both setups are:

- Monostatic MST is more complex since it requires the addition of a circulator, which
  for high frequency is also bulky and expensive
- Monostatic MST can be perceived as more sensitive since the measured voltage is the
  squared field value, whereas for bistatic the relation is one to one (excluding any
  calibration constants, etc.)
- Bistatic could take more space since it requires an additional antenna but the need of an expensive and bulky circulator is also eliminated.

Upon first look it appears that Monostatic MST is the obvious choice. It is more sensitive and more compact. However since at the time of setting up the measurement setup limited equipment was available, it was more convenient to used bistatic configuration. For this reason a Bistatic MST setup, at least for the initial testing and measurement, was used. This will not influence the validity of the research done, since the principles of operation of the system can be tested and evaluated with any of the two MST setups. It is of course self-explanatory that eventually we would move to Monostatic MST configuration in the later stages. This as mentioned will provide for greater sensitivity, which in term could allow for a smaller probe. The next chapter is a detailed discussion of the MST probe implementation cycle.
4 MST probe implementation

4.1 MST Probe Mark 1

A simple probe design is utilized, inspired by [42]. It uses a minimum amount of components and a standalone power supply (battery). This minimize costs, increases system robustness as well as minimizes any field perturbation due to electrical components in the near field region.

An idealized render of the probe is presented in Figure 6. It is a representation of the optimum solution for such a MST probe when it comes to dimensions and shape. The main design attributes are as follows:

![Figure 6: Dipole antenna MST probe](image)

A dipole antenna constructed of two copper wires is the main field scattering component. The wires are soldered on a dielectric piece of substrate, which serves as the probe mechanical base as well as for mounting the electronic components which are required for the probe to function. The entire probe occupies a volume of dimensions $1 \times 2 \times 3$ cm(w x l x), including the battery positioned on the back of the substrate and the dipole wires, which are made intentionally short even though for the aforementioned frequency ranges to be investigated they are nowhere near the optimum $\lambda/2$ length. This sacrifice is necessary in order to make the probe reasonably small in order to fit inside the body/phantom and the negatives effects it carries are discussed in the upcoming chapters.

The receiver need to be able to distinguish between the field reflected from the probe and any reflection coming from other sources, for instance mismatch of the antenna or other reflecting object in the vicinity of the system. As we stated in Chapter 3 this is done via electrically modulating the RF wave impeding on the probe.

An analog switch run by a microcontroller is used to realize the modulation functionality. The most commonly used modulation devices is usually a semiconductor element, like a diode or a transistor. The switch, however allows for easier implementation since no matching is require in order for it to be run. This means that is more flexible, however less precise, but as stated we are proving principles so precision is not the main concern.
The switch is operated in such a way that it alternated between open and closed state with a frequency of 1 kHz, which is a commonly used low frequency of modulation for MST probes. The 1kHz square wave is generated by a microcontroller, which has a built in programmable timer.

The effect is that essentially the impeding RF wave experiences the probe as two different devices depending on the state of the switch. In case the wires are shorted, the probe act as a single long wire and in the case the circuit is open, there are two shorter wires. In other words, the effective dipole is changed, resulting in change of probe interaction with the incoming field. Those are perceived as the two modulation levels by the receiver detector circuitry.

The resulting reflected signal is an AM (amplitude modulated) wave, which is received at the receiving antenna site. Since the carrier frequency is known at the receiver, a simple coherent detector can be used for demodulation.

Next two variations of the design of the Mark 1 probe are described, depending on different usage of the switch.

4.1.1 MST Probe Mark 1.1

The first implementation of the MST probe is the Mark1.1 probe, with some significant design points:

- The probe uses a analog switch
- The battery has been replaced by a wired power line
- The dipole wire length has been roughly chosen
- The probe was intended to serve initial testing and validation of the proposed measurement setup. No fine tuning has been done and it was design solely for the purpose of measurements in an air medium.

4.1.2 MST Probe Mark 1.2

The second MST probe uses a Video Systems switch, with a larger bandwidth compared to the analog switch of the Mark1.1 probe. Appendix B gives a more detailed description of the components level of the MST Mark 1.2 probe.

4.2 MST Probe Mark 2

The second iteration of the MST probe is a slight modification of the first iteration.

The main difference between the Mark 1 and Mark 2 probe is the antenna tuning. The Mark2 probe has the option of regulating the effective wire length in one of two way discussed in detail in Appendix B. This allows for measurements to be performed at the desired frequency, where the antenna is tuned for $\lambda/2$. Tuning the antenna improves sensitivity significantly.
4.2.1 MST Probe Mark 2.1

The first Mark 2.1 probe uses the analog switch. Appendix B provides detailed information about any relevant design modifications.

4.2.2 MST Probe Mark 2.2

The Mark 2.2 probe utilizes the video switch to realize its scattering functionality. There are some variations to the electrical component parameters with respect to the Mark 2.1 probe. Those are due to the difference between the two switches and are detailed in Appendix B.

4.3 MST Probe Mark 3

After doing extensive testing and measurements with the Mark 2 probe sufficient date was gathered in order to move to the next stage of the measurement cycle. The Mark 3 probe is an improved version of the Mark 2 probe. It provides better sensitivity due to proper tuning via $\lambda/2$ antenna length, allowing for measurements in a liquid medium representing human tissue and has minimized size. The Mark 3 probe is the last iteration of the MST probe for IBC channel measurements. A brief description of the introduced improvements is given in this chapter. More detailed information can be found in Appendix B.

Since the probe uses antenna rods tuned for $\lambda/2$ and in addition has to be liquid tight the wire length cannot be extended dynamically as was the case with the Mark 2 probe. For this reason several probes were built with different antenna lengths for the desired frequency of measurement. Also since now there is one device per frequency band, unlike the Mark 2 case where the same device was modified for each frequency, there might be a measurement error introduced by any differences in the probe design.

4.3.1 MST Probe Mark 3.1

The Mark 3.1 probe uses the analog switch again. However it has been modified in order for it to function as a measurement device in a human body tissue phantom of the liquid variety. This results in the use of a reduced substrate size, minimizing dimensions and weight. Another improvement is the location of the antenna pad terminals on the board, which are directly connected to the switch output. Those are now positioned in a symmetrical manner, as close as possible to each other and the switch, which should result in better matching an improved modulation depth boosting sensitivity. The probe has been made water tight in order not to short out the circuitry when submerged in the conductive body tissue phantom by the use of a plastic sealing agent.

4.3.2 MST Probe Mark 3.2

This the video switch version of the Mark 3.1 probe. As with the Mark 3.1 it is smaller, has improved antenna pad positioning and is liquid tight compared to the previous implementations. There are no significant differences in constructions or principles of operation between the Mark 3.1 and Mark 3.2.
5 Testing and measurement results

Measurements were performed using a bistatic setup. As mentioned before it is less sensitive compared with a monostatic setup, but at this stage insufficient equipment was available to construct the more complex monostatic setup. A disc cone antenna fed by a sinusoidal wave generator was used to irradiate the probe and an log-periodic receiver antenna connected to a Digital Spectrum Analyzer (DSA) measured the modulated signal. A snapshot of the setups can be seen on Figure 7.

![Figure 7: Bistatic MST Testing and Measurement Setup](image)

Medium – Air

5.1 MST probe Mark 1

5.1.1 Testing and measurement (Mark 1.1)

There are no viable measurement results to be presented here.

5.1.2 Testing and measurement (Mark 1.2)

5.1.2.1 Measurement at 1 GHz

Measurements at 1GHz for the Mark 1.2 probe were performed and are presented on Figure 8. After adjustment of the 'scale an enhanced version of the measurement plot was obtained and is presented on Figure 9.
The following comments can be made about the latter graph:

- X-axis presents the sweep time since the SA operates in 0 span mode
- Y-axis presents the received power level
- The two different operation modes can be seen (low and high signal level)
- This means that the AM modulated wave is detected and the probe is indeed scattering the impeding 1GHz signal and reflecting a modulated 1 kHz version of it.
- There is, however, a lot of noise distorting both the shape and level of the received square wave demodulated pulse

The conclusion is that the system works as expected. The MST probe works in scattering mode and the receiver is able to detect the reflected signal, even though noise distortion is significant.

Figure 8: MST probe Mark 1.2 measurement at 1GHz
5.1.3 Results, Discussion and possible improvement strategies

A short summary of the measurement results obtained with the Mark 1 probe is presented in order to discuss the main issues the design has and to propose an efficient strategy to solve them for the Mark 2 design.

First of all it is to be noted that the probe is not efficient enough. Many of the measurements done produced no detectable signal. It is clear that the signal level is too low to be separated from the noise floor by the receiver. The main reason is that the antenna length was not tuned for $\lambda/2$, leading to an ineffective radiator at the frequency of interest.

The probe, however, proved to function as expected. Even though a single successful measurement was performed it is enough to validate probe operation. The device indeed reflects the impeding RF wave, modulating it with a frequency of 1KHz, which is detected at the receiver. This easily proven by taking a look at Figure 9. One could also measure directly at the output of the timer for both Mark 1.1 and Mark 1.2 via a Digital Signal Oscilloscope (DSO) to verify probe operation (see Appendix B).

It was proven that the design is functional, the probe operates in the MST mode, however it is far too inefficient a system yet to produce viable channel measurement data. Improvements need to be made in order to have a detectable signal.

Difference in input resistance of the two switches, is why only the Mark 2.1 probe worked.

There are two approaches for improving the system efficiency. Either the probe has to be made more efficient or the receiver has to be made more efficient.
For this thesis we have chosen to work on the first approach. This will allow for efficient measurements under a number of different setups and make the system more flexible. Also it will reduce costs since in order to have a good I/Q demodulator which boosts receiver sensitivity custom components are required. The measurement setup in this thesis uses a general type DSA, found in any RF laboratory.

For the above reason the main focus when designing the Mark 2 probe was to improve the antenna efficiency, and make it possible for it to be tuned for \( \lambda/2 \) electrical length for each of the respective frequency bands.

The method we have used for tuning the antenna length is presented in Appendix B.

5.2 MST probe Mark 2

Since for the Mark 1.1 probe the only viable results were at 1 GHz, it was logical to start the next measurement cycle at the aforementioned frequency in order to compare the performance of the two probes. Measurements result at 400MHz and 2 GHz are also available.

5.2.1 Testing and measurement (Mark 2.1)

5.2.1.1 Measurement at 1 GHz

The very first measurement with the Mark 2 probe was done with its iteration utilizing the design employing the coils pair for antenna tuning. It was expected for the received signal to be detectable since now the antenna length should have been electrically tuned for \( \lambda/2 \) via the coils. There was not distinguishable signal present however.

Next measurements were performed with the probe with the physical extension wires attached. As expected there was quite the improvement in the signal level and it could be clearly distinguished this time, although the level is pretty low still. A plot of the measurement can be seen on Figure 10.

The measurement was repeated and even further improvement can be seen on Figure 11. The peak to peak signal level is now 1dBm, as compared to that on Figure 10, which was lower than 0.1 dBm.
5.2.1.2 Measurement at 400 MHz

Much like for the previous measurement frequency, wire extensions were used, so that the dipole antenna is a total of 37.5 cm in length ($\lambda/2$ length at 400 MHz), effectively tuning the antenna for 400 MHz. There were no other changes in the setup as compared to the previous one. However this was only done after attempting a measurement with the probe using the matching coil network. As expected there were no viable results.
The results produced by the tuned probe are presented on Figure 12. Those show even greater improvement than what one would expect based on probe performance at 1 GHz. An improvement in the signal peak to peak ratio is noticeable, which is now 2 dB (twice as much as for the 1 GHz case). The probe allows for a measurement with a greater peak to peak ratio, for 400 MHz than for 1 GHz, since the path loss effectively decreases for the lower frequency.

Figure 12: MST probe Mark 2.1 (tuned) measurement at 400 MHz

5.2.1.3 Measurement at 2 GHz

After validating the probe operation at 400 MHz and 1 GHz, a measurement was performed at 2 GHz. Although this is a frequency range outside the scope of most WBAN and ICB devices it is still important for the purpose of this thesis. Measurement at such a high frequency is crucial to validate probe operation. Also it is very close to the 2.4 GHz ISM band, making it a source of valuable insight as to how the probe would function in this type of environment.

The graph on Figure 13 is a representation of the measurement with the tuning wires attached and the antenna matched for 2 GHz. Obviously there is a significant reduction of the peak to peak ratio as is to be expected since attenuation increase with frequency. Even so the signal is still clearly distinguishable. This is exactly what one would expect when taking into account the previous two measurements at 1 GHz and 400 MHz.
5.2.2 Testing and measurement (Mark 2.2)

5.2.2.1 Measurement at 1 GHz

After the series of successful measurements with the Mark 2.1 probe it was clear that the probe redesign has resulted in a good merit of performance improvement. The same tuning wire set used for both the Mark 2.1 and Mark 2.2 probes, since they operate under the same principles. Since the probes using the matching coils network did not provide a sensitivity boots, measurement with those was omitted for the case of the Mark 2.2 probe design case, as the same behavior was expected as for the Mark 2.1 case.

The measurement graph at 1GHz frequency is presented on Figure 14. The Mark 1.2 performed better than the Mark 1.1. This is also the case for the Mark 2.1 vs. the Mark 2.2. The peak to peak ratio, though not significantly is still a bit higher for the probe using the video switch. The video switch makes for a better reflecting probe, because of its low input resistance in the on state.
5.2.2.2 Measurement at 400 MHz

At 400 MHz results were as expected for the Mark 2.1 probe. There was a very clearly detectable signal due to its peak to peak ratio of 4 dB as seen on Figure 15. It is a significant improvement over the initial Mark 1.2 probe where the signal was barely distinguishable, thanks to both the tuning and the use of the low resistance video switch.
5.2.2.3 Measurement at 2 GHz

Finally a measurement at 2GHz was performed. From Figure 16 a peak to peak ratio of around 1 dB at 2 GHz is evident, which is lower than the 4 dB at 400 MHz as expected. One would expect it to also be lower than the 1 GHz, however this was not the case. It is to be investigated in future research why that is.

![Figure 16: MST probe Mark 2.2 (tuned) measurement at 1GHz](image)

5.2.3 Results, Discussion and possible improvement strategies

Several things can be noted after the Mark 2 measurement cycle’s completion.

First of all the system’s performance has improved significantly. All the measurements produced viable results for both the Mark 2.1 and Mark 2.2 probes. An exception are those made with the matching coils probes, which obviously could not tune the antenna precisely enough due to the fact that the coils inductance values were calculated with an error. This can be attributed to the unknown switch impedance, since it was not available in the manufacturer’s datasheet.

Precisely tuned extension wires results in a great increase of the received signal level. Furthermore, the modulation depth was increased as well. As expected with the increase of frequency the level decreases as is expected due to the higher attenuation. So, the MST system behaves as a regular probe working in the receiver mode.

An exception is the case where the signal level at 2 GHz measured with the Mark 2.2 probe exceeds the one registered at 1 GHz, which is counterintuitive since there is a decrease in attenuation due to the lower frequency. At this point it cannot clearly be stated why this effect is occurring.
In the end the Mark 2 probe can be considered to fulfill the requirements for the MST probe functionality, however at the expense of increasing physical size due to the extension of the antenna.

The next step is to validate the already functioning Mark 2 probe for the case of human tissue as a propagation medium, which is presented in the next chapter.
6 Body Phantom Experimental Testing and Measurements

For the experiment the bistatic setup was used again, similar to the air measurements. There are no significant differences in the measurement setup as compared to the previous one. Of course since now a phantom is used, the presence of the tank filled with the human tissue phantom liquid will impact the measurements. The location of the probe with respect to the receiving and transmitting antenna remains the same, which means that attenuation due to path distance remains the same and any additional signal degradation will be due to mismatch and reflection due to the new materials introduced in the setup. In this way a comparison between the measurements in air and in human tissue phantom can be made based solely on the different medium properties, since the other setup parameters remain the same. The setup is described in greater detail in Appendix B.

6.1 MST probe Mark 3

Water is used as medium in the first experiment. The disadvantage is unknown precise permittivity. However taking into account that demineralized water has $\varepsilon_r \approx 78$, we can say we are working in a medium with much higher permittivity than the air we used in the previous measurements, which is also much closer to this of human tissue, which has $\varepsilon_r \approx 40 \div 70$. This falls well within the main task, which is to validate the system operational principles, rather than do precise measurements.

6.1.1 Testing an measurement for tap water as a medium

As mentioned before normal tap water is used since it is easy to acquire. The disadvantage is that when it is a cheap base for the measurement phantom its precise permittivity is not known. However taking into account that demineralized water has $\varepsilon_r \approx 78$, we can definitively say we are working in a medium with much higher permittivity than the air we used in the previous measurements, which is also much closer to this of human tissue, which has $\varepsilon_r \approx 40 \div 70$. This falls well within the main task, which is to validate the system operational principles, rather than do precise measurements.

6.1.1.1 Testing and Measurement (Mark 3.1)

6.1.1.1.1 Measurement at 400 MHz

The measurement data can be seen on Figure 17. Clearly the signal is detected, as well as the switching pattern. The probe is functional. Experiments were performed with different versions of the probes (e.g. different antenna lengths).

Optimal results were obtained using longer antenna lengths than half wavelengths, due to the unknown precise permittivity of the phantom.
6.1.1.1.2 Measurement at 600 MHz

The next measurement was performed at 600 MHz. The same probe, that was tuned for 200 MHz produced received signal with higher level at 600 MHz as compared to 400 MHz. Measurements results at 600 MHz are presented on Figure 18.

Figure 17: MST probe Mark 3.1 at 400 MHz in tap water (200MHz tuning)

Figure 18: MST probe Mark 3.1 at 600 MHz in tap water (200MHz tuning)
6.1.1.1.3  Measurement at 800 MHz

Measurements at 800 MHz are presented in Figure 19. Signal level is lower than at 600 MHz, which is expected.

![Figure 19: MST probe Mark 3.1 at 800 MHz in tap water](image)

6.1.1.2  Testing and Measurement (Mark 3.2)

Measurements at various frequencies were not successful. At 900 MHz some results were obtained.

The measurement plot at 900 MHz is presented in Figure 20. In closer inspection the shape seems to resemble a series of rectangular pulses. A zoomed version of the reading is presented on Figure 21. The rectangular pulses are clearly distinguishable, even though their shape appears to be distorted by noise.

The received signal is obviously very weak, the peak to peak ratio is on the order of 0.01 dB, which is a lot smaller than anything considered to be readable up to now.
6.1.2 Results, Discussion and possible improvement strategies

From theory it is known that the dipole antenna is most efficient at half wavelength, however this was not the case when measuring with the Mark 3 probe. This is due to the fact that tuning has been done assuming a value of \( \varepsilon_r = 78 \) for the phantom solution. The result is that the \( \lambda/2 \) length assumed for the antenna length calculation is shifted with respect to the actual
value. For this reason the antenna is actually tuned for a frequency shifted from the one the devise was initially tuned for.

However the probe still performed a dipole antenna according theory. It was best tuned around a central frequency and with increase/decrease in frequency, performance degraded.

The Mark 3 probe clearly displayed such behavior. It performed bad at 400 MHz, improved at 600 and degraded with frequencies of 800 MHz and above. This was the case for the Mark 3.1 probe, which appears to be tuned for frequencies around 600 MHz. The Mark 3.2 probe was tuned for frequencies above 900 MHz, which we could not validate due to the generator’s limitations. This being said we can conclude that tuning was not done properly, due to the fact that we had no precise measured value of the tap water permittivity. Assuming that \( \varepsilon_r \) of the medium was 78 was obviously wrong.

The measurement cycle for the Mark 3 probe did however prove the system works in the case of a human bod tissue phantom, however one consideration for future research is that deviations in the permittivity of the solutions could lead to great antenna mismatch. This in turn can lead to the received signal dropping below the noise floor and making it undetectable. It would seem this is a serious issue and should be investigated in greater detail in future research.
7 Conclusion and Future Work

The extensive testing and measurements performed over the course of this thesis are the base for the following conclusions.

The proposed measurement system functions well. It is able to measure signals both in air and in a human body tissue phantom. After proper tuning sensitivity is boosted enough to compensate for attenuation and reflection effects. The probe meets the requirements for IBC systems – it is small, power efficient and due to the use of MST field perturbation is not an issue.

The novel method of using MST for the purpose of intra-body channel characterization in WBAN and IBC systems proposed is not perfect.

There are still issues with the probe design to be solved.

The issue of investigating the measured with the Mark 2.2 probe signal at 2 GHz, exceeding the one registered at 1 GHz needs to be investigated further. This case is counterintuitive since there is a decrease in attenuation due to the lower frequency. Future research could include measurements over a greater number of frequencies around the 1GHz and 2 GHz bands in order to characterize why the aforementioned phenomena occurred.

Proper tuning for $\lambda/2$ length in the case of the liquid phantom is still required. This should be done by first measuring the medium permittivity via a permittivity probe and then calculating the wave phase velocity precisely. In this way accurate tuning can be implemented and sensitivity boosted further.

The onboard circuitry should be integrated into a single chip, reducing size and power consumption.

A major design improvement for a future system would be to make the MST probe 3D. Also the measurements should be performed in an anechoic chamber in the future.

Possible improvements could include the introduction of a more complex phantom. A multilayered inhomogeneous structure would provide for a better representation of the human body. In addition sensitivity could be boosted by employing an I/Q demodulator at the receiver as well using the MST in monostatic configuration.

Another improvement could be the use of a monostatic setup, instead of bistatic. This will require additional equipment, however it will boost sensitivity.
8 Appendix

Appendix A – Debye – Stogryn method and Matlab script calculation

8.1 Debye – Stogryn Model Calculation Algorithm

According to the Strogyn model [37], the complex permittivity of saline water is represented by the Debye equation of the form:

\[ \varepsilon = \varepsilon_\infty + \frac{\varepsilon_0 - \varepsilon_\infty}{1 - i\omega \tau} + \frac{i\sigma}{2\pi \varepsilon_0 \omega f} \]

where

\[ i = \sqrt{-1} \]

\( \varepsilon_0 \) static dielectric constant of the solvent (water in this case)

\( \varepsilon_\infty \) optical dielectric constant of the solvent

\( \tau \) relaxation time

\( \varepsilon_0^* \) permittivity of free space (air) - \( \varepsilon_0^* = 8.854 \times 10^{-12} \text{[F/m]} \)

\( \sigma \) ionic conductivity of the dissolved salt (NaCl in this case) [mho/m]

\( f \) RF wave frequency [Hz]

The values of \( \varepsilon_0 \) and \( \tau \) as a function of temperature and normality are given by:

\[ \varepsilon_0(T, N) = \varepsilon_0(T, 0)a(N) \]

\[ 2\pi\tau(T, N) = 2\pi\tau(T, 0)b(N, T) \]

where

\( T \) temperature of water \( \text{[°C]} \)

\( N \) normality of the solution

According to [36] the following can be calculated:

\[ a(N) = 1.00 - 0.2551N + 5.151 \times 10^{-2}N^2 - 6.889 \times 10^{-3}N^3 \]

\[ b(N, T) = 0.1463 \times 10^{-2}NT + 1.000 - 0.04896N - 0.02967N^2 + 5.644 \times 10^{-3}N^3 \]

\[ \varepsilon_0(T, 0) = 87.74 - 0.40008T + 9.398 \times 10^{-4}T^2 + 1.410 \times 10^{-6}T^3 \]

The relaxation time for pure water is obtained by [36]:

\[ 2\pi\tau(T, 0) = 1.1109 \times 10^{-10} - 3.824 \times 10^{-12}T + 6.938 \times 10^{-14}T^2 - 5.096 \times 10^{-16}T^3 \]
The normality can be calculated from the salinity $S$ as:

$$ N = S(1.707 \times 10^{-10} + 1.205 \times 10^{-5}S + 4.058 \times 10^{-9}S^2) $$

where

$S$ salinity in parts per thousand ($0 \leq S \leq 260$)

For NaCl solutions the ionic conductivity $\sigma$ from the basis Debye equation is calculated as [37]:

$$ \sigma_{NaCl}(T, N) = \sigma_{NaCl}(25, N)[1.000 - 1.962 \times 10^{-2}\Delta + 8.08 \times 10^{-5}\Delta^2 $$

$$ - (3.020 \times 10^{-5} + 3.92 \times 10^{-5}\Delta) + N[1.721 \times 10^{-5} - 6.584 \times 10^{-6}\Delta] $$

where

$\Delta = 25 - T$, and

$$ \sigma_{NaCl}(25, N) = N[10.394 - 2.3776N + 0.68258N^2 - 0.13538N^3 + 1.0086 \times 10^{-2}N^4] $$

The last remaining parameter to fill in the Debye equation is the optical dielectric constant of the solvent $\varepsilon_\infty$. There is no evidence in literature that $\varepsilon_\infty$ depends on salinity. There is some evidence that it slightly depends on the temperature of the solution, however not all of the research agrees on whether it increases or decreases with temperature. This is the reason that normally a constant value is used, which in this case is $\varepsilon_\infty = 4.5$, since it has also been chosen in [37] and agrees well with measurement data.

Based on the presented algorithm and the measurement values for the salinity and the temperature of the solution one can easily calculate the permittivity. Chapter 8.2 presents the specifics of the algorithm used in this thesis.
8.2 Debye – Stogryn Model Calculation Algorithm – Matlab Implementation

Below is a copy of the script, used for calculating the permittivity of the phantom liquid, based on the measured salinity and temperature. It is of two identical parts, which allow for varying both the salinity and temperature independently and observing the results on the permittivity of the medium

```matlab
% Start of Script
close all;
clear all;
clc;

% Parameter Definition

% Frequency of Operation in Hz
freq = 4*10^6;
% Permittivity of Free Space in F/m
eps_fs = 8.854*10^-12;
% Permittivity for Optica Frequencies 4.5 or 4.9 depending on source)
eps_inf = 4.9; % (4.5 / 4.9)
% Temperature in Degrees Celsius
temp = 25;
% Random parameter (no idea what it represents)
delt = 25 - temp;
% Salinity of the Solution in parts per Thousand (0<salin<260)
salin = [0.5 2 4 8];

% Parameter Calculation for the Strogyn method

for i = 1:length(salin)
    norm(1,i) = salin(1,i)*(1.707*10^-10 + 1.205*10^-5*salin(1,i) + 4.058*10^-9*(salin(1,i)^2));
    sigma_NaCl_25_norm(1,i) = norm(1,i)*(10.394 - 2.3776*norm(1,i) + 0.68258*(norm(1,i)^2) - 0.13538*(norm(1,i)^3) + 1.0086*10^-2*(norm(1,i)^4));
    sigma_NaCl_temp_norm(1,i) = sigma_NaCl_25_norm(1,i)*(1.000 - 1.962*10^-2*delt + 8.08*10^-5*(delt^2) - (3.020*10^-5 + 3.92*10^-5*delt) + norm(1,i)*(1.721*10^-5 - 6.584*10^-6*delt));
    c_2_pi_tau_temp_0 = 1.1109*10^-10 - 3.824*10^-12*temp + 6.938*10^-14*(temp^2) - 5.096*10^-16*(temp^3);
    eps_0_temp_0 = 87.74 - 0.40008*temp + 9.398*10^-4*(temp^2) + 1.410*10^-6*(temp^3);
end
```

53
\[ b_{\text{norm temp}}(1,i) = 0.1463 \times 10^{-2} \cdot \text{norm}(1,i) \cdot \text{temp} + 1.000 - 0.02967 \cdot (\text{norm}(1,i))^2 + 5.644 \times 10^{-3} \cdot (\text{norm}(1,i))^3; \]

\[ \alpha_{\text{norm}}(1,i) = 1.000 - 0.2551 \cdot \text{norm}(1,i) + 5.151 \times 10^{-2} \cdot (\text{norm}(1,i))^2 - 6.889 \times 10^{-3} \cdot (\text{norm}(1,i))^3; \]

\[ c_{\text{2 pi tau temp norm}}(1,i) = c_{\text{2 pi tau temp 0}} \cdot b_{\text{norm temp}}(1,i); \]

\[ \epsilon_{\text{0 temp norm}}(1,i) = \epsilon_{\text{0 temp 0}} \cdot \alpha_{\text{norm}}(1,i); \]

\[
\begin{align*}
% Debye equation and Permittivity Printing \\
\text{display('---------------------------------------------------------------------')} \\
\text{display('Permitivity Calculations in Accordance with the Debye - Strogyn Model')} \\
\text{display('---------------------------------------------------------------------')} \\
\text{display('Temperature of solution in C'); disp(temp);} \\
\text{for } j = 1: \text{length(salin)} \\
\epsilon(1,j) = \epsilon_{\text{inf}} + (\epsilon_{\text{0 temp norm}}(1,j) - \epsilon_{\text{inf}})/(1 - 1i \cdot c_{\text{2 pi tau temp norm}}(1,j) \cdot \text{freq}) + \\
(1i \cdot \sigma_{\text{NaCl temp norm}}(1,j))/(2 \pi \cdot \epsilon_{\text{fs}} \cdot \text{freq}); \\
\text{display('Salinity of the solution in grams per litre'); disp(salin(1,j));} \\
\text{disp(eps(1,j));} \\
\text{end} \\
\end{align*}
\]

\[
\begin{align*}
% Parameter Definition \\
% Frequency of Operation in Hz \\
freq = 4 \times 10^6; \\
% Permittivity of Free Space in F/m \\
\epsilon_{\text{fs}} = 8.854 \times 10^{-12}; \\
% Permittivity for Optica Frequencies 4.5 or 4.9 depending on source) \\
\epsilon_{\text{inf}} = 4.9; % (4.5 / 4.9) \\
% Temperature in Degrees Celsius \\
\text{temp} = [25 35 45 55]; \\
% Random parameter (no idea what it represents) \\
delt = 25 - \text{temp}; \\
% Salinity of the Solution in parts per Thousand (0<salin<260) \\
salin = 0.5; % (0<salin<260) \\
\end{align*}
\]

\[
\begin{align*}
% Variable preallocation (for speeding up script) \\
norm = \text{zeros}(1, \text{length(salin)}); \\
sigma_{\text{NaCl temp norm}} = \text{zeros}(1, \text{length(salin)}); \\
\text{b norm temp } = \text{zeros}(1, \text{length(temp)}); \\
\alpha_{\text{norm}} = \text{zeros}(1, \text{length(salin)}); \\
\end{align*}
\]
% Parameter Calculation for the Strogyn method

for i = 1:length(temp)
    norm = salin*(1.707*10^-10 + 1.205*10^-5*salin + 4.058*10^-9*(salin^2));
    % Normality of the Solution, Based on Measured Salinity (by a scale 0.1 resolution scale)
    sigma_NaCl_25_norm = norm*(10.394 - 2.3776*norm + 0.68258*(norm^2) - 0.13538*(norm^3) + 1.0086*10^-2*(norm^4));
    sigma_NaCl_temp_norm(1,i) = sigma_NaCl_25_norm*(1.000 - 1.962*10^-2*delt(1,i) + 8.08*10^-5*(delt(1,i)^2) - (3.020*10^-5 + 3.92*10^-5*delt(1,i)) + norm*(1.721*10^-5 - 6.584*10^-6*delt(1,i)));
    c_2_pi_tau_temp_0(1,i) = 1.1109*10^-10 - 3.824*10^-12*temp(1,i) + 6.938*10^-14*(temp(1,i)^2) - 5.096*10^-16*(temp(1,i)^3);
    eps_0_temp_0(1,i) = 87.74 - 0.40008*temp(1,i) + 9.398*10^-4*(temp(1,i)^2) + 1.410*10^-6*(temp(1,i)^3);
    b_norm_temp(1,i) = 0.1463*10^-2*norm*temp(1,i) + 1.000 - 0.04896*norm - 0.02967*(norm^2) + 5.644*10^-3*(norm^3);
    alpha_norm = 1.000 - 0.2551*norm + 5.151*10^-2*(norm^2) - 6.889*10^-3*(norm^3);
    c_2_pi_tau_temp_norm(1,i) = c_2_pi_tau_temp_0(1,i)*b_norm_temp(1,i);
    eps_0_temp_norm(1,i) = eps_0_temp_0(1,i)*alpha_norm;
end

% Debye equation and Permittivity Printing

for j = 1:length(temp)
    eps(1,j) = eps_inf + (eps_0_temp_norm(1,j) - eps_inf)/(1 - 1i*c_2_pi_tau_temp_norm(1,j)*freq) + (1i*sigma_NaCl_temp_norm(1,j))/(2*pi*eps_fs*freq);
    display('Temperature range of the solution in C');
    disp(temp(1,j));
    display('Permittivity of the solution for the aforementioned temperature in C and salinity in gramms per litre');
    disp(eps(1,j));
end

% End of Script
Appendix B – Board Schematics/Layouts, Supplementary Graphs

8.3 MST Probe Mark 1.1

8.3.1 MST Probe Mark 1.1

Below is a short list of the components used to build the MST Mark 1.1 probe. Those are for the main functionality, not including any resistors, capacitor or inductors required for the proper function of the microcontroller and switch (those can be seen from the circuitry schematic below as well from the manufacturer datasheet).

- **Switch:**

A Texas Instruments SPST CMOS Analog Switch (TS12A4514) has been used. It has more than 40 dB insulation, fast switching time (10 ns) and low power consumption of several $\mu A$. It can operate in several modes depending on the power supply voltage – this thesis uses the switch in its 3V power supply mode, which is the lowest available.

- **Timer IC:**

The timer generating the rectangular pulses running the switch is realized by a microcontroller of the type Microchip 12F509. It has been programmed to run at a frequency of 1kHz with a duty cycle of 50% in order for it to produce the rectangular pulses needed by the analog switch in order for it switch between the two states. This is essentially the modulation frequency for the AM signal.

Measurement at the output of the timer validates the pulse parameters. The measurement data from a digital oscilloscope, which has been reformatted via Matlab is shown on Fig. 22.

![Figure 22: Switch Timer output voltage](image-url)
The code used to program the IC logic is presented below:

```c
#define GP0 PIN_B0
#define GP1 PIN_B1
#define GP2 PIN_B2
#define GP3 PIN_B3
#define GP4 PIN_B4
#define GP5 PIN_B5

setup_oscillator(OSC_4MHZ);
#endif

#use delay(int=4000000)
#endif

void main()
{
output_low (GP0);
while (TRUE)
{
output_high (GP1);
delay_us (500);
output_low (GP1);
delay_us (500);
}
}
```

- Power supply

A standard CR2032, 200 mAh battery has been chosen to power the probe (required due to the presence of the analog switch and the microcontroller timer). There are far smaller options, which would require less space, making it possible to reduce probe dimensions. The cost of those is also similar to the CR2032 in general. However in most cases they have significantly reduced capacity and would not allow for prolonged probe operation. While this would not be a problem for an actual IBC devices like a capsule endoscopy probe for example this is not the case for a measurement setup. Testing and measurements are performed over the course of hours, spanning throughout multiple days/weeks. This coupled with the fact that the probe is encapsulated in plastic in order to make it water tight, makes it impossible to replace any components without breaking the insulation layer. Needless to say replacing the battery every couple of days is not efficient an gives reason to why such a big battery, with large capacity was chosen.
For the first build of the probe the battery has been replaced by a wire connection to an external power supply in order to be able to build a working probe with greater ease and speed. It is also a better solution to feed the device with a power line when tuning it and then replace the connections with the proper battery later on, when the devices was tuned to work as intended.

- **Dipole**

The dipole antenna wires are made of an annealed single-stranded copper conductor with a diameter of 1.5 mm and length of 15 mm each. It is of the type coated with polymeric insulation commonly used for coils/transformer winding. This particular type has been chosen since it offers good conductivity as well as mechanical rigidness due to the fact of being single-stranded. The coating provides the benefit of electrical insulation and gives additional protection in the case of imperfect water insulation to the circuitry.

- **PCB board**

For the substrate of the PCB board FR4 material has been used, with dimensions: 1 x 20 x 30 (w x l x h in mm).

A schematic of the circuitry responsible for the switching functionality is presented on Fig. 23 and on Fig. 24 is the PCB Layout of the circuitry. A one sided design has been utilized. It is of low complexity, there are no via holes, it is functional and simple to construct.

Unfortunately no snapshot of the Mark 1.1 probe is available since no photographic equipment was available at the time of testing and later on The Mark 1.1 probe was modified in order to produce the Mark 2 probe. However the only structural difference between the Mark 1.1 and Mark 1.2 is in the switching component due to the different package dimensions. Overall the two probes look very similar in shape and size.

![Figure 23: MST probe Mark 1.1 Circuitry](image-url)
8.3.2 MST Probe Mark 1.2

- **Switch:**

  The Video Switch is Texas Instruments Quad SPDT Wide-Bandwidth Video Switch with Low On-State Resistance (TS5V330). The fact that it has a low on-state resistance means that it will more effectively short the two wires, in order to better simulate the behavior of a single wire. This should improve sensitivity, since it will lead to a larger modulation index.

- **PCB board**

  Besides the difference in the switch, the two probes are identical with the exception of the onboard track routing due to the difference in package of the two switches (one is an 8 pin the other is 16 pin), the board dimensions are still the same. The schematic for the MST Probe Mark 1.2 is presented on Fig. 25, whereas MST probe Mark 1.2 PCB layout can be seen on Fig. 26, Figure 27 is a snapshot of the Mark 1.2 probe.
The MST Mark 2 probe is an improved version of the Mark 1 probe. It has been designed with one major goal in mind – to make it possible to flexibly regulate the electrical length of the dipole antenna in order to match it for $\lambda/2$, resulting in a sensitivity boost.

There were two variations depending on how the above was done and each has been implemented for both the Mark 2.1 and Mark 2.1 probe in order to investigate their effectiveness. This lead to building four separate devices, the characteristics of which are discussed below.

First let us present the two antenna tuning methods used:
8.4.1 **Antenna tuning – electrical method**

Initially since probe dimensions were considered to be one of the most important design parameters, a solution was used that changed the electrical length of the antenna without influencing physical size. Two identical, adjustable coils were utilized in order to change the effective electrical length of each of the antenna rods. The values for the respective coil inductance has been specifically chose. The procedure is described in great detail in [43], which is why we will only give a summary of it for the purpose of this thesis.

There are two parameters that are needed for the antenna tuning algorithm [43], the input impedance of the switching element, and the dipole characteristic impedance.

The input impedance of both switches was obtained via the same measurement procedure. A Vector Network Analyzed was used to measure the S11 parameter of the two devices. A conversion was made via a numerical method later on to antenna input impedance. It is to be noted that a custom made SMA cable was used to connect the VNA to the switch via soldering it directly to the board. This means there was no proper calibration, making this procedure the main contributor to any errors as far the matching coil inductance calculation goes. The antenna impedance was calculated numerically [44],[45].

8.4.2 **Antenna tuning – mechanical method**

The other possibility for tuning was to physically increase the length of the antenna wires. This is a rude solution but it provides flexibility, since attaching a piece of wire to the antenna via a crocodile is an easy and quickly accomplished task. In addition several wire lengths can be used and easily manufactured in order to quickly change the antenna tuning frequency in order to measure in another frequency band. This type of solution does have the disadvantage that for microwave frequencies the $\lambda/2$ wavelength is on the order of several tens of centimeters and makes the devise incomparably large to what one would expect for a IBC system. A snapshot is presented on Fig. 28, where one can see the wire extensions.
8.4.3 MST Probe Mark 2.1

The updated schematic for the Mark 2.1 probe is presented on Fig. 29, its PCB Layout on Fig. 30, and a snapshot of the complete probe on Fig. 31. Upon inspection of Fig. 31 one can see both the pair of tuning coils and the extensions added to the antenna wires where the crocodile connections are to be attached. Another notable fact is that the coils are quite small and do not increase the footprint of the probe, unlike the antenna extensions.
8.4.4 MST Probe Mark 2.2

For the case of the MST probe Mark 2.2 the same investigation was performed as for the Mark 2.1 as mentioned before. The antenna impedance has been calculated numerically and the switch on and off state impedance measured. As expected since we have different parameter value, due to the different switch impedance, the coil inductance value was also different. The Mark 2.2 Circuitry Schematic is presented on Fig. 32, its PCB Layout on Fig. 33, a device snapshot can be seen on Fig. 34.
Figure 32: MST probe Mark 2.2
Circuitry

Figure 33: MST probe Mark 2.2
PCB layout

Figure 34: MST probe Mark 2.2 - Snapshot
8.5 MST Probe Mark 3

First of all, the tuning coils were removed from the Circuitry Schematic, since they did not serve their intended purpose as explained before. This resulted in the need to modify the board design, by shorting the pads used for the now removed coils. This was found not to influence probe operation in any way.

Second a way to make the probe water tight had to be thought of, in order not to short out the circuitry when it was fully submerge in a substance primarily composed of water. Simply coating the whole device, including the board, components, battery and the antenna rods with a water tight plastic seemed like the most efficient solution. This has been done by pouring melted plastic using a hot glue gun. The process is easy to implement, is not time consuming and minimizes costs.

While it has a number of disadvantages since it creates antenna mismatch and adds an additional medium for RF wave propagation it comes with its benefits as well. It provides water resistance and a way to affix the probe so it does not float in the phantom liquid holding container, providing for a more accurate measurement.

A snapshot of the setup is presented on Fig. 35, the glue sealed probe can be seen on Fig 36, the probe inside the holding container is presented on Fig. 37

![Figure 35: Bistatic MST Testing and Measurement Setup](image)

Medium – Tap Water
8.5.1 MST Probe Mark 3.1

As mentioned in the previous section there were some changes in the MST probe design. On Fig. 38 one can see the modified circuitry of the Mark 3.1 probe. Since the coils were removed, the design reverted to the initial one, thus it is identical to the one for the Mark 1.1. The PCB Layout however is different, compared to both the Mark 1.1 and Mark 2.1. The component placement has been optimized. As a result the board dimensions have been
reduced, making the probe even smaller with dimensions closer to those of a IBC device such as a capsule endoscopy probe for example. Another important advantage that comes with the new design is that now the two pads connected to the switch output are now positioned closer. In a sense this makes it so that when the switch is closed the dipole is a much closer representation of a conventional antenna, where there is no switch in-between the wires and perfect symmetry is obtained. What this would lead to is better separation between the two state level, providing for a higher modulation index. The probe PCB Layout is presented on Fig. 39.

In order for the reader to obtain insight in what the new measurement setup looks like, two snapshots of it are presented for the Mark 3.1 probe case. Fig. 40 shows the whole setup with the probe container including the lid, which is present during the actual measurement process. Fig. 41 is a close up of the container with the lid remove, where one could see the phantom liquid solution the probe is submerged in. As it can be seen a simple plastic container initially meant for holding food items was used, however since it is made of radio transparent plastic material it is quite suited for initial testing. Also it provides an addition medium boundary, worsening propagation conditions, adding more reflections.
Figure 40: MST probe Mark 3.1 - Snapshot

Figure 41: MST probe Mark 3.1 – Snapshot Close-up
8.5.2 MST Probe Mark 3.2

The case for the Mark 3.2 probe is the same as the one for the Mark 3.1 probe. The coils have been removed from both the Circuitry Schematic presented on Fig. 42 and the PCB Board Layout presented on Fig. 43. The board dimensions have been optimized, antenna pads have been relocated closer together. Snapshots of the device can be seen on Fig. 44 and Fig. 45.

It is to be noted that a container made by a different manufacturer was used, with slightly different dimensions and composition, however still made from a radio transparent material. The reason for using a different container was to prove that the system functionality is not influence by the container shape or material.

![Figure 42: MST probe Mark 3.2 Circuitry](image)

![Figure 43: MST probe Mark 3.2 PCB Layout](image)
Figure 44: MST probe Mark 3.2 – Snapshot

Figure 45: MST probe Mark 3.2 – Snapshot Close-up
8.5.3 Comparison between probe designs.

Figure 46, 47, and 48 present the MST Mark 1, 2, and 3 probes in order of their construction. This way the reader can gain more insight into how the design evolved and improved.

Figure 46: MST Probe Mark 1

Figure 47: MST Probe Mark 2
Figure 48: MST Probe Mark 3
Appendix C – Measurement Phantom Physical Characteristics

Below is a general outline of the relevant body phantom parameters. Initially the phantom has been designed to be as simple as possible both from geometrical perspective and tissue layering. It is of the liquid type and is of a single layer. This design has the advantage of flexibility, since it can easily be extended to a more complex scenario in order to more accurately represent the human body.

This can be done via modifying the liquid phantom with a thickening agent (e.g. agarose), making it a gel one. Such a phantom allows for multi-layering and better probe mechanical stability. The latter will eliminate any unwanted perturbation due to movement.

A short summary of the main characteristics of both the phantom solution itself and the phantom holding container is presented in Table. 3 and Table. 4, respectively.

<table>
<thead>
<tr>
<th>Human Tissue Phantom</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Type</td>
<td>Liquid</td>
</tr>
<tr>
<td>Material</td>
<td>Water + NaCl</td>
</tr>
<tr>
<td>Layering type</td>
<td>Single</td>
</tr>
<tr>
<td>Structure</td>
<td>Homogeneous</td>
</tr>
</tbody>
</table>

Table 3: Phantom parameters

<table>
<thead>
<tr>
<th>Holding Container</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Material</td>
<td>PET</td>
</tr>
<tr>
<td>Shape</td>
<td>Generic square box</td>
</tr>
<tr>
<td>Dimensions</td>
<td>N/A</td>
</tr>
</tbody>
</table>

Table 4: Phantom holding container parameters

Several important considerations for the phantom design are to be mentioned:

First it should not be too large, since this will require a large volume of the solution to completely fill the vessel in order to have a single layered homogeneous structure. At the same time it should not be too narrow or short since there will not be enough material for the wave to propagate through and later on if more layers are introduced and the position of the transmitter is changed there might be constraint issues. Also the size should be in order with what an average adult human body is with respect of dimensions.

Second there is the option to adjust the electromagnetic properties of the solution dynamically. The addition of NaCl in precise quantities (measured by a 0.1 g resolution scale) regulates the imaginary part, whereas changing the temperature of the liquid by heating/cooling it (measured by a thermo coupler) regulates the real part of the permittivity of the phantom.

Last it should be noted that the constructed phantom is suited for the purpose of the thesis. It provides a dynamically adjustable electromagnetic medium, which is easy to construct and
operate. It is, however a simplified representation of the complex human body, because of its single layer homogeneous structure, which makes it not particularly accurate.

Possible improvements for future work, wielding a more precise measurement data set: the precision of the measurement, include: the use of a multilayered, inhomogeneous material gel, with precisely adjusted dimensions and holding container made from a material representing human skin for example.
9 References


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