Channel Characterization and Signal Propagation Studies for Wireless Galvanic Coupled Body Sensors

Meenupriya Swaminathan, Ferran Simon Cabrera, Gunar Schirner, Member, IEEE, and Kaushik R. Chowdhury, Member, IEEE

Abstract

New medical procedures involving continuous patient monitoring through human body sensors are becoming commonplace with micro-scale implanted sensors transferring information to on-surface macro-scale sensors for further data retrieval and analysis. Traditional forms of radio frequency-based wireless communication find limited use in such scenarios owing the limited penetration of electromagnetic waves through human tissue, and the need for frequent battery replacements. Instead, we propose a radically different form of wireless communication in this paper that involves galvanic coupling using extremely low energy electrical signals. The main contributions in this paper are: (i) developing a theoretical model suite to estimate the channel gain, where the body itself serves as the communication channel, (ii) obtaining an estimate for the observed noise and achievable data rates, and (iii) identifying the optimal transmission frequency and electrode placements for signal propagation through tissue. We propose two equivalent circuit models to characterize the channel, based on the theories of two-port and lumped element circuit design, which are then validated through extensive simulations using finite element method and known experimental measurements. Our results reveal a close agreement between theory, simulation and experimental findings, suggesting a promising case for the adoption of galvanic coupling-based communication for future intra-body sensors.
Channel Characterization and Signal Propagation Studies for Wireless Galvanic Coupled Body Sensors

Index Terms

Body coupled communication, galvanic coupling, channel model, implanted sensors, signal gain analysis.

I. INTRODUCTION

Body area networks (BANs) promise to usher in dramatic improvements in personalized medicine, implant-based in-situ monitoring, and controlled drug injection, among others. It establishes a communication channel between a remote monitoring entity and the body sensors engaged in actuating or in gathering data in the above applications. This ensures not only offline analysis, but also an opportunity for specialists to immediately detect and react to any abnormal changes by issuing actuating directives to the deployed sensors, such as releasing controlled amounts of a drug. In many cases, retrieval of these sensors for battery replacements is impractical, requiring efforts on prolonging their network lifetimes. In this paper, we envision a network of sensors without radio frequency (RF) communication, and instead, use weak electrical signals passing through the human body that dramatically lowers energy consumption. The key contribution in this work is building an electrical-equivalent channel model that is able to predict the gain of the signal propagation across the skin as well as the inner tissues, for a variety of transmitter-receiver distances, frequency and tissue thicknesses. The macro-scale sensors are operated on the surface to aggregate the information from micro-scale implants. We adopt the concept of galvanic coupling to realize this energy efficient non-RF communication paradigm.

A. Wireless communication through galvanic coupling:

In galvanic coupled communication, a pair of electrodes directly couple weak electric signals in the order of 1 mA at 0.05 V to body [5], as shown in fig.1. The induced magnetic field is well below 2mA/m, the permissible limit within the body [17]. Majority of the induced current coupled to the body passes through the return path of transmitter (represented by black arrow). A part of coupled signal propagates through the body and reaches the receiver electrodes (illustrated by gray arrows). The difference in voltage is detected by a corresponding pair of receiver electrodes that belong to a second macro-sensor, also placed on the skin. Note that there is no common ground required here, as in the case of capacitive coupling [1]. A characteristic feature of galvanic coupled communication is that the signal has a dominant component that passes to the receiver electrodes through the inner tissue layers, even when the transmitter is placed on the skin. Thus, apart from being energy efficient when compared to RF, the
communication in this scenario also becomes less impacted by environmental noise. A carefully designed galvanic coupling with selected signal amplitude and frequency, has a dominant component that propagates through specific tissue layers. Thus more than one transmission along similar pathways is possible. Note that this behavior is in contrast to the case of RF propagation, wherein other transceivers must be silenced owing to the broadcast nature of the medium.

Recent efforts have demonstrated the feasibility of using galvanic coupling for intra-body communication. A signal transfer model assuming null inductance and capacitance effect of a single tissue layer below 100 MHz was pioneered in [8], [12]. A transfer function for multi-layer tissue impedance using a 3D circuit model was derived in [9], and it was found that an increase in longitudinal (measured along the surface length) skin impedance lowers the upper cut-off frequency, i.e., the upper signal frequency bound that resulted in correct reception. Similarly, an increase in transverse (measured along the tissue depth) impedance shifts the lower signal cut off frequency up on the frequency scale. They also investigated the channel transmission characteristics for implantable electrodes, and compared the results with a numerical model and a phantom reflecting typical muscle-tissue properties at 27 MHz [10]. [6] presents an equivalent distributed parameter based circuit model for surface sensors, in which the propagation characteristics in a frequency range from 10 kHz to 1 MHz are calculated.

B. Research motivation:

While the suitability of galvanic coupling for BAN has been demonstrated, the state of the art has been limited to on-surface communication, i.e., with the transmitter and receiver placed on the skin. To the best of our knowledge there is no work that studies the scenarios of implanted micro-sensors, where the sources may be placed deep inside body tissues, like the muscle, with the receivers being at the same embedded micro-level or serving as macro-scale signal pick-up points on the skin. The field distribution arising out of galvanic coupling for the embedded sensors in the inner tissues has not been adequately studied, and no reproducible theoretical model exists that has been verified through experiments.

The main contributions of our work are as follows:
We derive two equivalent electrical-circuit models for analyzing the channel gain, for wireless communication through the dry skin surface and inner tissue-layers, such as muscle and fat for the human arm. Our theoretical approach is validated with previously conducted experiments in [9] and [11] for on-skin cases. There is no published work yet for measurements on intra-tissue signal propagation.

For verifying the accuracy of the multi-tissue analysis, we construct a 3D model of the human arm using Ansys HFSS, a high-performance full-wave electromagnetic (EM) field simulator that uses Finite Element Method (FEM). The simulator design captures minute aspects of the signal propagation through the inner tissues. The results are also well aligned with the experimental findings. This allows the simulation to be used for initial analysis of future network designs for situations where intra-body testing is not immediately feasible.

We provide insights on suitable implant positions inside the body tissues, and the corresponding optimal transmission frequency ranges that provide the best performance. We also formulate closed form expressions for the noise that impacts correct signal reception, which is the first step towards the further work for estimating signal-to-noise ratios and selection of suitable modulation schemes.

The rest of the paper is organized as follows. We formulate the equivalent circuit models for determining the signal transmission gain in the human arm and an approximation of possible noise power in Section II. The finite element simulation model is described in Section III, with the validation of the models given in Section IV under a variety of tissue types, macro-micro sensor placements, electrode sizes, and sensor separation distances. Finally, Section V concludes the paper.

II. EQUIVALENT CIRCUIT MODEL OF HUMAN ARM

Our work on a theoretical model for the human tissue communication channel is motivated by the fact that in-vivo tissue experiments are not always possible, and commercially available phantoms do not accurately reflect the electrical propagation characteristics over a wide frequency range. These practical limitations pose challenges in determining the optimal signal frequency of operation, and the signal power levels. The model will determine the loss in signal for a given choice of input frequency, transmitter-receiver distance (D) and separation between their electrodes ($E_S$).

We propose two equivalent circuit models for human tissues based on the two-port [14] and lumped element [15] approaches. The two-port model represents individual tissues as black-boxes, which are constructed from the electrical properties of real tissues, configurable with tissue dimensions corresponding to a specific human subject. Though this model adapts easily for different parts of the body with varying tissue thicknesses, it is more accurate for signal transmissions and receptions within the same tissue layer. On the other hand, the lumped element model provides the flexibility to analyze the effect of sensor placements at different tissue depths. Thus using this latter model, it is possible to identify the optimal positions and operating conditions of implants. We describe these models next using the frequency dependent electrical properties of tissues defined as follows:
Living tissue composes both movable charges and movement restricted dipoles and hence can be characterized as an imperfect dielectric medium. Tissues consist of an array of conducting cells. When excited by an external electrical signal, each cell activates its neighbor, enabling signal propagation through different paths dictated by the cell structure and the frequency of operation. Low frequency signals cannot penetrate the high impedance cell membrane, and so it takes the circuitous path through extra-cellular fluid. As opposed to this, high frequency signals pass through intra-cellular fluid by penetrating the cell membrane. Thus, the cell membrane gives a capacitance effect allowing only high frequency components.

Using the frequency dependent electrical properties of live tissues [2], [3], a simple biological cell can be modeled as parallel combination of Resistance $R_{ext}$ (representing dissipation loss), and a capacitor $C_m$ (representing the charge holding ability), as given in fig.2(a). Such materials can be completely described in terms of conductivity ($\sigma$) and relative permittivity ($\epsilon$) or in terms of complex relative permittivity ($\epsilon^*$) using Debye or Cole-Cole model [2]–[4], [18] given by,

$$\epsilon^* = \epsilon' - j\epsilon''$$

where, $\epsilon'$ is the dielectric constant and $\epsilon''$ is the out of phase loss factor, expressed as,

$$\epsilon' = \epsilon_\infty + \frac{\epsilon_s - \epsilon_\infty}{1 + \omega^2\tau^2}$$

$$\epsilon'' = \frac{(\epsilon_s - \epsilon_\infty)\omega\tau}{1 + \omega^2\tau^2}$$

where $\epsilon_\infty$ and $\epsilon_s$ are dielectric constants at very high and very low frequencies, $\omega$ is the angular frequency measured as $2\pi \times \text{frequency}$ and $\tau$ is the dielectric relaxation time given by $X/R$. The dielectric loss tangent ($\tan \delta$) represents inherent dissipation of electromagnetic energy given by $\frac{\omega\epsilon'' + \sigma}{\omega\epsilon'}$. Using (2) and (3), the tissue admittance using
the parallel RC elements (fig. 2 (a)) can be calculated as,

\[ Y = \frac{1}{Z} = G + j\omega C = F_W K(\omega \epsilon'' + j\omega \epsilon') \]  

(4)

where \( Z \) is the impedance, \( G \) is the conductance, \( K \) is the ratio of area and length of tissue decided by the direction of impedance measurement as explained in section II-B and \( F_W \in (1, 4) \) is the correction factor accounting for variation in dielectric properties with respect to extra-cellular and intra-cellular water distributions that can be determined using bioelectrical impedance analysis (BIA) [19], [20]. Thickness of a tissue (especially fat tissue) of a person can be approximated using body mass index (BMI), or triceps skin fold thickness. The overall electrical behavior of tissue depends on the structure and distribution of cells, along with its extra-cellular water content that varies from person to person.

B. Two-Port Equivalent Circuit Model

Prior to considering a human body part, equivalent circuit of a single galvanic coupled tissue is formulated using three impedances. The impedance values are derived based on the three paths taken by an injected current marked as A,B and C in fig.2.b.

- Path A is the primary return path offering the direct impedance \( Z_D \), that channels the majority of current from terminal to ground electrodes in the transmitter. In this case, factor \( K \) given in (4) takes the form \((E_L \times T)/E_S\), where, \( E_L \) is the electrode length, \( T \) is tissue thickness and \( E_S \) is the terminal-reference electrodes separation.
- Path B serves as a pathway for a small portion of current directed towards the receiver electrodes through longitudinal impedance \( Z_L \), between transmitter and receiver electrodes. \( K \) of \( Z_L \) is calculated as \((E_L \times T)/D\), where, \( D \) is the transmitter-receiver separation.
- Path C is the electric current propagation path to adjacent tissue layer through transverse impedance \( Z_T \). To compute this impedance, \( K \) is substituted with \((E_L \times E_B)/T\), where, \( E_B \) is the electrode breadth.
Additionally, there is a constant electrode coupling impedance $Z_C$ at the interface of tissue and electrode. A load impedance $Z_{Load}$ is added across the receiver electrodes, with the dots representing the possibility of attaching $Z_C$ and $Z_{Load}$ to any single tissue under study. For developing a tractable model, we assume uniform transverse tissue depth $Z_D$, indicated by $\oplus$ in fig.2.b. This assumption would yield identical impedance between transmitter and receiver locations. Similarly, the impedance between terminal electrodes and the impedance between reference electrodes of transmitter and receiver, indicated by $\otimes$ in fig.2.b, are also assumed to be equal. These assumptions enable simple modeling of each tissue as a balanced network. The transmitter electrodes, attached on the tissue, form the in port, and the receiver electrodes form the out port.

We approximate the galvanic coupled human arm as cylindrical dielectric block as given in fig.3. The cylindrical model of 400 mm length has four concentric tissue layers - outer dry skin, fat, muscle and cortical bone (hard outer covering of bone) layers of thickness 1 mm, 9 mm, 25 mm and 20 mm respectively. The variable factors such as $T$, $D$, $E_S$, $E_L$ and $E_B$ are added as variables in impedance calculation.

A sample case of the transmitter coupling location on skin surface is shown in fig.3. In the following multi-layer discussion the superscript $i$ and $j$ denote a specific tissue layer, i.e., $i, j \in \{S, F, M, B\}$, with the substitutions of $S$ for skin, $F$ for fat, $M$ for muscle, and $B$ for bone. The single tissue impedance $Z_D$ and $Z_L$ in fig.2 become $Z_D^{i-1}$ and $Z_L^{i-1}$ and $Z_T$ takes the form $Z_T^{i-j}$, denoting path from layer $i$ to $j$. Additional pathways for multi-tissue propagation are also shown in the diagram. This concept is clarified further in fig. 4 and some sample cases are discussed next.

**Example 1:** When the transmitter and receiver nodes are both on the skin surface (S-S Channel for macro-sensors), the impedance of tissue layers beneath the skin are in series with the transverse impedance of the skin $Z_{S-F}^T$.

**Example 2:** When the sensors are moved to the muscle tissue (M-M Channel for micro-sensors), the impedance offered by the skin and fat tissues are now in parallel with that offered by the bone. The resulting impedance is now in series with muscle’s transverse impedance $Z_{T-B}^M$.

Similar cases can be constructed for other locations of the tissues. The benefit of this black-box approximation of 2-port model is a simple first-approximation for the voltages and currents that are likely to be observed within the given tissue layer during communication. We now represent the tissue in terms of its specific electrical properties defined by the $Z$-parameters, and represented in a $2 \times 2$ matrix of complex numbers. The relation between impedance, in-port ($V_1$), out-port ($V_1$) voltages, and currents ($I_1$ & $I_2$) is given by,

$$
\begin{bmatrix}
V_1 \\
V_2
\end{bmatrix} =
\begin{bmatrix}
Z_{11} & Z_{12} \\
Z_{21} & Z_{22}
\end{bmatrix}
\begin{bmatrix}
I_1 \\
I_2
\end{bmatrix}
$$  \hspace{1cm} (5)

The $Z$ parameter of each layer can be calculated as,

$$
[Z] =
\begin{bmatrix}
Z_{11} & Z_{12} \\
Z_{21} & Z_{22}
\end{bmatrix} =
\begin{bmatrix}
\frac{2Z_1Z_2}{Z_1 + 2Z_2} & \frac{Z_1Z_2}{Z_1 + 2Z_2} \\
\frac{Z_1Z_2}{Z_1 + 2Z_2} & \frac{Z_1Z_2 + Z_2^2}{Z_1 + 2Z_2}
\end{bmatrix}
$$  \hspace{1cm} (6)
where \( Z_1 = \left( \frac{2Z_L Z_T}{Z_T - Z_L} \right) \) and \( Z_2 = \left( \frac{Z_T Z_D}{Z_T + Z_D} \right) \).

We will now develop detailed analytical models for skin-skin and muscle-muscle signal propagation in this section.

1) Skin-skin model: Here, the transmitter and receiver sensor electrodes are positioned on the skin surface with coupling impedance \( Z_C \). Though the primary current flow is through skin, secondary parallel paths exist through fat, muscle and bone layers. In this case, the total skin transfer impedance, \( Z_{S-F} \) is calculated in series with fat, muscle and bone as follows. Bone is assumed to be surrounded by muscle and therefore, the input impedance of bone is included in transverse impedance of muscle as,

\[
Z_{T-B} = Z_{T-B} + Z_{in}^B
\]  

(7)

where input impedance of bone \( Z_{in}^B \) is given by

\[
Z_{in}^B = Z_{11}^B - \frac{Z_{12}^B Z_{21}^B}{Z_{22}^B + Z_L}
\]  

(8)

Consequently, the equivalent circuit model is reduced by one layer as the bone’s impedance is included in muscle’s impedance. In the same way, Z parameters of Fat [\( Z_F \)] is determined from its direct, longitudinal and transverse impedance where the transverse impedance \( Z_{T-F} \) includes muscle’s input impedance, \( Z_{in}^M \).

\[
Z_{T-F} = Z_{T-F}^F + Z_{in}^M
\]  

(9)

Input impedance of fat is then included in the transverse impedance of skin layer as,

\[
Z_{T-F}^S = Z_{T-F}^S + Z_F
\]  

(10)

Input impedance of muscle and fat are calculated in same way as that of bone as given in (8). As a result, the human arm equivalent circuit given in fig. 3 for multi-layer tissues is reduced to single tissue impedance equivalent
Fig. 5. Lumped Element Circuit model for human arm

to fig.2. The voltage gain can be calculated from the Z parameter of total equivalent impedance \([Z]\) (or \([Z_S]\)) as,

\[
G = \frac{V_{out}}{V_{in}} = \frac{Z_{21}Z_{Load}}{(Z_{11} + Z_E)(Z_{22} + Z_{Load}) - Z_{12}Z_{21}}
\]  

(11)

2) Muscle-muscle model: To study the channel response at the muscle layer, the transmitter and receiver sensor electrodes are moved inside muscle tissue, in which the dominant path of the current also lies. In this case, the secondary current flows through fat and skin layers and also into the bone. The arrangements of the respective blocks of the 2-port model are shown in fig.4. Here skin input impedance \(Z_{in}^S\) is included in the transverse impedance \(Z_{F-M}^T\) of the fat layer. Now \(Z_{F-M}^T\) is in parallel with the transverse impedance of bone \(Z_{B-B}^T\), and therefore, their admittance parameters is aggregated as:

\[
\frac{1}{Z_{F-M}^{T\text{sum}}} = \frac{1}{Z_{F-M}^T} + \frac{1}{Z_{B-B}^T} 
\]

(12)

The resulting impedance is included in the transverse impedance of muscle, \(Z_{M-B}^T\). As all other layers are reduced to \(Z_{M-B}^T\) other than muscle, the gain at the muscle layer is calculated similarly as (11).

Both the skin and muscle signal propagation models are verified later in Section IV. The 2-port model assumes a black-box behavior and makes it easier to add, remove or modify tissue layers within the model. However, it cannot account for the changing impedance when the transmitter and receiver sensor electrodes are moved within the tissue (say, from 10 mm to 15 mm depth inside muscle). In the next section, we explain a second approach using Kirchoff’s current law (KCL) that provides the ability to move the transmitter and receiver electrodes anywhere
inside tissues, and also consider the impact of inter-tissue sensor placement.

C. Lumped Element Equivalent Circuit Model

Under 100 MHz, the dimensions of human body and implants are small compared to the signal wavelength, and hence, we can undertake the analysis using lumped elements. The impedance calculations corresponding to the three paths of current (shown in fig.3) are performed similar to the 2-port model.

The circuit in fig.5 is used to model the flow of current coupled by electrodes and solved using Kirchhoff’s Current Law (KCL). The circuit shown has 4 tissue layers and 15 nodes with 15 tensions and 15 equations, one for each node as,

$$\mathbf{M}_{Ad} \mathbf{v}^T = \mathbf{i}^T$$  \hspace{1cm} (13)

where $\mathbf{M}_{Ad}$ is the admittance matrix for each node, $\mathbf{v}^T$ is the vector with tensions that needs to be found, and $\mathbf{i}^T$ is the vector with the sum of currents through each node. A node equation can be formed in matrix form as follows.

$$\mathbf{M}_{Ad} = \begin{bmatrix}
\sum_{i=1}^{n} \frac{1}{Z_{1i}} & -\frac{1}{Z_{12}} & \cdots & -\frac{1}{Z_{1n}} \\
-\frac{1}{Z_{21}} & \sum_{i=1}^{n} \frac{1}{Z_{2i}} & \cdots & \frac{1}{Z_{2n}} \\
\vdots & \vdots & \ddots & \vdots \\
-\frac{1}{Z_{n1}} & -\frac{1}{Z_{n2}} & \cdots & \sum_{i=1}^{n} \frac{1}{Z_{ni}}
\end{bmatrix}$$  \hspace{1cm} (14)

where $Z_{nm}$ is the impedance between node $n$ and node $m$. From this node equation, the voltage vector $\mathbf{v}$ and current vector $\mathbf{i}$ representing the sum of currents entering or leaving node can be derived as

$$\mathbf{v} = \begin{pmatrix}
V_1 \\
V_2 \\
\vdots \\
V_n
\end{pmatrix} \hspace{0.5cm} \& \hspace{0.5cm} \mathbf{i} = \begin{pmatrix}
\frac{V_{IN}}{R_s} \\
0 \\
\vdots \\
0
\end{pmatrix}$$

where $V_n$ is the voltage at node $n$. The position of $V_{IN}/R_s$ depends on the position of the source. The transfer function from the circuit in fig.5 is calculated using

$$G(w, E_L, D, E_S, [T]) = 20 \log_{10} \left| \frac{V_O(w, E_L, D, E_S, [T])}{V_I} \right|$$  \hspace{1cm} (15)

where $[T]$ is the vector of tissue thicknesses for skin, fat, muscle and bone, $V_O$ is the potential difference observed in load obtained from (13) and $V_I$ is the source voltage. We tracked the phase shift information using the following equation.

$$\text{Phase} = \arctan \left( \frac{\text{Im}(V_O)}{\text{Re}(V_O)} \right)$$  \hspace{1cm} (16)

As in the case of the 2-port model, the models based on lumped element circuits for S-S and M-M channels (as well as other inter-tissue propagation cases such as S-M/M-S) are verified in Section IV. We use the lumped element model for the rest of the analysis in this paper, as it is more expressive of the impact of tissue depth, which
becomes a key factor in the design of implanted sensors.

D. Noise estimation and signal to noise ratio

To fully determine the ability of the receiver to decode the signal and determine the achievable rate, the estimation of noise is of critical importance. Once this noise level is known, the signal to noise ratio (SNR) can be computed, and the impact of various modulation schemes can be studied. To quantify the noise level, we model the noise by approximating the power spectral densities (p.s.d) of thermal noise, electrode coupling noise and RF radiation interference as described below.

**Thermal Noise** \(N_T(f):\) The thermal noise depends mainly on frequency and temperature and can be calculated as:

\[
N_T(f) = \sqrt{4KT} \frac{R}{W/\sqrt{Hz}}
\]

where \(T\) is the absolute temperature in Kelvin, \(K\) is Boltzmann constant, and \(R\) is electrode and tissue resistance in ohms.

**Electrode coupling noise** \(N_E(f):\) This noise occurs at the interface where the electrode is attached to the tissue. The skin-electrode interface noise can be related to the real part of the skin-electrode impedance, and is equivalent to the thermal noise at high frequencies [28]. We use the noise p.s.d of surface electrodes as approximated in [27], [28] as:

\[
N_E(f) = \frac{1}{f^\alpha}, \quad 1.5 < \alpha < 2.0
\]

where \(\alpha\) is a correction factor that depends on the gel type and skin properties.

**RF Radiation interference** \(I_o:\) RF radiation from sources such as TV and radio broadcast signals, transmissions in the so called *lost band* (160 - 190 KHz), non-directional radio beacons (NDBs) (190 - 435 kHz), amateur radio (135.7 - 137.8 KHz) and top band radio (1.8 MHz - 2 MHz) that are in 100 KHz to 10 MHz range might be a potential source of interference. We approximate these interference sources as Additive White Gaussian Noise, by assuming that the contributions from each of these sources are independently and identically distributed (i.i.d) Gaussian random variable \(N(0, \varphi^2)\), with zero-mean and standard deviation \(\varphi\).

The channel’s Signal to Noise Ratio (SNR) and maximum capacity using Shannon - Hartley theorem can be calculated as follows.

\[
SNR = \frac{P_t \cdot G(w, E_L, D, E_S, [T])}{(N_T + N_E + I_o)\Delta f}
\]

\[
C = \Delta f \log_2(1 + SNR) \, [\text{bits/s}]
\]

where \(\Delta f\) is the bandwidth. Fig.9 shows the channel capacities that are theoretically achievable with different \(\varphi\) values. This estimation of noise is used to determine the standard bit error rate (BER) for a given modulation type. We provide results for BPSK and QPSK modulation specifically in table.I.
III. SIMULATION OF SIGNAL PROPAGATION IN TISSUE USING FINITE ELEMENT METHOD

In this section, we describe our evaluations using the Ansys HFSS [21], which allows us to perform full-wave electromagnetic (EM) simulations for arbitrary 3-D models. It allows detailed computational analysis of field distribution at various locations inside the human tissues using Finite Element Methods (FEM), and is especially useful when experimental results are not easily obtained for intra-body channels.

We model the human arm with dimensions as described in Section II. A pair of copper cuboids of dimension $10 \times 10 \times 1 \text{mm}^3$ are used as the terminal and reference electrodes that are connected by a complex impedance defined lumped port. The source current of $1 \text{mA}$ is set at the lumped port (input). Around 1 foot distance around the human arm model, we emulate a boundary as an open electrical circuit. The frequency dependent electrical properties of dielectric tissue blocks are configured using (1)-(3) for the frequency range of 10 KHz to 10 MHz.

HFSS transforms the 3-D tissue model into a mesh of tetrahedron structures, with a high density of mesh points at critical positions like the electrode-tissue interface (fig.6(left)). To estimate the field strength across these tetrahedrons, complex EM field values at each vertex of tetrahedron is computed using Maxwell’s partial differential equations. The normal electric field ($E$) on skin surface is measured as surface integral over an area equivalent to the surface area of a receiving electrode and the magnetic field ($H$) is measured as surface integral of its tangential component. The current through surface $S$ at distance $l$ from the source can be obtained from Ampere’s law as $I_{\perp S}$ at $l = \oint H.dl$.

For emulating the signal received at the implanted micro-sensor, we move the transmitter electrodes and port deep into the tissue and also draw a line approximation for the receiver sensor lead. The $H$ field strength measured tangential to this receiver line are used to calculate the current induced in it as given through Ampere’s law above.

The gain through the tissues can be calculated as follows.

$$ G_E(dB) = 20 \log_{10} \left( \frac{E_{\text{Detector}}}{E_{\text{Coupler}}} \right) $$

(21)

The simulation is repeated for different $E_S$ (distance between the terminal and reference), and D (different distances between the transmitter and receiver) for varying [T] (thickness of tissues) at frequencies ranging from 10KHz to
10MHz.

IV. CHANNEL GAIN & MODEL VALIDATION

This section validates the theoretical models using the 2-port model (Section II-B) and the lumped parameter model (Section II-C) with the simulator design described in Section III, as well as prior experimental studies in the literature. For the experimental results, we use the findings described in the existing works [9] and [11]. We conduct the following evaluations: (i) variation of gain with frequency, (ii) phase shift of the signal with frequency, (iii) impact of frequency of the signal on energy dissipation.

The simulation environment adheres to the International Commission on Non-Ionizing Radiation Protection (ICNIRP) guidelines [17] that limits the current density through the human body to $5\text{mA/m}^2$ for electrical signals in the frequency range of 100 KHz through 60 MHz. The energy absorbed by the tissue is proportional to the conductivity of the medium. At lower frequencies such as 100 KHz, conductivity and therefore, the absorption is low and it results in no impact on live tissue. The ability of tissue to carry electrical signals as a channel is measured in terms of channel gain. The channel gain obtained on the S-S channel with D being 100 mm and 200 mm using the two models, i.e., the 2-port model (11), lumped parameter model (15), and our simulator (21) are presented in fig. 7. We see a good agreement among the three plots and with prior experimental results from literature with a maximum difference of 4 dB in gain among all the results.

We next present results for the signal gain when the sensors are at the micro-scale within the tissue, all the other parameters being unchanged. As there are no published experimental data on the signal gain with the transmitter and receiver placed within the tissue, we limit the studies to the two theoretical models and the same FEM simulation that accurately matched the on-skin experimental data. The new results for the propagation through muscle tissue only (i.e., the M-M channel) are given in fig. 8, which verifies the accuracy of the model through simulation.
A. Variation of gain with frequency and distance

In this section, we vary both the distance between the transceiver and receiver, as well as the frequency of the signal, and study their cumulative impact on the signal gain. Our studies for two different distances, 100 mm and 200 mm for the S-S channel are presented in fig. 7. We observe that the channel gain increases with frequency from 10 KHz to 5 MHz, and especially, in the range 100 KHz to 1 MHz, the overall increase in gain is about 5 dB. Within the muscle tissue, as we vary the transmission frequency between 100 KHz to 1 MHz, the difference in gain is less than 2 dB. Thus, M-M channel is more robust to the choice of the frequency. While the gain varies in a similar trend in the muscle tissue as is seen for the skin, interestingly, the comparative lower attenuation in muscle ($\approx 22 - 24$ dB) than that on skin ($\approx 34 - 42$ dB) is because muscle traps most of the energy. Thus, when the coupling electrodes are placed in the muscle, very small amount of energy is dissipated into the neighboring tissues.

Transmitted signals suffer a natural attenuation with distance as the longitudinal impedance, $Z_L$ gets larger. For an increase in D of 100 mm on S-S channel, the signal gain drops by a factor of 10. In the M-M channel,
there is about 7 dB decrease in gain for increase of 100 mm of D. The S-M and M-S channels have the similar relationship with frequency and distance, with channel gain lying between S-S and M-M channels. Interestingly, with the transmitter on skin, the gain at the receiver placed deep within the muscle (S-M channel) is about 10 dB more than that of the S-S channel. However, below 50 mm of D, the S-S channel dominates at all frequencies, which makes on-surface communication preferable at short distance. Table I summarizes the channel capacity and possible error rates achievable through different channels at 100 mm and 200 mm with two extreme values of noise variances as \(1 \mu W/\sqrt{Hz}\) and \(1 nW/\sqrt{Hz}\) using BPSK and QPSK modulation schemes.

B. Phase shift of the signal with frequency

We next study the impact of phase change on the transmitted signal using (16), when the transmitter and receiver are placed on/within a single layer, i.e., the skin (S-S) and muscle (M-M), and also when the signal moves from one layer to the other, i.e., muscle to skin (M-S) and skin to muscle (S-M). Fig. 10 shows the results of the phase shift when the signal frequency varies in the range of 100 KHz – 1 MHz. We observe that the phase shift on the S-S channel varies from 20 to 30 degrees, whereas for M-M, there is less than 5 degrees of phase shift. Thus, the muscle tissue serves as a better channel, offering less phase shift to the original signal. The phase change for the tissue cross-over cases S-M and M-S is in-between these two cases, with a deviation from 8 to 15 degrees. This phase information is critical for training the receiver with the corresponding phase shifts depending on the frequency used for the transmission.

C. Impact of operating frequency

To identify the ideal range of the transmission frequency, we consider two factors: (i) frequency of the signals naturally generated by the human body, and (ii) signal loss caused by dissipation for a given frequency within the tissue. The electrical signals within the human body comprise of neural impulses, ECG, and EEG signals that operate at a frequency lower than 1 KHz, and therefore, we avoid the frequencies \(\leq 1\) KHz for intra-body communication.
Also, as the channel characteristics are frequency dependent, we need to identify the ideal operating frequency that reduces signal loss.

The signals transmitted into the tissue results into two current components, i.e., the conduction current and displacement current. At lower frequencies, the conduction current that is caused by the movement of charges is high. This enables energy concentration inside the tissue, resulting in higher signal strength at the receiver end. At higher frequencies, due to increase in the capacitance effect, there is less movement of electrons with increasing electric field strength. This ultimately results in the signal dissipating from the body into the surrounding region, possibly causing interference externally, as well as limiting the energy incident on the receiver electrode. This situation is demonstrated in fig.11 using field distributions observed using the FEM model, as explained below.

Above 1 MHz, the conductivity remains constant and therefore the conduction current also remains fixed. However, due to capacitance effect at higher frequencies, the displacement current grows larger with frequency above 1 MHz resulting in the field distribution spreading out of the human body to a distance of few centimeters. For instance, at 100 KHz, (fig.11) the H field in the surrounding the body is in the order of $\mu A/m$, extending to less than 50 mm at the exterior. On the other hand, at 10 MHz, the H field surrounding the body is higher by two orders of magnitude, extending to about 3 feet away from the body. The signal spreading out of the body is considered wasted, as it cannot reliably be detected at the embedded receiver. Thus, the signal loss is minimized as long as the operating frequency is restricted in such a way that the conduction current dominates the displacement current. This is true when the relationship $\sigma \omega \epsilon' > 1$ holds, i.e., when we limit the frequency lower than 2 MHz. Thus, to ensure that the dissipation loss is at minimum, the maximum frequency of operation is set at 1 MHz.

D. Effect of Tissue Thickness on Signal Gain for Intra-body Sensor Placement

In this section, we investigate the impact of fat and muscle tissue thickness on the signal gain. When a signal propagates along a tissue, it attempts penetrating into neighboring layers with less impedance. Each point of penetration becomes a signal source with respect to the neighboring tissue. As sensors are placed either on the skin (simple placement with external access) or in the muscle (best propagation characteristics), the intermediate

Fig. 11. Magnetic Field spreading out of body at 100 KHz (left most), 500 KHz, 1 MHz and 10 MHz (right most)
fat tissue and its thickness play a crucial role when the signal transcends the tissue boundaries. In general, fat acts as a barrier between skin and muscle tissues, allowing either tissue to either retain the energy (in case of a thick fat layer) or allowing more current to pass through (in case of thin fat layer).

For the channel gain results given in fig.7 and fig.8, we considered an average value of arm fat thickness as 9 mm. The transmitter and receiver are positioned on the skin with $D = 200\ mm$ (i.e., S-S channel) for varying fat thickness from -90% to +90% in the frequency range of 100 KHz and 1 MHz. Then, the receiver alone is moved inside the muscle (i.e., S-M channel) with the same longitudinal distance as 200 mm. The channel gain observed using these two set-up is shown in fig.12.

- **Fat thickness:** The variation in fat thickness offers a signal gain difference of around 15 dB in each layer except for M-M channel. The M-M channel always out-performs other channels, and is also not affected by the variation in fat thickness. The M-S and S-M channels exhibit better gain for less fat thickness, whereas the S-S channel performs better when the fat thickness is above 11.5 mm. For average thickness of fat, S-S channel covers less distance because of low gain, when compared with the other cases. We can conclude that for thick fat layer, receiver should be positioned in the same tissue layer as the transmitter for better channel gain. As signal leakages between the S-M and M-S channels are non-negligible for any fat thickness, simultaneous communication on the

---

**TABLE I**

<table>
<thead>
<tr>
<th>D (mm)</th>
<th>Freq. (KHz)</th>
<th>Channel Capacity (bps)</th>
<th>BER (BPSK)</th>
<th>BER (QPSK)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>S-S</td>
<td>S-M/M-S</td>
<td>M-M</td>
</tr>
<tr>
<td>100</td>
<td>1kHz</td>
<td>2.9 × 10^4</td>
<td>1 × 10^5</td>
<td>3.2 × 10^5</td>
</tr>
<tr>
<td>1000</td>
<td>1kHz</td>
<td>3.5 × 10^4</td>
<td>1.7 × 10^5</td>
<td>4.2 × 10^5</td>
</tr>
<tr>
<td>200</td>
<td>1kHz</td>
<td>3 × 10^3</td>
<td>3 × 10^4</td>
<td>1.7 × 10^5</td>
</tr>
<tr>
<td>1000</td>
<td>1kHz</td>
<td>4 × 10^3</td>
<td>4 × 10^4</td>
<td>1.9 × 10^5</td>
</tr>
</tbody>
</table>

| 100    | 1MHz        | 2.5 × 10^5 | 5 × 10^5 | 7.7 × 10^5 | 7 × 10^{-4} | 1 × 10^{-17} | 1 × 10^{-132} | 0.01 | 1 × 10^{-10} | 1 × 10^{-72} |
| 1000   | 1MHz        | 3 × 10^5 | 6 × 10^5 | 9 × 10^5 | 2 × 10^{-6} | 1 × 10^{-33} | 1 × 10^{-160} | 8 × 10^{-4} | 1 × 10^{-18} | 1 × 10^{-130} |
| 200    | 1MHz        | 6 × 10^4 | 3 × 10^5 | 5.8 × 10^5 | 0.3 | 1 × 10^{-15} | 1 × 10^{-35} | 0.7 | 1 × 10^{-12} | 1 × 10^{-17} |
| 1000   | 1MHz        | 9 × 10^4 | 3.5 × 10^5 | 6.8 × 10^5 | 0.09 | 1 × 10^{-17} | 1 × 10^{-45} | 0.2 | 1 × 10^{-14} | 1 × 10^{-24} |

---

Fig. 12. Gain in 4 channels vs. % of Variation in Fat Width
skin and within the muscle cannot coexist at the same frequency. Thus for multiple pair of sensors placed on the skin as well as between implants within the muscle to be active, a multi-access scheme is required. For covering longer distances, and if the values of BMI indicate thick fat layer, the M-M and S-M/M-S channels are preferable.

- **Muscle thickness:** We undertake a similar study for varying muscle thickness (fig.13). Interestingly, the M-M channel gain increases exponentially with muscle thickness, as it is able to trap a large amount of signal. The S-S, S-M or M-S channels are not affected significantly by the variation in muscle thickness. In addition to tissue thickness, there are other prominent factors that influence gain, such as the electrode dimensions and separation between terminal and reference electrodes ($E_S$) in the transmitter and receiver.

- **Electrodes Separation:** Fig. 14 illustrates the effect of varying $E_S$ of the transmitter and the receiver for the S-S sensors. With the increase in $E_S$, the gain increases, which is more prominent with thin fat tissue. Therefore, moving the electrodes far apart, such as for the separation achieved by positioning one on the top surface and the other one on the bottom surface of the arm, and coupled with the presence of a thin fat tissue, the gain dramatically increases to a maximum of 25 dB. On the contrary, $E_S$ increase may contribute to about 6 dB increase in gain for fat thickness that is above average level.

  We observe similar trends when the separation distance is varied within muscle (i.e., the M-M case). For larger $E_S$, the gain is higher, which becomes more significant for thick fat layers. By parting electrodes more than 10 mm, the gain increases from 5 to 10 dB for average fat width.

- **Electrode dimensions:** The electrode size also has same effect as that of $E_S$. Larger electrode dimensions offer larger gain. For instance, an increase of 10 mm in $E_L$ and $E_B$ of electrode brings in 8 dB of improvement in gain. However, larger on-skin macro or implanted micro sensors may cause discomfort. Thus, a compromise between electrode size and gain can help decide the transmitter - receiver distance, the next hop sensor and the need for relay nodes and its best possible location.

![Graph showing the variation in muscle thickness for different channels and frequencies.](image)

**Fig. 13.** Gain in 4 channels vs % of Variation in Muscle Width
V. CONCLUSIONS

In this paper, we devised and verified two different electrical circuit equivalent models for characterizing the physical layer of the body communication channel based on galvanic coupling. We conducted extensive studies regarding the gain and phase-change in the transmitted signal under varying operating frequencies, tissue dimensions, sensor placements, electrode separation distances and dimensions, among others, to comprehensively characterize the body channel, while respecting permissible current limits. We found that a maximum of 15 dB in channel gain could result from variation in tissue properties from person to person. We identified the optimal frequency to be between 100 KHz to 1 MHz for both on skin and in muscle channels, and determined that placing both the sender and receiver sensors within the muscle offered better channel propagation characteristics, as opposed to on the skin. The results on SNR studies and the achievable capacity for BPSK and QPSK modulation will help in selecting sensor locations within the body and well as determining number of relay nodes to meet the bandwidth needs of the application. We will investigate higher layer protocol design as future work, using the physical characteristics of the body channel derived in this paper.

ACKNOWLEDGEMENTS

This material is based on work supported by the U.S. National Science Foundation under Grant No. CNS-1136027. The authors are grateful for the helpful discussions and inputs provided by Deniz Erdogmus from Northeastern University and Taskin Padir from Worcester Polytechnic Institute.

REFERENCES


