Ingestible Electronics Without Batteries: Wireless Power and Communication for Gastroresident Devices

by

Abubakar Abid

Submitted to the Department of Electrical Engineering and Computer Science
in partial fulfillment of the requirements for the degree of Master of Engineering in Electrical Engineering and Computer Science

at the

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Abstract

In this thesis, I introduce a novel ingestible electronic device designed to reside inside the stomach for weeks or longer with both wireless communication and wireless powering capabilities. The device is powered by a transmitter outside of the body via wireless power transfer through 5 to 6 cm of tissue, while keeping under specific absorption of radiation limits. Electromagnetic theory and microwave simulations identified the optimal region of operation for transmitting power through tissue as around 1 GHz. Small loop antennas fabricated to fit onto a circuit board the size of a pill capsule exhibited power transfer efficiencies of around -45 dB when tested ex vivo in pig stomach tissue. Choice of electronic components for rectification, sensing, and wireless communication are also discussed, as well as electrical and material characterization of an encapsulated device, to show that an end-to-end wireless ingestible electronic device is feasible for gastroresident applications.

Thesis Supervisor: Dr. Giovanni Traverso
Title: Research Fellow
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Chapter 1

Introduction

This thesis introduces the first end-to-end ingestible electronic device that is designed to reside inside the stomach for weeks or longer, and is entirely wireless, in both power delivery and communication.

Recent technological developments have made it possible to envision electronics that can be swallowed and retained inside of the stomach, devices that I refer to as *gastroresident ingestible electronics* (GIEs). Applications of these devices are numerous, including continuous monitoring of vital signs, sensing micronutrients or toxins in the gastrointestinal tract, and even actively delivering therapeutics directly into the stomach.

In this thesis, I show that such devices can be more than envisioned; they can be designed, fabricated, and tested in environments that mimic the gastric environment. Furthermore, they can be constructed without batteries, which provides further advantages in miniaturization, biocompatibility, and retention lifetime.

This work is part of the larger framework of implantable and ingestible medical devices that enable continuous monitoring of the body’s physiological and pathophysiological signals. Just as neural probes and implantable pacemakers have allowed clinicians to probe with unprecedented precision and respond real-time to events in the cardiac and nervous systems, GIEs have the potential to improve diagnosis and treatment of diseases in the gastrointestinal
system and perhaps other parts of the body as well.

1.1 Ingestible Electronics

The history of ingestible electronics starts remarkably early in the 1950s, when “radio pills” consisting of electronic components fitted inside of gel capsules were developed to monitor physiological signals in the gastrointestinal tract. These circuits generally consisted of oscillators and amplifiers made from only discrete transistors and passive components. Variations in capacitance and inductance due to changes in pressure, temperature, or pH resulted in variations in transmission frequency, which could be detected by an external receiver [6].

Over time, tireless improvement in the semiconductor industry drove the miniaturization of transistors in accordance with Moore’s Law [7], and it became possible to put more complex sensors and entire integrated circuits inside of pills. Pills that combined different sensing modalities in one pill soon followed, as did pills that could perform wireless video capsule endoscopy of the entire gastrointestinal tract [8]. Successful commercialization of these devices spurred further interest in developing new applications for ingestible electronics. In 2012, Proteus Digital Health received FDA approval for a small chip to be integrated within drug capsules to notify caretakers when patients had taken their medicine [9].

Despite the varied landscape of ingestible electronics in literature and on the market, none are designed to stay in the body beyond the regular gastric transit time, roughly two days – none are *gastroresident* devices.

1.2 Gastroresident Devices

Drug delivery researchers have long recognized that in spite of the convenience and safety of orally-administered medication, short gastric transit times reduce the efficacy of ingested drugs [10]. Since ingested substances generally make their way through the entire gastric tract in less than two days, drug release is far from complete by the time the pill exits the
body. Thus, over the last few decades, several different gastroretentive drug delivery systems (GRDDs) have been developed to stay in the stomach for weeks or months [10].

A variety of mechanisms have been exploited to increase gastric retention time. One such method relies on decreased or increased density of GRDDs. Low-density gel-forming cellulose or matrix-forming polymers can float in the stomach above gastric contents [11]. A similar effect can be achieved in effervescent systems that usually contain liquids that gasify in body temperature. Conversely, a high-density capsule that is ingested can sink through gastric contents until it sinks into the folds of the stomach wall, and be retained for an extended period of time as well [11].

Another kind of GRDD relies on swelling or expanding to achieve retention inside of the stomach. In such systems, the capsule contains polymer that expands in the stomach, either because it absorbs water, or because it unfolds once the gel capsule has dissolved [11]. If it becomes too large to pass through the pylorus, connecting the stomach to the small intestine, it will stay in the stomach for a long period of time.

GRDDs, in general but particularly those that rely on expansion, pose several safety risks that must be navigated in clinical applications. If the gastroretentive device blocks the entire opening of the pylorus, it can act as a plug and prevent gastric contents (e.g. food) from leaving the stomach altogether. In addition, if the device enters the pylorus in a swollen state, it can damage the narrow walls of the small intestine. Recent work from the Langer Lab has addressed both of those concerns through the development of an enteric polymer that breaks down in the neutral pH of the small intestine, but retains a loop shape (or other geometry) inside the stomach [1]. The geometry of the polymer can be designed to allow passageways for food and contents of the stomach to pass even when the polymer is in the stomach. The device can fit into a standard 000 pill capsule, as illustrated in Figure XX, and such devices can be retained for upwards of five days in the stomach [1].
The development of safe gastroretentive platforms provides an opportunity to fabricate electronic devices that can be embedded onto these platforms and be retained in the stomach as well. Such devices have numerous applications. Continuous monitoring of vital signs including core body temperature and cardiac and respiratory rhythms [12] would allow physiological status monitoring for people in high-risk situations. Biochemical sensors embedded in the stomach could measure ingested micronutrients and/or toxins, allowing real-time feedback for patients seeking to adopt certain diets. Ingestible electronics loaded with drugs could enable “smart” and precisely-controlled drug delivery, in response to specific physiological events.

Towards these ends, I worked with a modified version of the drug delivery platform developed in the Langer Lab, in the shape of a three-legged star, as shown in Figure 1-2. The star shape retained the flexibility to fit into a standard 000 gel capsule, while expanding to become larger than the opening of the pylorus upon the dissolution of the gel capsule in the stomach. The thicker leg of the star provided space for a small circuit board, that could house a microcontroller, sensors, antennas, and a battery, if desired. Functionally, this star provided us a means to deliver an ingestible electronic device to the stomach, where it could be retained for an extended period of time.

1.3 Gastroresident Ingestible Electronics (GIEs)

Figure 1-1: A gastroretentive device made of an enteric elastomer that fits inside a 000 capsule, adapted from [1]
1.4 Wireless Power for GIEs

GIEs are designed to reside inside the stomach for months, if not longer, thus requiring an energy source that can power these devices for a sustained period of time. For most GIE applications, particularly those that require sensing and wireless telemetry, or active drug delivery, including a battery severely limits the lifetime of the device, making gastric residence fruitless. Furthermore, embedding a battery also poses challenges to biocompatibility [13] and increases the physical size of the GIE.

To overcome these challenges, I investigated the possibility of powering the electronic device from outside of the body, specifically by using wireless power transfer to sustain the GIE indefinitely. Wireless power transfer comes with its own set of challenges, including optimizing the antennas and operating frequency given the tight size constraints of the GIE and minimizing absorption of radiation in the tissue. Solving these problems is my primary novel contribution through this thesis, as I was able to identify, fabricate, and test a wireless power transfer system that provided enough energy to power a GIE for sensing and telemetry applications.

We begin in chapter 2 by considering the problem of transmission of electromagnetic energy through tissue from the perspective of frequency and mode of propagation. Chapter
3 investigates the choice and characterization of antennas used in propagation. Rectification of RF power to DC is considered in Chapter 4, and detailed descriptions of the design of the receiver circuit board is discussed in Chapter 5. Characterization and testing is discussed in Chapter 6. Chapter 7 concludes and presents next steps.
Chapter 2

Wireless Power Transfer

Nikola Tesla first demonstrated wireless power transfer around the turn of the 20th century when he used inductively coupled coils to power incandescent bulbs from a distance [14]. The principle behind his experiment continues to be the basis of near-field wireless power transfer, in which the power is transferred over a distance, \( d \), much smaller than the wavelength, \( \lambda \), of the electromagnetic waves involved. In contrast, a different mode of wireless power transfer is far-field power transfer, in which power is sent over large distances relative to the wavelength (\( d \gg \lambda \)); radio broadcast stations typically broadcast electromagnetic radiation in the far field, for example, although for the purpose of transmitting signal, not power. Between the two regimes lies mid-field coupling, in which the electromagnetic wavelength and distance of separation between the transmitter and receiver are comparable. Because the three kinds of coupling require different hardware, I studied and assessed each of them before proceeding to the design of the wireless power transfer network.

2.1 Coupling Regimes

In this section, I discuss the three broad regimes of wireless power transfer in more depth, and assess which of the three regimes of operation is most suited to wirelessly powering GIEs.
2.1.1 Near-Field Coupling ($d \ll \lambda$)

Near-field coupling is used in a variety of medical and non-medical devices for wireless powering over short distances, such as Qi cellphone charging pads [15], certain pacemakers, and neural probes that are powered from outside of the skull [16]. In medical applications, near-field power transfer is based on the coupling of evanescent electromagnetic waves between a transmitter circuit outside of the body, and a receiver inside of the body. While the coupling can be either capacitive (electric field-based) or inductive (magnetic field-based), the latter is chosen for most medical applications because the magnetic field can penetrate into tissue without adverse effects [16].

In Fig. 2-1 (left), an inductively-coupled wireless powering scheme is shown. When an alternating current passes through the transmitting coil, an alternating magnetic field is generated. Part of this oscillating field passes through the receive coil, generating an alternating voltage that can be rectified and used to drive a load. Because the coupling depends on overlapping reactive waves, which do not radiate, this method is only effective if the distance over which power is being transferred is smaller than the size of the coils [15].

![Figure 2-1: Circuit implementations of inductive power transfer (left) and resonant inductive power transfer (right)](image)

To improve the efficiency of this circuit, resonant inductive coupling can be created by including capacitive elements to resonate both the transmit circuitry and receive circuitry at the frequency, $\omega_0$, at which the current is alternating through the transmit circuit. This allows the transmitter to “store” the energy in the transmit circuitry over multiple cycles until it is delivered to the receiver (generally this corresponds to operating at a frequency...
Even the efficiency of resonant inductive coupling degrades quickly with increasing distance. The efficiency, \( \eta \), is derived in [17], and can be written for our system as:

\[
\eta = \frac{(kQ)^2}{(kQ)^2 + 2(1 + \sqrt{1 + (kQ)^2})}
\]

(2.1)

where \( Q \) is the quality factor of each coil and \( k \) is the coupling factor between the two coils. This expression makes use of the assumptions that (1) the coils resonate at the same frequency, (2) the coupling factor is small \((k^2 \ll 1)\), and (3) the load resistance has been optimized for maximum efficiency. In reality, the efficiency will be lower, but this expression provides an upper bound.

The quality factor for an inductor with inductance \( L \) and series resistance \( R \) at its resonant frequency, \( \omega_0 \), is defined [17] as:

\[
Q = \frac{L\omega_0}{R}
\]

(2.2)

The coupling coefficient between the two coils depends on the radius of the coils and the distance between them. For two single-turn coils, it can be approximated [18] as:

\[
k = \frac{1}{[1 + (\sqrt{2}d/r)^{3/2}]^{3/2}}
\]

(2.3)

where \( d \) is the distance between the coils and \( r \) is the radius of the coils.

Because the internal diameter of the 000 capsule limits the board width to about 8 mm, the largest coil size that can be printed is about 6.5 mm wide (the margin is due to fabrication process limits). We can compute the theoretical maximum near-field efficiency of two 6.5 mm coils by using (2.1). If we assume optimistically [19] that our printed coils have a \( Q \) of up to 70, we see the efficiency curve as a function of distance and \( Q \) in Fig. 2-2.

Within a couple of centimeters, near-field transfer is very efficient - however, the efficiency
Near-Field Power Transfer Between 3.5-mm Coils

Figure 2-2: The near-field wireless power transfer efficiency, \( \eta \), between two 3.5-mm coils as a function of distance and \( Q \)

quickly degrades as distance between the coils increases. Even using coils with a \( Q \) of 70, the efficiency at 6 cm is -50 dB, or 0.001%. In addition, near-field coupling suffers from one qualitative limitation, and that it has little directivity [16], meaning that it does not allow focusing of power transfer in specific directions, as an antenna array does. While this is not a problem if a stationary biomedical implant needs to be powered, it can be helpful to have some steerability in the case of gastroresident devices, which may move around inside of the stomach. Furthermore, increased directivity allows us to focus electromagnetic fields from multiple transmitters and achieve further gains in efficiency [20].

26
2.1.2 Far-Field Coupling \((d \gg \lambda)\)

The weak directivity and limited efficiency of near-field coupling over large distances prompts one to consider operating at higher frequencies, in the regime of far-field coupling.

In a far-field system, power at high frequencies (generally in the GHz range) is first generated, then passed through a transmitting antenna. The antenna radiates power through the medium in the form of propagating electromagnetic waves. Depending on the directivity of the transmitting and receiving antennas, some fraction of the waves are absorbed by the receiving antenna, and are converted to either alternating or direct current to power a load.

An equivalent circuit is shown in Figure 2-3.

![Circuit diagram](image)

Figure 2-3: Circuit implementations of far-field wireless power transfer over antennas

Neglecting inefficiencies in the generation of RF power, as well as the inefficiency of the rectification of RF waves in the receiver circuitry, the efficiency, \(\eta\) of wireless power transfer in air is given by Friis equation [21]:

\[
\eta = \frac{\lambda^2 G_t G_r}{(4\pi d)^2} \quad (2.4)
\]

where \(d\) is again the distance between the coils and \(\lambda\) is wavelength of the radiation. \(G_t\) and \(G_r\) are gains of the transmitting and receiving antennas respectively (gain is a property that takes into account both the directivity and efficiency of the antennas).

To come up with an estimated efficiency of far-field coupling, let us assume that the transmitter and receiver antennas are half-wavelength dipole antennas in air, constrained to be 6.5 mm or less (to fit in the 000 capsules). This corresponds to antennas that are...
tuned to roughly 24 GHz or higher. The gain of an ideal half-wavelength dipole is 2.19 dB, corresponding to efficiency shown in Fig. 2-4 for various frequencies.

![Far-Field Power Transfer Between 3.5-mm Coils](image)

Figure 2-4: The far-field wireless power transfer efficiency, $\eta$, between two 3.5-mm coils as a function of distance and frequency

What we see is that although the far-field efficiency at small distances (1-2 cm) is not as high as in near-field powering, the efficiency also drops much slower as compared to near-field. Thus, at 6 cm, far-field at all of the frequencies shown in Fig. 2-4 is more efficient than near-field for all of the $Q$'s shown in Fig. 2-2.

There are several downsides to this increased efficiency. Operating at higher frequencies requires more specialized equipment (e.g. network analyzers, signal generators) than lower frequencies. While smaller circuits can be designed at higher frequencies, correspondingly smaller capacitors and inductors are required, and parasitic effects are more costly.

Furthermore, our analysis thus far has assumed that antennas are operating in air. In
reality, there will be dielectric losses if these antennas are used to transmit wireless power to GIEs, as there will be absorption in stomach tissue. The fact that human tissue is far more conductive at higher frequencies (Fig. 2-5), and absorption of radiation inside tissue is proportional to the conductivity of the tissue 2.10, means that we can transmit far less at power into the body while keeping under absorption safety limits (this is discussed in more detail in section 2.2.4) if we choose to operate in the far-field region.

![Figure 2-5: Conductivity of stomach tissue as a function of frequency, showing a significant increase in conductivity in the low-GHz range. Data obtained from [2].](image)

### 2.1.3 Mid-Field Coupling ($d \approx \lambda$)

Given the trade-offs between near-field and far-field coupling, it may be advantageous to operate in between these regimes. This is known as mid-field coupling, and at the distances of transmission we are interested in, corresponds to frequencies of operation around 1-10 GHz.

Mid-field coupling has properties of both near- and far-field coupling: it has moderate directivity, and if frequencies of around 1-2 GHz are chosen, then the absorption in stomach tissue can be kept low. An analytical expression for the efficiency of power transfer in the
case of mid-field powering is more difficult to derive than in the other two cases, because
the more complicated shapes of the electric and magnetic field in this regime do not lend
themselves to neat abstractions that we can readily apply.

Nevertheless, we can write general expressions in terms of the electromagnetic fields
themselves. From [20], the efficiency of a conjugately-matched two-coil system can be written
as:

\[ \eta = \frac{1}{4R_L} \frac{\omega \mu_0^2 A^2 |\vec{H}(d) \cdot \hat{n}|^2}{\int_V \text{Im}[\epsilon(\vec{r})]|\vec{E}(\vec{r})|^2 d\vec{r}} \]  

(2.5)

Where \( R_L \) is the conjugately-matched load resistance, \( \omega \) is the operating frequency, \( \mu_0 \)
is the permeability of free space, \( A \) is the receiving coil area, and \( \vec{H}(d) \cdot \hat{n} \) is the component
of the magnetic field that passes through the receiving coil, assumed constant over the area
of the coil. The product of the imaginary part of \( \epsilon \), the permittivity of the tissue involved,
and \( \vec{E} \), the electric field strength, is integrated over all of the tissue into which radiation is
transmitted.

As can be seen in (2.5), increasing the frequency of operation increases the power effi-
ciency, but in the case of tissue, that gain is eventually offset due to increasing dielectric
losses, as discussed in section 2.1.2. The optimal frequency of operation depends on the
tissue composition and distance between the coils, and has been solved for various tissues in
[20].

Neither \( \vec{E} \) nor \( \vec{H} \) can be easily determined in general for transmitting or receiving anten-
tenas in the midfield, so to get a sense of the efficiency of mid-field powering, I will use
numbers that were derived for the case of a particular transmitting source and receiving
antennas [3]. These antennas were roughly the same size as the GIEs that I am developing
(to be more specific, the size of the transmitter is larger – 6 cm – while the receiving coils
are smaller – 2 mm. However, these differences in sizes, besides being in opposing directions,
are negligible compared to the radiation losses in tissue. Given the area ratios, I believe that
the theoretical curve is accurate to within 10 dB when applied to my antennas).
That efficiency is displayed in Fig. 2-6, and suggests that operation at 1.6 GHz yields reasonable efficiencies, even at large distances of up to 10 cm. Furthermore, the same authors also showed that midfield powering has a moderate amount of directivity, and can be used to focus electromagnetic waves in different areas.

![Mid-Field Power Transfer Between 3.5-mm Coils](image)

Figure 2-6: The far-field wireless power transfer efficiency, $\eta$, between two 3.5-mm coils as a function of distance and frequency, adapted from [3]

### 2.1.4 Comparison of Regimes

Fig. 2-7 shows schematically the three different coupling regimes that we have discussed thus far. To decide which of these regimes is most appropriate for powering GIEs, we can overlay the plots of efficiency for each of the domains. Such a plot is presented in Fig. 2-8.

We are interested in efficiency over many centimeters, suggesting that far-field and midfield are better regimes to operate in. But efficiency is not the only consideration in deciding which regime to use for powering GIEs. Other qualitative considerations, along with efficiency, are summarized in Table 2.1. In particular, I note that mid-field powering allows us...
Figure 2-7: Different configurations of a two-coil system will yield near-field, mid-field, or far-field coupling, depending on the distance between them and operating frequency.

Figure 2-8: The power transfer efficiency, $\eta$, as a function of distance in each of the three different coupling regimes. This graph is a composite of figures 2-2, 2-4, and 2-6.
Table 2.1: Qualitative differences between wireless power transfer across tissue in near-field, mid-field, and far-field domains.

<table>
<thead>
<tr>
<th></th>
<th>Near-field</th>
<th>Mid-field</th>
<th>Far-field</th>
</tr>
</thead>
<tbody>
<tr>
<td>Efficiency</td>
<td>Low</td>
<td>High</td>
<td>High</td>
</tr>
<tr>
<td>Directivity</td>
<td>Low</td>
<td>Moderate</td>
<td>High</td>
</tr>
<tr>
<td>Absorption in Tissue</td>
<td>Low</td>
<td>Moderate</td>
<td>High</td>
</tr>
<tr>
<td>RF Equipment Required</td>
<td>No</td>
<td>Yes</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 2.1: Qualitative differences between wireless power transfer across tissue in near-field, mid-field, and far-field domains.

to achieve high efficiency and moderately high directivity, while keeping absorption in the tissue to a reasonable minimum.

Based on these considerations, I decided to choose mid-field powering to deliver wireless energy to GIEs. The next section quantifies many of these considerations and determines the optimal frequency within the range of mid-field frequencies through simulation.

### 2.2 Simulations in Tissue

In section 2.1, I showed that theoretical considerations strongly favored operating in the mid-field regime. However, several questions need to be answered more precisely: what exact frequency should be used? How much power can be transmitted at that frequency? And how much power do we expect to be received by the GIE at that frequency before breaching safety limits? In this section, I answer those questions by building a tissue model that I then use for microwave simulations.

#### 2.2.1 Distance of Transmission in Tissue

An axial CT scan of the abdomen is shown in Fig. 2-9. The image shows that the stomach is generally found pressed against one side of the abdominal wall. From this scan, we can see that a device that is inside the stomach is roughly one quarter of the diameter of the
waist away from the wall. Given that the average waist size in men is 94 cm (37 in.), I chose a distance of 6 cm (2.4 in.) as being the average distance that wireless power would have to be transferred across, assuming that the external transmitter is pressed against the skin of the abdomen closest to the stomach.

Figure 2-9: A labeled axial CT scan of a normal abdomen, adapted from [4].

The 6 cm that separates the transmitter and receiver coils consists of skin, muscle, fat, and stomach tissue. However, since the electrical properties of various kinds of soft tissue are fairly similar [2] (as they mostly consist of water) and the thickness of skin and muscle is relatively small, I decided to model the tissue environment as consisting solely of stomach tissue. This model, shown in Fig. 2-10, is used for the remainder of this section.
2.2.2 S- and Z-Parameter Simulation

To characterize the wireless power transfer, I built a geometric model consisting of stomach tissue in CST Microwave Studio, using the built-in tissue properties available in CST. I then created two generic 6.5-mm square loops made from 35-micron thick copper, separated by a distance of 6 cm. Each loop is 0.25 mm wide and includes two small 1.6-mm gaps, in one of which I placed a lumped port that an RMS voltage of 1 Volt, and in the other, I placed a lumped capacitor to resonate the coil at a specific frequency. This setup is illustrated in Fig. 2-10.

![Figure 2-10: Screenshot of setup in CST Microwave Studio consisting of stomach tissue model and two 6.5 mm-diameter coils. [4].](image)

Resulting plots of the magnetic field (shown in Fig. 2-11) demonstrate that the coils are far enough apart for magnetic fields to propagate, so the operative coupling is mid- or far-field.

Power transfer efficiency is typically characterized through $S$-parameters. $S$-parameters represent the power received at a port, as a proportion of the power that is inputted into the (same or different) port, as is defined through the subscripts. As such, the $S_{11}$ parameter
Figure 2-11: Results from CST simulation of the magnetic field direction and magnitude (top) and the magnitude of x-component of the magnetic field (bottom).
is defined as the fraction of power that is reflected by the first port when some amount of power is inputted into it. The $S_{21}$ parameter correspondingly refers to the amount of power transmitted to the second port as a fraction of the amount of power inputted into the first port.

As such, the standard measure of power transfer efficiency in our system is $S_{21}$ (or, equivalently, $S_{12}$). However, in our simulations, we cannot use $S$-parameters, because we are sweeping across a frequency range. The amount of power that is actually coupled into the tissue from the first coil depends on how much can actually be transferred into the coil from the port (the rest of it is reflected back). That in turn depends on the impedance of the coil, which is frequency-dependent, thus making $S_{21}$ an inaccurate basis of comparison across frequencies.

Instead, I use $Z$-parameters. $Z$-parameters are defined similarly to $S$-parameters, but are based on ratios of voltages and currents, which can be constrained to defined values on each coil. As such, the $Z_{11}$ is the ratio of the voltage and the current on the first port. The $Z_{21}$ parameter correspondingly refers to the voltage on the second port as a ratio of a fixed current at the first port. Though $Z$-parameters are not used when making physical measurements (e.g. with a network analyzer) because voltages and currents are difficult to measure individually in high-frequency circuits, they can be readily used in simulation.

The equivalent measure of efficiency is $Z_{12}$, but a more useful value in our case is

$$\gamma = \frac{|Z_{21}|^2}{R_1}$$

(2.6)

where $R_1$ is the real part of $Z_{11}$. Maximizing this value corresponds to maximizing the ratio of the power received by the receiving coil to the total power coupled into the tissue by the transmitting coil [3].

First, as a check on simulation, let us take a look at Fig. 2-12, which shows a plot of $S_{11}$ as a function of frequency for various capacitor values. As expected, we find that the dip in $S_{11}$ (the resonant frequency) changes as the capacitor values change. However, as we will
see, the choice of the capacitance does not significantly change the $Z_{21}$ and $\gamma$ curves.

Figure 2-12: Simulations show that the resonant frequency (the minimum of the $S_{11}$ curve) varies as the value of the tuning capacitor is changed.

Next, Fig. 2-13 plots the value of $Z_{21}$ as a function of frequency and capacitor values. We see that almost regardless of value of the lumped capacitor, the frequency at which $Z_{21}$ is maximized is very close to 1.1GHz. This is similarly true for the plot of $\gamma = \frac{|Z_{21}|^2}{R_1}$ shown in Fig. 2-14.

Since the curves for $Z_{12}$ and $\gamma$ produce optimal frequencies that are independent of the tuning capacitor, we can be confident that these curves are indicative of which frequency is best based on tissue properties, rather than the resonance of the coils themselves. Based on these simulations, we can conclude that the optimal frequency for mid-field operation is around 1.1 GHz.
Figure 2-13: The $Z_{12}$ impedance curves show that regardless of the value of the tuning capacitor, the curves take a maximum close to 1.1 GHz.

### 2.2.3 Efficiency at Optimal Frequency

Neither $|Z_{12}|$ nor $\gamma$ directly provide us with values of power transfer efficiency that are comparable to those values theoretically calculated in section 2.1 and plotted in Fig. 2-8. Rather, the efficiency is given [3] by

$$\eta = \frac{|Z_{12}|^2}{4R_1R_2} \quad (2.7)$$

where $R_2$ is defined (analogous to $R_1$) as the real part of $Z_{22}$. Plots of the real part of $Z_{11} = Z_{22}$ (for identical coils) are shown in Fig. 2-16.

We are interested in the best possible efficiency when the coils are operating at 1.1 GHz, so we should have the coils resonate at that frequency. From [22], we can see that the inductance of a square loop with the dimensions we have chosen is 17 nH. Finding the capacitance value to resonate with that inductance is straightforward:
Figure 2-14: The plotted graphs of $\gamma$ show that regardless of the value of the tuning capacitor, the curves take a maximum close to 1.1 GHz.

$$C = \frac{1}{(2\pi f)^2 L} \approx 1.2 \text{ pF}$$  \hspace{1cm} (2.8)

Of the tuning capacitances that we have chosen, the closest to this value is the green curve, in which the tuning capacitor is 1.12 pF. We can confirm that the coil does indeed resonate at 1.1 GHz (among other frequencies) with this tuning capacitor by plotting the imaginary part of $Z_{11}$ and noting that it approaches zero at 1.1 GHz, as seen in 2-15.

We can thus identify the values of $R_1 = R_2$ by looking at the real part of the $Z_{11}$ of the coils at 1.1 GHz when the tuning capacitor is set to 1.12 pF.

Now, we can substitute into (2.7) values from Fig. 2-13 and Fig. 2-16 to calculate the efficiency as:

$$\eta = \frac{(1.22)^2}{4(128)(128)} \approx 1.9 \cdot 10^{-5}$$  \hspace{1cm} (2.9)
Figure 2-15: The curves show the imaginary part of the impedance $Z_1$ for different frequencies and tuning capacitors.

Table 2.2: Simulated value of power transfer efficiency and comparison with theoretical value.

<table>
<thead>
<tr>
<th>Property</th>
<th>Theory (dB)</th>
<th>Simulation (dB)</th>
<th>Simulation (magnitude)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\eta$</td>
<td>-35</td>
<td>-47</td>
<td>$1.9 \cdot 10^{-5}$</td>
</tr>
</tbody>
</table>

The value is boxed in Table 2.2, and is less than, though comparable to, predicted by theory, which had been calculated assuming a more complex transmitting source, designed to focus radiation into the body (cf. Fig. 2-8).

2.2.4 Specific Absorption Rate

We have shown that the optimal frequency of operation is 1.1 GHz, and at the expected efficiency between the transmitter and receiver is about -47 dB, but to assess the feasibility of wireless powering, we must answer: how much power can we transmit from outside the
Figure 2-16: The curve shows the real part of the impedance $Z_{11}$ for different frequencies and tuning capacitors. When the tuning capacitor is 1.12 pF, the real impedance is 128 ohms at 1.1 GHz.

The limiting factor that determines the amount of power that can be coupled into the tissue is the specific absorption rate (SAR), defined [5] as follows:

$$\text{SAR} = \frac{1}{V} \int_{V} \frac{\sigma(\vec{r}) |E(\vec{r})|^2}{\rho(\vec{r})} d\vec{r}$$  \hspace{1cm} (2.10)$$

where $V$ is the volume of the integration, $\sigma$ is the conductivity, $|E|^2$ is the norm-squared of the electric field, and $\rho$ is the mass density. Regulatory groups including the IEEE have set limits to the SAR allowed in human tissue. These limits are summarized in Table 2.3. IEEE distinguishes a "low-tier" standard from a "high-tier" standard; the high-tier standard refers to the maximum SAR in controlled settings, while the low-tier standard includes a margin of safety to account for imprecisely-controlled settings. Both standards are for locally-averaged values of SAR over 10 grams of tissue [5].
<table>
<thead>
<tr>
<th></th>
<th>Low-tier</th>
<th>High-tier</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum SAR over 10 g of tissue</td>
<td>2 W/kg</td>
<td>10 W/kg</td>
</tr>
</tbody>
</table>

Table 2.3: IEEE limits on SAR in human tissue, see [5]

To simulate the SAR in human tissue due to a transmitting antenna, we turn to COMSOL Multiphysics Suite. A detailed model of the transmitting and receiving antennas is first built and meshed in COMSOL, as shown in Fig. 2-17.

![Figure 2-17: The COMSOL model of the antennas on FR4 substrate, placed in stomach tissue, including a box (purple) that represents 10 g of tissue.](image)

Next, one of the antennas is stimulated at various power levels, and the resulting electric field is computed throughout the (background) tissue. A boxed volume (2.15 cm on a side) is added 1 mm from the transmitting antenna to represent the 10 g of tissue. From the resulting electric field (Fig. 2-18), the SAR can be computed and averaged within the 10 gram of tissue.
Figure 2-18: The log (base 10) of the calculated electric fields (V/m) inside the stomach tissue due to the transmitting antenna.

The resulting SAR is plotted as a function of power levels in Fig. 2-19. From the plot, we see that we need to emit below about 28 dBm to keep under the high-tier standard, and about 20 dBm to keep under the low-tier standard.
Figure 2-19: The averaged value of SAR as a function of transmit power from the antenna outside stomach tissue. Horizontal lines represent IEEE limits [5].
Chapter 3

Antenna Design

The results of chapter 2 indicate that wireless power transfer across stomach tissue occurs most efficiently in the mid-field regime. The measured efficiency will depend on the specific antenna structure utilized. In this chapter, I discuss the choice and design of the antenna and matching network, and characterize them in air and in stomach tissue.

3.1 Small Loop Antennas

An antenna operating frequency of around $\nu = 1$ GHz, as determined in chapter 2, corresponds to a wavelength of $\lambda = c/\nu = 30$ cm. Even a quarter-wavelength resonant antenna is far too large to fit inside an ingestible capsule that is roughly 2 cm long. Instead, we must use a small non-resonant antenna in which the antenna resonates not due entirely to its physical size but because of the addition of external loads, generally lumped elements [23].

Among choices for small antennas (e.g. short dipole, short monopole, small loop), the small loop antenna stands out for several reasons. The small loop antenna is a magnetic antenna, whose near field is dominated by magnetic fields, not electric fields. Because the permeability of the body is close to the permeability of free space, small loop antennas are less affected by being close to the body, and in turn affect the body to a lesser extent [16]. Furthermore, small loop antennas can be fabricated without a ground plane (freeing up more
space on a two-sided printed circuit board). These advantages outweigh the disadvantage of small loop antennas: their relatively low radiation resistance, which corresponds to a low radiation efficiency.

The small loop antennas that I used were square loops printed using copper trace on an FR4 substrate, as shown in Fig. 3-1.

Figure 3-1: A 6.5 mm per side square loop antenna, printed on FR4 substrate, along with tuning capacitor and an MMCX connector.

To resonate with the inductance of a small loop antenna, a capacitor was included in parallel with the loop antenna. The inductance, $L$, of a square loop of width, $w$, and trace thickness, $d$, is given by [22]:

$$L = \frac{\mu_0}{\pi} \left[ (-4 + 2\sqrt{2} + 2\ln(1 + \sqrt{2}))w + 2w(\ln\frac{2w}{d}) \right]$$

(3.1)

For a trace thickness of 0.25 mm and a side length of 6.5 mm, the square loop has an inductance of about 17 nH in free space. As such, a lumped capacitor of value 1.2 pF is needed to resonate the loop at 1.1 GHz. Since lumped capacitors only occur at discrete values, I chose a 1 pF tuning capacitor, resulting in a resonant frequency of 1.24 GHz.

Because of the small size of the loop antenna, there was the possibility of the wiring picking up more radiation than the antenna itself, and thus incorrectly inflating the efficiency of the system during testing. Thus, I chose to use small MMCX connectors instead of the more standard SMA connectors (Fig. 3-2) to attach to the board.
3.2 Antenna Matching Networks

Because of the low radiation resistance and ohmic resistance of small loop antennas, a matching network is needed to interface the antenna to the 50-Ω connectors and test equipment.

The radiation resistance and ohmic resistance of the small loop antenna can be given in [24] as:

\[ R_R = 3.92 \cdot 10^{-30} \left[ \frac{\Omega}{m^4 \cdot Hz^4} \right] (\nu \cdot w)^4 \approx 0.05 \Omega \]  
\[ R_L = 1.29 \cdot 10^{-4} \left[ \frac{\Omega}{Hz^{0.5}} \right] \sqrt{\frac{\nu \cdot w}{d}} \approx 0.45 \Omega \]

respectively. The total resistance is thus approximately 0.5 Ω. To match this to 50 Ω, we can use a variety of matching networks. The simplest matching network – or at least the one with the fewest lumped elements – is the L-network, illustrated in Fig. 3-3. The relevant equations for the L-network are found in [25] and reproduced below.

\[ Q = \sqrt{\frac{R_L}{R_S}} - 1 \]  
\[ L = \frac{Q R_S}{\omega} \]
\[ C = \frac{Q}{\omega R_L} \]  \hspace{1cm} (3.6)

Figure 3-3: An L-network topology that converts a higher source resistance \( R_S \) into a lower load resistance \( R_L \), while the position of the reactive elements also serves to block DC current.

These equations produce values of 0.72 nH and 29 pF for the inductor and capacitor. Again, due to the nature of discrete lumped elements available, we end up using 1 nH and 33 pF. A small loop antenna that has been fabricated completely its matching network is shown in Fig. 3-2 (right).

3.3 \textbf{S-parameter Characterization}

I then proceeded to characterize the antennas using \( S \) parameters measured from a calibrated HP 8753C Network Analyzer. The \( S_{11} \) parameter measures how much power is reflected from the antenna back into the incident port that is powering the antenna. As such, it is a measure of how well the matching network and tuning capacitor cause the antenna to radiate power at the desired frequency.

The \( S_{11} \) was measured with the antenna embedded in stomach tissue (after the antenna had been wrapped in insulating tape, to simulate the encapsulation of the GIE). The the setup is shown in Fig. 3-4. The results are presented presented in a log-magnitude plot in Fig. 3-5. We can see from this plot that the antenna resonates at 1.457 GHz. Multiple fabrications of this antenna showed that the antenna frequency was consistently within 10 MHz of 1.457 GHz. The reason that the resonant frequency was higher than our target value
is likely due to the non-exact values of capacitors and inductors used in the L-network and for tuning the antenna.

The reflection $S$ parameter is shown for a second antenna also embedded in tissue (Fig. 3-6). In this case, the resonance is not as peaked (return losses are higher). This is likely due to variations in the capacitor and inductor values used for the L-network, as variations in the lumped elements will cause some matching networks to perform better matches than other networks.

![Figure 3-4: Two square loop antennas placed in stomach tissue 3 cm apart in order to characterize antenna S-parameters.](image)

After the reflection characteristics of the antennas had been characterized, I also measured the transmission characteristics between two antennas. This was measured by the $S_{12}$ parameter, a measure of what fraction of the power transmitted from the port connected to one antenna was received by the port connected to the second antenna.

Results from the $S_{12}$ measurement taken when the antennas were separated by 2 cm are presented in a log-magnitude plot in Fig. 3-7. We can see from this plot that the at the resonant frequency of 1.457 GHz, the efficiency of transfer is -32 dB, or just under 0.1%.
Figure 3-5: $S_{11}$ characteristics of first small loop antenna show a resonance at 1.46 GHz of reflection magnitude -21 dB.

3.4 Efficiency as a Function of Distance

By repeating the $S_{12}$ measurement at various distances, we can measure the efficiency as a function of distance. Furthermore, we can compare our measured efficiencies to those from theory with ideal components. Such a plot of measured and theoretical efficiencies is shown in Fig. 3-8. We see that at 4 cm apart, the efficiency is about -40 dB, and it drops to -45 dB at 6 cm.
Figure 3-6: $S_{22}$ characteristics of first small loop antenna show a resonance at 1.46 GHz of reflection magnitude -12 dB.

Figure 3-7: $S_{12}$ characteristics of the antennas show that at 1.47 GHz, the efficiency (at 2 cm apart) is -32 dB.
Figure 3-8: Measured $S_{12}$ efficiency as a function of distance in stomach tissue. Each dot represents a single measurement, while the solid blue line passes through the median measurement at each distance. The dashed red line is theory, adapted from section 2.1.3.
Chapter 4

Rectifier Network

The antennas propagate electromagnetic radiation through tissue in the form of RF waves. At the receiver, the radio-frequency waves need to be converted to direct current for most usable applications. This process of rectification is critical as inefficient rectifiers can significantly degrade the overall efficiency of power transfer. This section discusses RF rectifier design and simulation.

4.1 Background and Design

Growing interest in harvesting energy from ambient RF sources have resulted in a proliferation of circuit topologies for the rectification of radio-frequency electromagnetic waves [26]. Depending on the operating frequency, received power, and required DC voltage, different topologies may be used.

For GIEs operating at 1.46 GHz, the requirements are summarized in Table 4.1. The -17 dB received power level is a consequence of the results found in chapters 2 and 3 (28 - 45 = 17 dBm). The 1.8 V of required DC voltage is based on the choice of microcontroller, which is discussed in more detail in section 5.2.

Given these requirements, I decided to construct a multi-stage voltage doubler circuit, as found in [27]. A schematic of one unit of this circuit is found in Fig. 4-1. The basic voltage
doubler unit is two Schottky diodes connected in anti-parallel. Two capacitors are included, the first of which stores power during one half-cycle, while the second stores power during the other half-cycle.

![Diagram of voltage doubler schematic](image)

**Figure 4-1:** A single-stage voltage doubler schematic consisting of two Schottky diodes and two capacitors.

For the diodes, I chose HSMS-282C Schottky diodes because of their low forward voltage and high switching time (to rectify low power and high frequency signals). The capacitors were set to 68 pF based on [27], while the number of stages was left as a design parameter that I optimized with the use of simulation.

### 4.2 Simulation

Voltage-doubler circuits can be stacked in series to achieve a higher DC voltage [27], as shown in Fig. 4-2. The number of stages, $N$, determines the output voltage that will drive...
the load. Because the output voltage rises over the course of just a few microseconds (see Fig. 4-3 for the output of a 1-stage voltage doubler), it was sufficient to run a transient simulation for 1.4 $\mu$s to determine the output voltage.

![Figure 4-2: A N-stage voltage doubler schematic consisting of 2N Schottky diodes and 2N capacitors. The stages are connected in series.](image)

The output voltage as a function of the number of stages is shown in Fig. 4-4. As can be seen from the figure, it is possible to achieve an output of 1.8 V with a 9-stage voltage doubler, thus showing that RF power 1.457 GHz can be converted to usable energy to run a microcontroller.
Figure 4-3: The rectified voltage rises to a stable value of about 0.3 V from a single-stage voltage doubler within a few microseconds.

Figure 4-4: The rectified voltage increases as the number of stages increases.
Chapter 5

Circuit Board Integration

The purpose of the wireless powering system that we have developed over the last few chapters is to provide enough energy to power a circuit board for GIE applications. In this chapter, we discuss and design such a circuit board for a representative application: continuous measurement of core body temperature and wireless transmission of this information to a receiver outside of the body.

Besides being a proof-of-concept device, this circuit board also serves as a prototype that can be characterized for its electrical and material properties, as is carried out in chapter 6.

5.1 Size Constraints

Our mechanism of choice for gastroretention is the three-arm elastic polymer that is folded into a pill and unfolds inside of the stomach. Because the largest pill typically ingested by humans is the 000 capsule (see Fig. 1-1), this sets limits to the size of the circuit board that can house the components on the receiver.

In particular, if the circuit board is designed to fit within the one thick arm of the three-legged structure as shown in the photo of the mock devices in Fig. 5-1 (the thick arm is colored black in device to the left), then the circuit board can be at most 23.5 mm long along the sides, 7 mm wide, and 3 mm thick. The exact dimensions can be seen in in Fig. 5-2.
Figure 5-1: Two model GIEs are shown. The mock circuit board is included in the thicker arm of each GIE, and is thus limited by its dimensions.

Figure 5-2: The dimensions of the thicker arm are as shown here, and so the complete circuit board must fit within these dimensions.
Within these dimensions, I needed to fit the essential electronic components of the GIE: a microcontroller, temperature sensor, and the parts required for wireless communication. A variety of manufacturers produce parts that are general-purpose microcontrollers, sensors, and antennas, but in this case, I needed to consider the tight design constraints of the GIE.

5.2 Microcontroller

The factors that constrain the choice of a microcontroller suited for GIE applications include size, low power consumption, ease of programming, and radio capability. Among radio-enabled microcontrollers available in the market that could fit within the size constraints in Fig. 5-2, two stood out: the TI CC2460 and the Nordic NRF51822. The CC2460 has a 4-mm x 4-mm footprint, while the NRF51822 has a 6-mm x 6-mm footprint. Both are ultra low-power devices, with standby currents of around 1 µA. However, the NRF51822 can be programmed much more easily (using freely available software such as Keil µVision and NRFGo) as opposed to needing proprietary TI software for the CC2460 - thus, I chose the NRF51822 (Fig. 5-3, top left).

5.3 Temperature Sensing

Although the NRF51822 includes an on-chip temperature sensor, the resolution of the sensor is fairly limited (0.25 °C), so an external temperature sensor is needed to measure fluctuations in core body temperature. This sensor needs to be small, consume little power, and be able to interface with the microcontroller. For these reasons, I chose to use a TI TMP102 with 0.06 °C resolution and an active (max) current draw of 85 µA, pictured in Fig. 5-3 (top right).
5.4 Wireless Communication (Bluetooth)

The NRF51822 microcontroller is radio-enabled, but still requires an external 2.4 GHz antenna and matching network for wireless communication. A small ceramic chip antenna, the RFANT801008A3T, was chosen due to its 8 mm length and its 3dB-bandwidth of more than 100 MHz, which makes it resilient to detuning in the body. It is shown in Fig. 5-3 (bottom left). A balun and 50-Ω matching network are needed to interface with the single-ended antenna. Both features are provided with the 2450BM14E0003 chip, which also features a low insertion loss (0.9 dB) at 50 Ω.

5.5 Battery

Although the GIE is designed to be wirelessly powered, the circuit board used for testing and characterization was battery-powered. The battery-powered prototype allowed me to fabricate and test of the non-wireless power components in the microcontroller, as well as to easily characterize the robustness of the device in gastric environments as discussed in Ch. 6. Furthermore, in a complete application, the GIE is likely to include both a wireless power receiver and a rechargeable battery that stores energy.

The biggest consideration for a rechargeable battery was finding one that could fit within the size constraint of the 000 capsule. One option was the Seiko MS621 coin cell battery, with a 6.8-mm diameter. The other option was thin-film batteries, such as the Cymbet CBC050. Although the Cymbet batteries were even smaller (seven of them could fit within the same space as one Seiko battery), they also had a much smaller battery capacity (all seven together had one tenth of the capacity of the coin cell), and for that reason, I ended up choosing the coin cell (see Fig. 5-3, bottom right).
Figure 5-3: A selection of components used in the assembled GCE: a QFN package NRF51822 illustrated with a shield (top left), the TI TMP102 illustrated with a shield (top right), the external RFANT8010080A3T antenna (bottom left), and the 6.8-mm Seiko MS621 coin cell battery (bottom right). Pictures adapted from commercial vendor pictures or datasheets.
5.6 Schematic and Layout

Once the components were specified, the schematic of the GIE was prepared. The oscillator and balun were connected to the designated pins on the NRF51822, while the temperature sensor was connected to general purpose input/output pins. Vias were added to allow external wires to connect to the power, ground, and programming pins. Pads for the coin cell were connected to power and ground as well, although the connection to ground included a 1 μF capacitor (C7) in series to provide power to the microcontroller during peak current consumption (during the transmit and receive portions of the Bluetooth protocol).

The completed schematic of the GIE is shown in 5-4, and the layout of both the top and bottom of the two-sided board is shown in 5-5. The board was fabricated by Sunstone (Mulino, OR, USA), and is shown in Fig. 5-6.
Figure 5-4: Schematic of the complete GIE showing how the electronic components are connected.
Figure 5-5: Layout of the complete GIE showing how the electronic components are laid out on the circuit board. The top side (left) and bottom side (right) are shown separately.
Figure 5-6: Pictures of the fabricated and assembled circuit board, its top side (top) and bottom side (bottom).
Chapter 6

Device Characterization

After the fabrication of the circuit board, I characterized the electrical and material properties of the board. Characterization of the power consumption was necessary to determine the whether the amount of power delivered the microcontroller would be able to sustain the microcontroller for an extended period of time. Basic material characterization was needed to ensure that the board remained robust in the gastric environment. This was tested by encapsulating the board with different materials and then placing it in simulated gastric fluid (SGF). Finally, in order to ensure reliable wireless communication with the power levels tested, radio strength of the board was measured in ex vivo and in vivo settings.

6.1 Power Consumption

To measure the power consumption of the board, it was programmed with a standard Blue-tooth test program and connected to an Agilent N6705B power analyzer. The voltage was fixed at 3 volts, while the current was measured. Because the current was not constant (it spiked during radio broadcasts, for example), 60-second readings were recorded, and the average current computed in post-processing.

The current consumption of the board varied depending on the Bluetooth radio broadcast level (programmable from -30 dB to +4 dB) and the time between radio broadcasts. It also
Table 6.1: Power consumption during advertisement mode. *Bluetooth Low Energy (BLE) standard specifies that the advertisement interval is < 10 seconds

<table>
<thead>
<tr>
<th></th>
<th>-30 dBm</th>
<th>-10 dBm</th>
<th>0 dBm</th>
<th>+4 dBm</th>
</tr>
</thead>
<tbody>
<tr>
<td>25 ms</td>
<td>0.65 mA</td>
<td>0.96 mA</td>
<td>0.93 mA</td>
<td>1.07 mA</td>
</tr>
<tr>
<td>1 s</td>
<td>0.023 mA</td>
<td>0.030 mA</td>
<td>0.030 mA</td>
<td>0.033 mA</td>
</tr>
<tr>
<td>10* s</td>
<td>0.008 mA</td>
<td>0.009 mA</td>
<td>0.009 mA</td>
<td>0.009 mA</td>
</tr>
</tbody>
</table>

Table 6.2: Power consumption during transmission mode. *BLE specifies that the connection timeout is < 32 seconds, after which advertisement resumes.

<table>
<thead>
<tr>
<th></th>
<th>-30 dBm</th>
<th>-10 dBm</th>
<th>0 dBm</th>
<th>+4 dBm</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 s</td>
<td>0.018 mA</td>
<td>0.018 mA</td>
<td>0.019 mA</td>
<td>0.020 mA</td>
</tr>
<tr>
<td>10 s</td>
<td>0.008 mA</td>
<td>0.008 mA</td>
<td>0.008 mA</td>
<td>0.008 mA</td>
</tr>
<tr>
<td>30* s</td>
<td>0.007 mA</td>
<td>0.007 mA</td>
<td>0.007 mA</td>
<td>0.007 mA</td>
</tr>
</tbody>
</table>

depended on whether the device was already paired over Bluetooth (in *transmission* mode) or was searching for a Bluetooth-enabled device to pair to (advertisement mode). Tables 6.1 and 6.2 summarize the current readings from the Agilent N6705B.

As a result of this characterization, I concluded that the average power consumption at the receiver could be limited to 0.024 mW (3V × 0.008 μA), corresponding to a value of -17 dB. (This is the current during advertisement when the advertisement interval is set to 10 seconds)

6.2 Integrity of Board in Simulated Gastric Fluid

To ensure that the circuit board operated without shorting in the gastric acid environment, the board was encapsulated in different materials before being placed in simulated gastric fluid. Three types of encapsulation were tested independently: PDMS alone, PDMS preceded by a layer of superglue (Scotch, MN, USA), and PDMS preceded by a layer of 5-minute
epoxy (Devcon, MA, USA). The application of superglue and epoxy was done with a brush to thoroughly apply a coating all over the microcontroller before it was encased in a layer of PDMS as shown in Fig. 6-1 (left).

The encapsulated boards were then placed in a jar of simulated gastric fluid (Fig. 6-1 [right]) and incubated at 37 °C. Every 12 hours, the encapsulated GIE was connected to an external powering source to test whether the integrity of the circuit board was compromised. I found that while PDMS encapsulation alone only provided a circuit board lifetime of less than 48 hours, a second layer of encapsulation with epoxy or glue extended it to beyond 2 weeks, after which testing ended. These results are shown in Fig. 6-2.
Figure 6-2: A circuit board that was directly encapsulated in PDMS only survived 2 days inside SGF before shorting, but the addition of a coat of superglue or epoxy extended the lifetime to more than 2 weeks (testing ended at 2 weeks).

6.3 Radio Strength in Simulated Gastric Fluid

Finally, we needed to ensure that the power consumption of the board was enough to wirelessly communicate with an external telemeter system. To determine whether this was feasible, radio strength was measured in three conditions: the power from a bare board, the power from an encapsulated board, and the power from an encapsulated board in a 1-L jar of SGF.

The results are presented in Fig. 6-3. And as can be seen from the received strength signal indicator (RSSI) values, in all three scenarios, wireless signal could be picked up by an external telemeter (in this case, a Samsung Galaxy S6 phone with a minimum receive threshold of -100 dBm) up to 10 ft (three meters) away.
Figure 6-3: Received strength signal indicator measured by a receiver as a function of distance from a GIE in different environments.
Chapter 7

Conclusion and Further Research

In the preceding chapters, I have designed and characterized (1) a gastroretentive ingestible electronic (GIE) device capable of measuring core body temperature and transmitting the measurement to an external receiver over Bluetooth, along with (2) a wireless mid-field transmitter and receiver system capable of transferring enough power through the body to sustain the GIE indefinitely inside of the stomach.

Let us review the key results of this thesis. In chapter 2, we demonstrated that it is possible to transmit up to 28 dBm of power into the body using 1.1 GHz small loop antennas, while remaining under IEEE limits on specific absorption of radiation. The efficiency of the power transfer at this frequency is around -45 dB, as measured through 6 cm of stomach tissue and discussed extensively in chapter 3. This power can be rectified to a usable DC voltage, based on the design similar to that presented in chapter 4. Assuming a completely efficient rectifier, this gives us $28 - 45 = -17$ dBm of usable power. The choice of electronics in the design of the GIE, as discussed in chapter 5 and 6 allows us to just achieve this bound, with a device that is stable in a simulated gastric environment, and powerful enough to communicate from inside of it.

The principal implication of my work is that wireless GIEs are a feasible technology. This work opens the doors to ingestible devices that reside inside the stomach – continuously
monitoring vital signs or tracking diet via biochemical sensors or delivering drugs on demand.

To continue developing wireless GIEs, immediate next steps are fabricating a device that includes the full receiving antenna and rectifier network and testing it *ex vivo*. Subsequently, animal experiment should be performed to confirm that midfield wireless is efficient even in the case when orientation and position of the receiving antenna cannot be controlled precisely. In parallel, further development of biochemical sensors and gastroretentive mechanisms will be important steps to ensure that the full potential of wireless GIEs is achieved.
Bibliography


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Optimization of the voltage doubler stages in an RF-DC convertor module for energy 