Abstract

In today’s world of ever-increasing low-power portable electronics, from implants to wireless accessories, powering these devices efficiently and conveniently is an escalating issue. The proposed solution is to wirelessly recharge these lower-power portable devices through a common magnetic link with a higher-power portable device, such as a smartphone. Such a method is convenient for users, environmentally friendly, and cheap to implement.

This portable-to-portable wireless charging application differs from conventional charging pad-based systems in that the transmitter is energy constrained, so system efficiency is key. Also, since both the transmitter and receiver are portable, loading on the transmitter changes dynamically, which affects efficiency and delivered power.

This thesis addresses these challenges through the design of an efficient and robust wireless charging system. The first half of the thesis presents a transmitter power amplifier control loop for increasing efficiency and balancing power across changing loading conditions. Mathematical analysis of the resonant inductive wireless power circuits shows the impact of changing conditions on power amplifier zero-voltage switching, and its effect on efficiency and power. The control loop adjusts the power amplifier shunt capacitance and series inductance to maintain zero-voltage switching while regulating delivered power.

The second half of the thesis presents the implementation of a resonant inductive wireless charging system operating at 6.78 MHz that transfers energy between portable devices with high efficiency. A custom integrated circuit designed in 0.18 µm HVCMOS implements the derived control loop by sensing for power amplifier zero-voltage switching and adjusting the power amplifier components. An end-to-end efficiency of 78% is achieved when delivering 200 mW over a 7 mm distance. Efficiencies over 70% are maintained over 4-12 mm distances.

A diverse set of applications are demonstrated that use a smartphone to wirelessly recharge a fitness tracker, a cochlear implant, an MP3 player, a calculator, a toy light, a wireless keyboard, and a bicycle light, charging most devices in 2 minutes for a typical day’s use.

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Chapter 1

Introduction

Since the 1800s when inventors harnessed electricity for the first telegraphs and light bulbs, development of modern electronics has relied on the availability and abundance of energy sources. These electronics operate with power delivered at appropriate times, whether it’s from an AC-DC adapter to perform a computation in a processor or from a transmission line to spin an electric motor. In many cases, power is transferred to the device through conductive wire, such as through the cables of the AC-DC adapter or the lines of the transmission grid.

Wireless power transfer is an old technology with new popularity. Its roots can be traced back to Nikola Tesla, who demonstrated in the late 1800s that light bulbs could be powered through wireless coupling of energy from giant high-voltage coils. At the time, wirelessly powered devices existed only in Tesla’s private labs and public lectures. Today, many products exist and much research is being undertaken relating to many implementations of wireless power transfer, at a variety of power levels from micro-watts to kilowatts.
1.1 Current Wireless Power Technologies

Figure 1.1: Overview of current wireless power technologies

Figure 1.1 shows the range of wireless power technologies currently being explored in both the commercial and academic arenas, over a wide range of power levels [5, 6, 7, 8, 4, 9].

On the low-power end, many medical implants rely on wireless power to function. One example is the cochlear implant, a device with implanted electrodes that stimulate auditory nerves to promote hearing [5]. According to the Food and Drug Administration, as of December 2012, about 324,000 people worldwide have received these devices [10]. Each implant is coupled through a magnet to a power transmitter outside the ear, which transfers milliwatts of power wirelessly through the skin. Researchers are also developing retinal implants that stimulate electrodes in the eye to promote sight [4]. These medical devices require on the order of 100 mW of wirelessly transmitted power. Outside the medical space, researchers are also investigating systems to wirelessly charge lithium ion batteries at 10 mW rates [8].

In the middle of the power spectrum, many commercial products are on the market today that provide wireless power for popular consumer electronics such as cell phones and tablets. One example is the Duracell Powermat, which is a charging pad plugged into an electrical outlet that recharges compatible consumer devices placed on top [6]. Devices such as cell phones may be outfitted with a special case to support wireless recharging. Another example
are the WiTricity charging pad products to transfer power to cell phones and tablets. WiTricity repeaters can be used to extend the wireless power range to charge devices from farther away [11].

These products are among many similar ones that adhere to specific industry standards for wireless power transfer at about 1-10 W levels. One such standard is the Power Matters Alliance, with members such as WiTricity, Duracell, and Qualcomm [12]. Another standard is the Wireless Power Consortium, or Qi, with members such as Texas Instruments, Energizer, and Panasonic [13]. Both standards cover resonant wireless power transfer with applications such as charging mobile devices on a charging pad over distances on the order of 1 cm.

On the high-power end, at about 1 kW levels, people in both academia and industry are developing technologies to power large machines. At CES 2011, Fulton Innovation demonstrated the technology to wirelessly charge batteries in an electric vehicle [7]. Researchers are studying system designs for higher efficiency power transfer to wafer processing machines [9].

1.2 Motivation for Wireless Charging of Portable Electronics

Given all the wireless power technologies currently in existence operating at various power levels, there is at least one power level regime that has not been heavily explored: the low-to-mid power region between 100 mW and 1 W. Many consumer portable electronics can be recharged at these levels, from wireless accessories, such as wireless keyboards, to AA(A) battery-operated devices, such as graphing calculators, to medical implants, such as cochlear implants or pacemakers.

These portable electronics are used on a daily basis, and thus need to be recharged regularly. But the current methods of recharging portable electronics are both inconvenient and environmentally unfriendly. Users either charge the device through a charging cable or directly recharge the batteries inside the device, and both of these methods involve external equipment (cables or chargers) not built into the portable that must be carried with the portable. This may be a hassle for users. In the case of medical implants, regularly wearing external equipment to charge or power these implants could be a source of social stigma as well. In addition, at the end of life, this external equipment contributes to a large amount of environmental waste,
including 179,000 tons of waste batteries and 51,000 tons of waste chargers per year in the United States [14, 15].

At the same time, companies and researchers are rapidly developing wireless power technologies for ubiquitous mobile devices such as cell phones and tablets, hence the industry standards described earlier. The wireless power circuitry in these higher-power portable devices could not only be used as receivers to charge these devices, but also as transmitters to charge lower-power portable devices.

A system that wirelessly recharges portable electronics using a higher-power mobile device is both convenient for users and environmentally friendly. The hassle of carrying external charging equipment is eliminated as portable devices would have a built-in common wireless interface with a mobile device for transferring power. As a result, associated waste is also eliminated. The system is also a convenient and relatively stigma-free solution for charging medical implants.

A wireless charging system using a higher-power mobile device is relatively cheap to implement relative to the cost of current charging methods. If the higher-power mobile device is already part of another system that wirelessly charges the mobile device, then the device only needs to be outfitted with a power amplifier and control logic circuits to transmit power to other portable devices. In essence the only added costs and electronics of the proposed system would be in the receiver portable devices. Today, people in the United States spend an estimated 10.6 billion dollars yearly on secondary lead-acid batteries to power portable devices, or about twenty dollars per person per year [16]. Considering that the proposed system involves only circuit-level additions to already-existing electronics, this price point is not difficult to beat.
1.3 Vision for Portable-to-Portable Wireless Charging

Figure 1.2: *Power sharing* structure for portable-to-portable wireless charging

Figure 1.2 shows the vision for wireless charging from a higher-power mobile device to a lower-power portable device. The vision for future wireless power transfer is inspired by the structure of data transfer in place today. Currently, large amounts of data are transferred in a server-client structure in which servers with larger data storage capacity transfer data to clients with smaller capacity through a common cloud interface. This structure is convenient for users because a client can retrieve data from virtually any server with a simple connection to the cloud. The common infrastructure also reduces redundancy and waste. The benefits of the data transfer structure can be extended to power transfer if a similar structure is adopted in which devices with higher power storage capacity transfer power to devices with lower power storage capacity through a common wireless interface. We define this structure for power transfer as *power sharing*.

1.3.1 Representative Use Cases for Portable-to-Portable Wireless Charging

To illustrate the convenience of the proposed wireless charging structure, consider an example use case. Suppose an MP3 player, strapped to a runner's arm, needs to be recharged. The runner is outside and does not have access to an outlet with which to use a regular charger, and the runner does not want to spend unreasonable amounts of valuable exercise time to recharge the music player. With power sharing, the runner simply removes his cell phone from his pocket and brings it close to the MP3 player on his arm, about 10 mm away. The cell phone starts
transferring power wirelessly to the MP3 player, taking about 2 minutes to charge. Afterwards, the runner can use the newly recharged MP3 player for the rest of his exercise routine.

As another example use case, suppose a person has a cochlear implant. With a conventional implant, she needs to wear a set of external equipment to constantly stream power to her implant. This equipment may not be waterproof, thus preventing her from engaging in water sports or limiting usage in the shower. This external equipment is bulky and thus may contribute to social stigma. With power sharing, she no longer needs to wear this external equipment. Instead, she uses her cell phone to wirelessly recharge her implant in the morning in a couple minutes for an entire day’s usage. Her cochlear implant is now truly invisible.

Table 1.1 shows a summary of the representative use cases for the proposed system.

<table>
<thead>
<tr>
<th>Device to be Charged</th>
<th>Charging Time</th>
<th>Usage Time</th>
</tr>
</thead>
<tbody>
<tr>
<td>MP3 Player</td>
<td>2 Minutes</td>
<td>30 Minutes</td>
</tr>
<tr>
<td>Cochlear Implant</td>
<td>2 Minutes</td>
<td>8 Hours</td>
</tr>
</tbody>
</table>

Table 1.1: Representative use cases for portable-to-portable wireless charging

1.3.2 User Metrics for Portable-to-Portable Wireless Charging

The performance of power sharing is dependent on some user-centered metrics. One important metric is end-to-end efficiency. Since the transmitter mobile device is a portable device, it is energy constrained, so it is critical that transferring power from this mobile device to a portable device does not significantly decrease the battery life of the transmitter, which is valuable to the user. In terms of the cochlear implant use case, the user desires high efficiency so that after power transfer, both the cell phone and the implant are almost fully charged and can be used for the entire day. Another important metric is delivered power. Delivered power is related to both the time it takes to charge a portable device and the amount of energy that is wirelessly transmitted. In terms of the MP3 player use case, the user desires to charge at a high power in order to charge quickly to minimize inconvenience, but at the same time the user desires to charge at a low power to minimize radiation for safety reasons. Overall, it is crucial that systems designed to share power can do so at high efficiency with regulated power delivery.
1.4 Thesis Contributions and Organization

This work develops circuits and systems that work towards the wireless power sharing vision and that help realize the use cases described in Subsection 1.3.1. The main contributions are in the following three areas:

1. **Transmitter power amplifier control loop for high efficiency and regulated power**: Chapters 2-4 present a theoretical analysis of the wireless power system to derive a transmitter-side power amplifier control loop for increasing system efficiency and balancing delivered power over changing conditions. Chapter 2 gives an introduction to wireless power transfer. Chapter 3 describes the general control loop design, highlights the objectives it satisfies, and compares it to previous work. The main design challenge is the variability of portable-to-portable wireless charging coupling and loading conditions, and the control loop addresses this through dynamic compensation of the power amplifier components. Chapter 4 provides control loop design details. The design is motivated by a detailed analysis of the efficiency and delivered power of a conventional wireless power transfer system over changing conditions. The analysis shows that power amplifier zero-voltage switching is critical to high efficiency, and this leads to a method adopted in the control loop of dynamic feedback and adjustment of the power amplifier shunt capacitance and resonant tank series inductance to maintain zero-voltage switching while balancing delivered power. Mathematical analysis is supported by simulation results.

2. **Integrated circuit and system design for wireless charging**: Chapters 5-7 present a full design of a wireless charging system incorporating the power amplifier control loop. Chapter 5 presents the design of an ultra low-power integrated circuit in 0.18 μm HVC-MOS that implements the control loop by sensing for power amplifier zero-voltage switching and dynamically adjusting the power amplifier component values. Chapter 6 provides the design of the board-level transmitter and receiver subsystems that use the fabricated integrated circuit in a functioning resonant inductive wireless charger system operating at 6.78 MHz. Chapter 7 provides measurement results. The system achieves 78% peak end-to-end efficiency transferring 200 mW over 7 mm, and achieves over 70% end-to-end efficiency transferring about 200 mW over 4-12 mm distances.
3. Medical, sensor, and consumer applications for portable-to-portable wireless charging: Chapter 8 presents seven system applications for portable-to-portable wireless charging, from medical devices such as a fully-implantable cochlear implant to sensor devices such as a fitness tracker to niche devices such as bicycle light. Each application is developed to support wireless recharging of the associated portable device using a smartphone in 2-15 mins for a typical day’s usage. The applications showcase the wide-ranging opportunities for power sharing.

Wireless power sharing from device to device will dramatically improve users’ experiences with portable electronics and contribute to greater environmental sustainability.
Chapter 2

Basic Wireless Charging Circuits

Before detailing the new techniques in this thesis, this chapter first presents a background in wireless power circuit basics. The following chapters build on these conventional circuits to improve performance in the key user metrics of high efficiency and consistent delivered power.

2.1 Basic Wireless Charging System Block Diagram

![Figure 2.1: Block diagram of the basic wireless charging system](image)

Figure 2.1 shows a block diagram of a foundational resonant wireless wireless power system. Other types of wireless power transfer techniques exist, namely capacitive coupling power transfer (in which energy is transferred through electric fields) [17] and inductive power transfer (in which energy is transferred through magnetic coils not tuned to a given frequency) [18], but resonant wireless power transfer is the most popular. Power starts in the transmitter mobile device battery and is converted into AC power through the power amplifier. The power is then
delivered into the wireless link, which consists essentially of two coupled resonant tanks, the primary-side resonant tank in the transmitter device and the secondary-side resonant tank in the receiver device. Each resonant tank includes a capacitor in series with a matched inductor, which is implemented in this case as a discrete coil. Power is transferred through the air between these two coupled coils. The received AC power is then rectified and is available to charge the portable device battery and power a load.

### 2.2 Basic Wireless Charging Circuit

![Circuit-level diagram of the basic wireless charging system](image)

Figure 2.2: Circuit-level diagram of the basic wireless charging system

Figure 2.2 shows a circuit-level implementation of the basic wireless charging system. This circuit is used as the foundation for the system developed in this work. In the figure, the battery and the power amplifier blocks in Figure 2.1 are not shown, and instead their output is modeled as the AC source $v_{AC}$ [19]. The battery voltage $V_{CC}$ is set to be 3.3 V for the system developed in this work, since it is a common voltage derived from a lithium-ion battery. The primary-side resonant tank consists of the elements $C_1$ and $L_1$. The secondary-side resonant tank consists of the elements $C_2$ and $L_2$. The resistances $R_S$, $R_1$, and $R_2$ represent parasitic resistances of the power amplifier, primary-side coil, and secondary-side coil, respectively. Let $R_{PRI} \equiv R_S + R_1$. The transmitter and receiver coils are modeled as the inductances $L_1$ and $L_2$, respectively, with a primary-to-secondary turns ratio of 1-to-$n$.

The rectifier is a full-bridge rectifier and the load consists of an energy-storage element, the ultra-capacitor $C_{OUT}$, and some load, modeled as a resistance. An ultra-capacitor is used because it charges and discharges quickly, and it is simpler to charge compared to other energy-storage elements, such as lithium-ion batteries. In addition, in some applications such as medical implants, implanted ultra-capacitors are preferred over batteries because the former has a
lifetime of many more charge cycles than the latter (millions to thousands) [20]. The ultra-
capacitor used in the system in this work is designed to be charged from a voltage of 2.5 V to
5 V.

A few other aspects of the basic wireless charging circuit are worth describing in more
detail in the following subsections, namely the operating frequency, the coil parameters, and
the coupling coefficient.

2.2.1 Operating Frequency

The operating frequency of the system developed in this work is 6.78 MHz. In other words,
the receiver resonant tank is tuned to resonate at 6.78 MHz, or \( \frac{1}{2\pi\sqrt{L_2C_2}} = 6.78 \text{ MHz} \). The
transmitter resonant tank is tuned to a slightly different frequency, the details of which are
described later. 6.78 MHz is chosen because it is an ISM band, an unregulated frequency spec-
trum range between 6.765 MHz and 6.795 MHz. The Federal Communications Commission
permits equipment used for “industrial, scientific, medical, domestic or similar purposes, ex-
cluding applications in the field of telecommunication” to transmit in this band with unlimited
energy [21]. Other ISM bands exist, such as at 13.56 MHz and 5.8 GHz. Poon et al. show that
operating in the gigahertz frequency range leads to lower dispersion in materials such as tissue
[22]. Using a higher operating frequency allows the coils to be made smaller, but also results in
higher losses in and more difficult design of the power amplifier. The 6.78 MHz band is chosen
as an acceptable tradeoff for this work, but the techniques developed can be applied to other
operating frequencies as well.

2.2.2 Coil Sizes and Inductances

The transmitter and receiver coils of the system developed in this work are sized after a common
wireless power application: a cochlear implant. In a conventional cochlear implant, the receiver
coil is about 25 mm in diameter [23]. Thus, the receiver coil in this system is designed to be
25 mm and the transmitter coil is designed to be slightly larger, 30 mm, for better coupling
with the receiver coil. At these dimensions, both coils are designed to have 5 uH inductance,
or \( L_1 = L_2 = 5 \text{ uH} \).
2.2.3 Coupling Coefficient

The primary-side transmitter coil and the secondary-side receiver coil are separated by material that is not magnetically ideal, so the coils are not ideally coupled. A widely known metric for the coupling quality of the two coils is the coupling coefficient $k \equiv \frac{M}{\sqrt{L_1 L_2}}$, where $M$ is the mutual inductance. The mutual inductance between two one-turn coils is

$$M_{i,j} = \frac{2\mu}{\alpha} \sqrt{r_i \cdot r_j} \left[ \left( 1 - \frac{\alpha^2}{2} \right) K(\alpha) - E(\alpha) \right]$$  \hspace{1cm} (2.2.1)

$$\alpha = 2 \sqrt{\frac{r_i \cdot r_j}{(r_i + r_j)^2 + d_{ij}^2}}$$ \hspace{1cm} (2.2.2)

where $r_i$ and $r_j$ are the radii of the two turns, $\mu$ is the permeability of the medium, $d_{ij}$ is the separation, and $K(\alpha)$ and $E(\alpha)$ are the complete elliptic integrals of the first and second kind, respectively [24]. To find the total mutual inductance between two multi-turn coils, pairwise sum the individual mutual inductances between each turn of the first coil and each turn of the second coil, or $M = \sum M_{i,j}$.

![Coupling Coefficient vs. Separation for 30 mm / 25 mm Diameter Coils](image)

Figure 2.3: Coupling coefficient variation with separation for 30 mm / 25 mm coils

Figure 2.3 shows the coupling coefficient change with separation between an 11.5-turn 30 mm diameter coil and a 17-turn 25 mm diameter coil, both with 5 uH inductances, based on
2.3 Basic Wireless Charging Circuit Performance

This section analyzes the performance of the basic wireless charging circuit with respect to the key user metrics of power and efficiency. To do this, a simplified equivalent circuit is derived and analyzed.

2.3.1 Rectifier Equivalent Resistance

The impedance looking into the full-wave rectifier of Figure 2.2 from the left can be modeled as a real resistance from the perspective of power transfer. As shown in [25], this resistance has value

\[
R_L = \frac{4V_{OUT}}{I_{SEC}}
\]

(2.3.1)

Figure 2.4 shows the wireless charging circuit with equivalent output resistance.

2.3.2 Primary-Side Equivalent Resistance

At resonance, real impedances can be reflected from the secondary side to the primary side. Given that \( L_2 \) and \( C_2 \) are tuned to resonate at the operating frequency, in other words, given that \( w = \frac{1}{\sqrt{L_2C_2}} \), then at resonance, the imaginary impedances on the secondary side cancel out, and thus \( R_2 + R_L \) is the pure resistance looking into the secondary side. Reflecting this
back to the primary side, the resistance becomes

\[ R_{EQ} = \left( \frac{k}{n} \right)^2 Q_L^2 (R_L + R_2) = \frac{k^2 L_1}{C_2 (R_L + R_2)} \parallel \frac{k^2 L_1}{C_2 R_L} \]  (2.3.2)

This is the same result as derived in [19] using a transformer model. On the primary side at resonance, the impedances of \( C_1 \) and \( L_1 \) may not entirely cancel, and there may be some excess capacitance or inductance. Thus, the voltage across the primary side resonant tank may lead or lag the current through it. However, a sinusoidal component \( v_S \) of the power amplifier output \( v_{AC} \) is at the resonant frequency and in phase with the primary-side current. This is the only component that can deliver power to \( R_{EQ} \) and thus the output.

### 2.3.3 Delivered Power

![Primary-side equivalent circuit for power transfer](image)

Figure 2.5: Primary-side equivalent circuit for power transfer

Figure 2.5 shows the equivalent circuit for the basic wireless charging circuit, keeping only the component of the circuit that transfers power to the output. From [19], the average output power is

\[ P_{OUT} = V_{S,\text{rms}}^2 \left( \frac{k^2 L_1 C_2 R_L}{(k^2 L_1 + C_2 R_{PRI} (R_L + R_2))^2} \right) \]  (2.3.3)

### 2.3.4 Efficiency

Figures 2.4 and 2.5 can be used to calculate the power efficiency of the basic wireless charging circuit. This ignores the efficiency of the power amplifier and full-wave rectifier, both important contributions that are considered later. The efficiency is simply the product of the primary and secondary efficiencies, or

\[ \eta = \left( \frac{R_{EQ}}{R_{EQ} + R_{PRI}} \right) \times \left( \frac{R_L}{R_L + R_2} \right) \]  (2.3.4)
2.4 Basic Power Amplifier Circuit

The power amplifier converts the DC energy from the transmitter’s battery to AC energy to drive the transmitter coil. A Class-E power amplifier is selected for the system in this work because of its high efficiency, ideally 100%.

![Figure 2.6: Basic Class-E power amplifier circuit](image)

Figure 2.6 shows a basic Class-E power amplifier circuit. The PA draws DC current from the battery $V_{CC}$ and converts it to AC current through the resonant tank created by the capacitance $C_R$ and inductance $L_R + L_{EX}$.

![Figure 2.7: Class-E power amplifier waveforms](image)

Figure 2.7 shows waveforms of the switch drive voltage (blue trace $v_{\text{gate}}$), the switch drain voltage $v(\omega t)$ (green trace $v_{\text{switch}}$), and the resonant tank current (red trace L1). The switch drain voltage reaches zero just as the switch turns on, satisfying the zero-voltage switching (ZVS) condition that gives the power amplifier perfect ideal efficiency. The switch drain voltage also leads the resonant tank current. This is caused by the excess inductance $L_{EX}$ ($C_R$ and $L_R$ resonate at the operating frequency of the power amplifier, or $\frac{1}{2\pi\sqrt{L_RC_R}} = 6.78$ MHz). Only with this excess inductance can ZVS be achieved.
In relation to the basic wireless charging circuit of Figure 2.2, \( v_{AC} = v(\omega t) \), \( C_1 = C_R \), \( L_1 = L_R + L_{EX} \), \( R_1 + R_S = R_{PRI} \), and \( R_{EQ} \) is the reflected load impedance at resonance. At the operating frequency, the imaginary impedances of \( C_R \) and \( L_R \) cancel but the impedance of \( L_{EX} \) remains. Thus, a portion (\( v_S \) in Figures 2.5 and 2.6) of \( v(\omega t) \) at the fundamental frequency and in phase with \( i_R \) is dropped across the resonant tank resistance and a portion out of phase is dropped across the excess inductance.

### 2.5 Chapter Summary

This chapter analyzes the operation of the basic wireless power circuit with respect to the key metrics of efficiency and power. The mathematical analysis in the following chapters shows the performance of previously published systems based on this circuit and derives improvements to this basic topology.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Battery Voltage</td>
<td>3.3 V</td>
</tr>
<tr>
<td>Operating Frequency</td>
<td>6.78 MHz</td>
</tr>
<tr>
<td>Primary-Side Coil</td>
<td>5 uH, 30 mm diameter</td>
</tr>
<tr>
<td>Secondary-Side Coil</td>
<td>5 uH, 25 mm diameter</td>
</tr>
<tr>
<td>Power Amplifier</td>
<td>Class-E</td>
</tr>
<tr>
<td>Rectifier</td>
<td>Full-wave discrete rectifier</td>
</tr>
<tr>
<td>Output</td>
<td>Ultra-capacitor (2.5 V - 5 V)</td>
</tr>
</tbody>
</table>

Table 2.1: Summary of basic parameters used in this work

Table 2.1 shows a summary of the basic parameters used for the system developed in this work.
Chapter 3

Design Overview and Previous Work

3.1 Wireless Charging System Design Overview

In light of the portable-to-portable wireless charging use cases presented in Chapter 1, some critical design challenges must be addressed by the wireless charging system in this work. Design objectives are derived to meet these challenges, and an overall system design that satisfies these objectives is presented.

3.1.1 Design Challenges

The portable-to-portable wireless charging application presents a number of obstacles to achieving good performance along the key user metrics of high efficiency and consistent delivered power. These obstacles come from the fact that both the transmitter and receiver in this application are portable, so the charging environment is highly dynamic. In contrast, the wireless power applications most consumer systems are designed for are relatively static, such as charging a cell phone sitting on a charging pad. In this more static case, the associated wireless power systems can be optimized around a more constant condition.

On the other hand, the transmitter-receiver coupling in a portable-to-portable charging application may vary dramatically, since both devices may move with respect to each other. In addition, the receiver is a low-power device and charges quickly (in a couple minutes), so the output loading changes quickly as well. In the case of the system developed in this work, the receiver charges an ultra-capacitor at the output, and the load voltage of this ultra-capacitor
changes relatively quickly.

All of these changing conditions have a significant effect on the end-to-end efficiency and delivered power. Thus, the main design challenge is to achieve high efficiency and balanced delivered power across these changing coupling and load conditions.

3.1.2 Design Objectives

The wireless charging system developed in this work meets the design challenge described in the preceding subsection. There are three general objectives for the system developed in this work:

1. **Rapidly recharge a receiver portable device in about two minutes.** This enables quickly recharging a portable device on the go, which is convenient for the user. The average delivered power specification is derived from being able to charge an MP3 player in two minutes, which is one of the representative use cases highlighted in Chapter 1. From measurements of an iPod Shuffle, a common MP3 player, it draws about 13 mW of power during operation. From the use case, this device should be able to be recharged in two minutes to last 30 minutes. Thus, the required average delivered power is $13 \times \frac{30}{2} \approx 200$ mW.

2. **High end-to-end charging efficiency.** This minimizes the battery drain of the transmitting mobile device, which is important for the user since mobile device battery life is valuable. The target peak end-to-end efficiency for this system is 75%.

3. **Maintain high efficiency and balanced power levels with changing conditions.** This system must maintain high efficiency and about 200 mW delivered power levels across the range of coupling and load conditions that the portable-to-portable application may entail. For the cochlear implant and MP3 player recharging use cases described in Chapter 1, typical separation distances may be 7-12 mm. Over these distances and averaged over the full charging range, the end-to-end efficiency should be greater than 70%. The delivered power has a relatively small impact on a short charging time, so
it can be allowed to vary slightly. The delivered power level should stay within a 40% variation across operating conditions, which is typical of previous work [3].

<table>
<thead>
<tr>
<th>Design Objective</th>
<th>Target</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Delivered Power</td>
<td>200 mW</td>
</tr>
<tr>
<td>Peak Efficiency</td>
<td>75%</td>
</tr>
<tr>
<td>Operating Separation Distances</td>
<td>7-12 mm</td>
</tr>
<tr>
<td>Efficiency, Power over Distance/Load</td>
<td>&gt; 70%, &lt; 40% variation</td>
</tr>
</tbody>
</table>

Table 3.1: Design objectives for wireless charging system in this work

Table 3.1 summarizes the design objectives for this work.

### 3.1.3 Design Overview

![Figure 3.1: Adaptive wireless charging system general diagram](image)

To address the design challenges and meet the objectives, a resonant inductive wireless charging system is designed using the conventional circuit topology presented in Chapter 2 but with an added power amplifier control loop. Figure 3.1 shows the overall system design. The control loop senses the power amplifier switch drain voltage and adjusts the power amplifier shunt capacitance and resonant tank series inductance to compensate for changing conditions. The details of this control loop are described in Chapter 4.

The nominal components of the power amplifier are designed to meet the 200 mW delivered power objective. The control loop is low-power and the system components are carefully selected to achieve > 75% peak system efficiency. Dynamic compensation by the control loop allows the system to maintain high efficiency and balance delivered power as coupling and output load...
conditions change. This control loop approach is compared to previous work in the next section.

3.2 Previous Implementations of Wireless Power Systems

Many researchers have developed wireless power systems at similar power levels and with similar dimensions as those of interest. Each system is analyzed with respect to the user metrics of end-to-end efficiency at a desired delivered power level. Four implementations are considered, each with a different approach to handling changing coupling and load conditions. The first work implements no dynamic control, the second work implements power amplifier sensing and a set delay, the third work implements dynamic power amplifier duty cycle control, and the fourth work implements dynamic power amplifier supply voltage control.

3.2.1 Lo et al., IEEE Biomedical Circuits and Systems 2013 [1]

This previous work describes the design of an epi-retinal and neural prostheses SoC incorporating wireless power transfer. Due to the many electrodes required to stimulate retinal neurons for acceptable granularity, this system requires relatively high constant power, about 100 mW. A circuit is designed to transmit this power from a 39 mm diameter transmitter coil to a 19 mm diameter receiver coil, separated by distances of 1-3 cm. A standard Class-E power amplifier is used in the transmitter at an operating frequency of 2 MHz. The outputs are fixed ±2V and ±12V supplies generated with CMOS rectifiers and on-chip regulators for constant power delivery to the load.

With this system design, the end-to-end efficiency is 24% at 10 mm coil separation, and the rectifier efficiency is 72%. In addition, the end-to-end efficiency increases to a peak of 40% at 17 mm separation before declining to 15% at 30 mm separation. The supply voltage of the transmitter power amplifier is manually adjusted to regulate the delivered power for each coil separation. This thesis differs from the previous work in that this thesis presents a dynamic power amplifier control circuit that achieves higher efficiency and automatically controls power levels over changing conditions.
3.2.2 Kendir et al., IEEE Circuits and Systems 2005 [2]

This previous work describes an efficiency-based design methodology for the components in a wireless power system. The system transfers 250 mW of power from a 36 mm diameter transmit coil to a 20 mm diameter receive coil, separated by 2-15 mm. A standard Class-E power amplifier is used in the transmitter at an operating frequency of 1 MHz. The output is a fixed 16 V supply generated with a discrete half-wave rectifier. The authors derive an optimal coil design considering the coil losses. In addition, the work proposes calculating the time delay between the point at which the resonant tank current reaches zero and when the power amplifier switch drain voltage reaches zero, sensing the zero-crossing point of the resonant tank current, and waiting that precise amount of time to turn on the switch for zero-voltage switching. Switching the power amplifier at zero voltage increases the system efficiency. The wireless power system implements this precise time delay calculated at a single coil separation of 7 mm.

The efficiency of the reported system is 65% at 2 mm separation, 67% at 7 mm separation, and 51% at 15 mm separation. The efficiency is optimized at the single separation of 7 mm, while zero-voltage switching is not satisfied at other separations. In this previous work the delivered power is kept constant across different separations by manual adjustment of the transmitter power amplifier supply voltage. This thesis differs from the previous work in that this thesis presents a lower power dynamic power amplifier control circuit that automatically maintains optimal efficiency and balanced power for a range of separations.

3.2.3 Baker, Sarpeshkar, IEEE Biomedical Circuits and Systems 2007 [3]

This previous work investigates feedback circuits in wireless power systems for bionic devices. The power circuits are designed to deliver up to 10 mW of power through 30 mm diameter transmitter and receiver coils separated by 1-10 mm. The output is a fixed 3.4 V supply generated with a discrete half-wave rectifier. The paper presents an enhanced Class-E power amplifier operating at 6.78 MHz. The PA incorporates a feedback circuit that detects when the switch drain voltage has reached zero to attempt to always turn on the MOSFET under zero-voltage switching conditions, which increases system efficiency. The feedback adjusts the
duty cycle and frequency of operation so zero-voltage switching can be achieved.

With this configuration, the system achieves end-to-end efficiency of 54% at 10 mm separation and 60% at 7 mm separation. The rectifier efficiency is 96%. At a transmitter power amplifier supply voltage of 2.35 V, the output voltage varies from 5.9 V at 5 mm separation to 5 V at 10 mm separation, so the delivered power varies by \( \frac{5.9^2 - 5^2}{5^2} = 39\% \). This thesis differs from the previous work in that this thesis presents a feedback circuit for the power amplifier based on adjusting the power amplifier component values instead of adjusting its switching duty cycle and frequency. The fixed frequency and fixed 50% duty cycle of the power amplifier in this thesis limits undesired out-of-band or harmonic energy. Adjusting the power amplifier component values also allows the power amplifier to have zero-voltage switching in a wider range of cases, especially when the switch drain voltage tends to be positive at switch turn-on. Also, adjusting the power amplifier component values keeps the delivered power consistent across changing conditions in addition to optimizing the system efficiency.

3.2.4 Wang et al., IEEE Circuits and Systems 2005 [4]

This previous work develops a wireless power system delivering 250 mW to a retinal implant. The authors use a 40 mm diameter transmitter coil and a 22 mm diameter receive coil. The transmitter includes a Class-E power amplifier operating at 1 MHz. The output is a fixed 15 V supply generated with a discrete half-wave rectifier and linear regulator circuits. The wireless power system incorporates a feedback circuit that determines the output power and communicates this information back through the power link using a load-shift keying technique. The transmitter circuit then adjusts the power amplifier supply voltage to regulate the transmitted power, so that the output voltage stays around 15 V.

With this design, the system achieves end-to-end efficiency of 65.8% at 7 mm separation and 36.3% at 15 mm separation. This thesis differs from the previous work in that this thesis presents a control loop that not only regulates power but also optimizes efficiency over changing conditions.
3.3 Chapter Summary

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Objective</td>
<td>Dynamic PA capacitance / inductance</td>
<td>200 mW</td>
<td>&gt; 75%</td>
<td>&gt; 70% 7-12 mm</td>
</tr>
<tr>
<td>Lo et al.</td>
<td>No dynamic control</td>
<td>100 mW</td>
<td>40%</td>
<td>24% 10 mm</td>
</tr>
<tr>
<td>Kendir et al.</td>
<td>Resonant current zero-crossing detect</td>
<td>250 mW</td>
<td>67%</td>
<td>65% 2 mm, 51% 15 mm</td>
</tr>
<tr>
<td>Baker</td>
<td>Dynamic PA duty cycle / frequency</td>
<td>10 mW</td>
<td>60%</td>
<td>54% 10 mm</td>
</tr>
<tr>
<td>Wang et al.</td>
<td>Dynamic PA supply voltage</td>
<td>250 mW</td>
<td>65.8%</td>
<td>36.3% 15 mm</td>
</tr>
</tbody>
</table>

Table 3.2: Comparison of this system design approach to previous approaches

Table 3.2 shows a summary comparing the approach of this work to previous work. The unique control loop proposed allows the system to achieve higher peak efficiencies and higher efficiencies over changing separation distances while delivering consistent power levels. The next chapter presents the mathematical analysis detailing the proposed control loop.
Chapter 4

Power Amplifier Control Loop for High Efficiency and Balanced Power

This chapter details the power amplifier control loop described in Chapter 3 for increasing system efficiency and balancing delivered power across changing conditions. A block-level efficiency analysis is first performed to show that power amplifier and wireless link block efficiency is important. This block is then analyzed to show that power amplifier zero-voltage switching optimizes its efficiency. Finally, a method is derived for dynamically tuning the power amplifier to maintain zero-voltage switching while balancing power across changing conditions. The power amplifier control loop adopts this feedback and compensation method.

4.1 Block-Level Efficiency Breakdown

Efficiency of the system at the block-level as illustrated in Figure 2.1 is determined through SPICE simulation. The combined transmitter and receiver system is broken into two major blocks: the power amplifier and wireless link block and the rectifier block. The system is separated this way because these two blocks are generally decoupled from each other.
Figure 4.1 shows the circuit for simulation based on the wireless power circuit and parameters of Chapter 2. The setup includes a circuit-level implementation of the power amplifier, which is realized as a Class-E amplifier consisting of the components $L_{CHOKE}$, $M_1$, $M_2$, $D_5$, $C_3$, $C_1$, and $L_1$. Typical parasitic resistances on the primary and secondary sides are modeled as 1.75 ohm and 1.5 ohm resistances in series with the coils, respectively. On the secondary side, diodes $D_1$ through $D_4$ form the full-bridge rectifier. At the output, the high-capacitance ultra-capacitor is modeled as a constant voltage source. All the components are sized so the power delivered to the output is approximately 200 mW.

Simulating this circuit gives the end-to-end efficiency to be 79.4%. In particular, the efficiency of the power amplifier and wireless link combination is 91.5% and the efficiency of the rectifier is 86.7%. Thus, both of these blocks are significant contributors to the end-to-end efficiency. Considering the rectifier block, virtually all of the efficiency hit comes from the conduction loss of the Schottky diodes. Much research has been conducted in the area of CMOS rectifiers to reduce this conduction loss [26, 27, 28, 29, 30]. Thus, increasing the efficiency of the rectifier block through CMOS techniques is not the focus of this work.

Instead, this thesis focuses on the efficiency of the power amplifier and wireless link block. Also, since the delivered power of this block is directly related to the overall delivered power, this thesis focuses on keeping the block power consistent. Both of these aims are achieved by
4.2 Power Amplifier and Wireless Link Analysis Strategy

The goal of the following analysis is to identify and optimize the factors that determine efficiency and delivered power of the power amplifier and wireless link block under changing load and coupling conditions. This is done in a few steps. First, zero-voltage switching (ZVS) is theoretically shown to be the optimal efficiency condition. Next, the effect of changing conditions on the primary-side equivalent resistance is analyzed. This is used to show the effect of changing equivalent resistance on power amplifier ZVS. Two types of ZVS are analyzed: full ZVS, in which the power amplifier switch drain voltage and slope are both zero at switch turn-on, and general ZVS, in which just the voltage is zero. In either case, ZVS can be maintained across changing conditions by dynamically tuning the power amplifier shunt capacitance and resonant tank inductance. Maintaining the full ZVS condition increases efficiency, but maintaining the general ZVS condition provides the flexibility to balance the delivered power at the same time. This analysis is used to derive a method to compensate the power amplifier capacitance and inductance that maintains high block efficiency while also balancing delivered power under changing conditions. Simulation results at the end of this chapter support the mathematical analysis.

4.3 Effect of Power Amplifier Zero-Voltage Switching on Efficiency

In general, the power amplifier drain voltage waveform $v(t)$ can be in one of three cases, as shown in Figure 4.2.
Figure 4.2: Class-E power amplifier switch drain voltage waveform cases

Case 1 is the ZVS case in which the voltage reaches zero at switch turn-on. At this point, the slope can be zero (termed full ZVS) or non-zero (termed general ZVS). Case 2 is the case in which the voltage is positive at switch turn-on, at which point the switch brings the voltage quickly down to zero. Case 3 is the case in which the voltage reaches zero before switch turn-on. As the voltage becomes negative, the switch body diode clamps the voltage to a value near zero until the switch turns on. Cases 2 and 3 both have negative effects on the efficiency and delivered power of the wireless power circuit, as well as other effects.

The efficiency in Case 2 suffers from switching losses from the parasitic and discrete capacitors at the output of the switch. In addition, just after the switch turns on, there is a low resistance path through the switch for the capacitors to discharge, causing a potentially destructive instantaneous current. Because of these two reasons, Case 2 should be avoided.

In Case 3, when the voltage becomes negative, the switch body diode (or any discrete diode across the switch) starts to conduct. Since the diode has forward voltage drop, this dissipates power. Another source of loss comes from the phase shift and higher order harmonics of the voltage waveform. As the zero-crossing time gets further from the switch turn-on time, the resonant tank voltage increasingly leads the resonant tank current. Also, since the voltage waveform changes faster, the energy of higher-order harmonics increases. The out-of-phase voltage and
higher-order harmonics both do not deliver power to the load. Another interpretation of this situation is that the power factor of the power amplifier is increasingly lower than one. Thus, in order to keep the same delivered power, the input power must be increased, which causes more losses due to the parasitic resistances of the power amplifier (such as the resistances of the choke inductor and power switch). Hence, keeping the same delivered power is at the cost of decreased efficiency, or keeping the same efficiency is at the cost of decreased delivered power. Both of these situations are undesirable. In addition, since the voltage is positive for a shorter time, the peak of the voltage waveform is higher. This is because since the average voltage across the choke inductor is zero, the average of the switch voltage is always equal to the DC voltage $V_{CC}$. This higher peak causes greater stress on the power switch, and to accommodate this higher stress a power switch might have to be chosen that has more parasitics than a switch that can tolerate lower stresses. For all of these reasons, Case 3 should be avoided.

In summary, the power amplifier and wireless link block achieves maximum efficiency across conditions when ZVS is maintained. Note that even when ZVS is maintained, delivered power can still vary with changing conditions, so the system must not only maintain ZVS but also regulate power. The following analysis shows how ZVS is affected by changing conditions and shows how to maintain ZVS in these situations.

### 4.4 Effect of Changing Conditions on Equivalent Resistance

![Figure 4.3: Basic wireless power circuit](image)

Figure 4.3 shows the wireless power circuit as discussed in Chapter 2. Chapter 2 shows that the secondary side presents an equivalent load resistance $R_{EQ}$ to the primary side. Changing output voltage $V_{OUT}$ and changing coupling coefficient $k$ both affect $R_{EQ}$, given by
\[ R_{EQ} = \frac{k\sqrt{L_1L_2w} \left( \frac{\pi}{4} \frac{|V_s|}{V_{OUT}} k\sqrt{L_1L_2w} - R_{PRI} \right)}{\frac{\pi}{4} \frac{|V_s|}{V_{OUT}} R_2 + k\sqrt{L_1L_2w}} \]  \hspace{1cm} (4.4.1)

Details are given in Appendix A.

![Figure 4.4: \( R_{EQ} \) versus coupling coefficient \( k \)](41)

![Figure 4.5: \( R_{EQ} \) versus output voltage \( V_{OUT} \)](41)

Figures 4.4 and 4.5 show how \( R_{EQ} \) is affected by changing coupling and load conditions.
Intuitively, increasing the coupling coefficient or decreasing the output voltage should increase the loading on the primary side, hence the higher $R_{EQ}$.

### 4.5 Maintaining Power Amplifier Full ZVS

The goal is to maintain the power amplifier in the full ZVS condition (drain voltage and slope are zero at switch turn-on) given changing primary-side equivalent resistance $R_{EQ}$. From Section 4.3, this leads to optimal efficiency of the power amplifier and wireless link block. The Class-E power amplifier is shown in Figure 4.6. The operation is dependent on two key parameters: the shunt capacitance $C_P$ and the excess inductance $L_{EX}$. In full ZVS, $C_P$ and $L_{EX}$ are given by

\[
\frac{X_P}{R_{PRI} + R_{EQ}} = \frac{\pi(\pi^2 + 4)}{8} \approx 5.447
\]

\[
\frac{X_{EX}}{R_{PRI} + R_{EQ}} = \frac{\pi^3}{16} - \frac{\pi}{4} \approx 1.152
\]

where $X_P = \frac{1}{wC_P}$ and $X_{EX} = wL_{EX}$. Details of this derivation are in Appendix A. This shows that to maintain full ZVS with a changing $R_{EQ}$, $C_P$ and $L_{EX}$ must be adjusted accordingly to unique values. This works for optimizing circuit efficiency, but does not leave any flexibility for regulating delivered power. Thus, in the next section, a more general ZVS condition is analyzed that provides more flexibility.
4.6 Maintaining Power Amplifier General ZVS

The goal is to maintain the power amplifier in the general ZVS condition (only drain voltage is zero at switch turn-on) given changing primary-side equivalent resistance $R_{EQ}$. This provides the efficiency benefits described in Section 4.3 but adds flexibility.

In general ZVS, the power amplifier operation is governed by the following equations

$$\begin{align*}
\phi &= \tan^{-1} \left( \frac{-2V_{CC}}{\pi V_{CC} + S} \right) \\
V_S &= \frac{2S \sin(2\phi)}{\pi \sin(\phi) + 2 \cos(\phi)} \\
V_{EX} &= \frac{2S \cos(2\phi) + \left( \frac{2}{\pi} - \frac{\pi}{2} \right) S}{\pi \sin(\phi) + 2 \cos(\phi)} \\
\frac{V_{EX}}{V_S} &= \frac{wL_{EX}}{R_{PRI} + R_{EQ}} = \cot(2\phi) + \left(1 - \frac{\pi^2}{4}\right) \csc(2\phi) \\
I_R &= -\frac{V_S}{R_{PRI} + R_{EQ}} \\
X_P &= \frac{\pi S}{I_R(\pi \sin(\phi) + 2 \cos(\phi))}
\end{align*}$$

(4.6.1)

where $\phi$ is the phase shift between resonant tank current and switch drive voltage, $S$ is the non-zero drain voltage slope at switch turn-on, and $X_P = \frac{1}{wC_{P}}$. Details of this derivation are in Appendix A.

Figure 4.7: $\frac{X_P}{R_{PRI} + R_{EQ}}$ vs. $S$ in general ZVS
From these equations, Figure 4.7 is a plot of \( \frac{X_P}{R_{PRI}+R_{EQ}} \) vs. \( S \). It is evident that given a value for \( R_{EQ} \), picking different values of \( S \) results in different values of \( C_P \) that satisfy the general ZVS condition. Figure 4.8 is a plot of \( \frac{X_{EX}}{R_{PRI}+R_{EQ}} \) vs. \( S \). It is evident that given a value for \( R_{EQ} \), picking different values of \( S \) results in different values of \( L_{EX} \) that satisfy the general ZVS condition.

These two figures show not only that \( C_P \) and \( L_{EX} \) can be tuned to maintain general ZVS across coupling and load conditions, but also that one of \( L_{EX} \) or \( C_P \) is flexible. The flexibility in \( C_P \) is used to regulate delivered power at the same time as maintaining ZVS, as described in the compensation method in the next section.

### 4.7 Power Amplifier Control Loop Design

From previous subsections, it is shown that the Class-E power amplifier parameters \( L_{EX} \) and \( C_P \) are key for maintaining ZVS and their values depend on the primary-side reflected equivalent resistance \( R_{EQ} \). This section presents a method for tuning \( L_{EX} \) and \( C_P \) to maintain ZVS across changing conditions while balancing delivered power.

From Figure 4.4, \( R_{EQ} \) at a nominal coupling coefficient of 0.2 is about 28 ohms, which
corresponds to a nominal $L_{EX}$ of 0.8 $\mu$H according to Equation 4.5.2. When $R_{EQ}$ changes due to changing coupling or load, $L_{EX}$ must adjust as well. The tuning range of $L_{EX}$ determines the range of $R_{EQ}$ and thus the range of conditions that the system can handle while maintaining zero drain voltage at switch turn-on. In this work, $L_{EX}$ is chosen to be tunable from 0.7 uH to 0.9 uH. For full ZVS, this corresponds to a range of $R_{EQ}$ between 24 ohms and 32 ohms. From Figure 4.4, for a nominal output load of $V_{OUT} = 3.8$ V, this corresponds to a coupling coefficient range between 0.175 and 0.22. From Figure 2.3, ZVS is maintained for a coil separation range between about 8 mm and about 10 mm. From Figure 4.5, for a nominal coupling coefficient of $k = 0.2$, ZVS is maintained for an output voltage range between 3.4 V and 4.5 V.

From the analysis in Section 4.5 of full ZVS, the tuning range of $L_{EX}$ uniquely determines the tuning range of $C_P$. While full ZVS optimizes efficiency, the delivered power can vary greatly with changing conditions.
Figure 4.9: Received power variations in general ZVS condition

Figure 4.9 shows the received power levels of the wireless power circuit for various coupling and loading conditions. The power levels are calculated from Equation 2.3.3 and indicated by the colors of each point in the plot. Points on each curve show circuits in which general ZVS is satisfied (power amplifier switch drain voltage is zero at switch turn-on). The asterisk point on each curve shows the circuit in which full ZVS is satisfied (power amplifier switch drain voltage and slope are both zero at switch turn-on). Each curve represents a different primary coil inductance value (which can be changed by tuning $L_{EX}$), and moving up on each curve
represents increasing the shunt capacitance $C_P$.

This figure shows that if the wireless power system attempts to maintain full ZVS, then as the coupling coefficient decreases or as load voltage increases, the received power increases dramatically. Thus, it is necessary to relax the conditions on the system and instead attempt to maintain general ZVS. This gives the added flexibility of being able to tune the shunt capacitance $C_P$ to regulate delivered power without significant effect on efficiency [31]. In particular, if $C_P$ is increased significantly as the coupling decreases or the load voltage increases, then the delivered power increases less. This is intuitively apparent, because increasing $C_P$ decreases its impedance, which shunts power away from the resonant tank.

In this work, a unique value of $C_P$ is chosen for each value of $L_{EX}$ such that as conditions change within the ranges in Figure 4.9, general ZVS is maintained to increase efficiency and balance power at the same time. Specifically, Table 4.1 shows five tunable settings adopted in this work.

<table>
<thead>
<tr>
<th>Compensation Setting</th>
<th>Excess Inductance $L_{EX}$</th>
<th>Shunt Capacitance $C_P$</th>
</tr>
</thead>
<tbody>
<tr>
<td>D</td>
<td>0.7 uH</td>
<td>240 pF</td>
</tr>
<tr>
<td>C</td>
<td>0.75 uH</td>
<td>190 pF</td>
</tr>
<tr>
<td>B</td>
<td>0.8 uH</td>
<td>160 pF</td>
</tr>
<tr>
<td>A</td>
<td>0.85 uH</td>
<td>140 pF</td>
</tr>
<tr>
<td>Z</td>
<td>0.9 uH</td>
<td>130 pF</td>
</tr>
</tbody>
</table>

Table 4.1: Tunable excess inductance and shunt capacitance values used in design

The aim of this work is to design a closed-loop control system for the basic Class-E power amplifier that senses the switch voltage at the turn-on time and determines if the power amplifier is in a ZVS condition. If not, the system adjusts the excess inductance $L_{EX}$ and shunt capacitance $C_P$ to bring the power amplifier closer to ZVS while keeping the power consistent.

### 4.8 Meeting the Design Objectives

This section shows that the wireless power circuit with Class-E control designed in the previous sections satisfies the design objectives presented in Chapter 3.

Combining the equations for efficiency and primary-side equivalent resistance from Subsection 2.3.4 and Section 4.4, the nominal efficiency of the power amplifier and wireless link block
\[ \eta = \frac{\pi |V_s| k \sqrt{L_1 L_2 w - R_{PRI}}}{\frac{\pi}{4} |V_{OUT}| k \sqrt{L_1 L_2 w + \left(\frac{\pi}{4} |V_{OUT}|\right)^2 R_2}} \]  

(4.8.1)

Using the basic parameters in Chapter 2 and Section 4.1, and the nominal values \( k = 0.2 \) and \( V_{OUT} = 3.8 \text{ V} \) (typical for the wireless charging application), the nominal efficiency is \( \eta = 92.0\% \). This value matches the simulation results of the same circuit in Section 4.1 very closely. A typical rectifier efficiency is 90\%, so the end-to-end efficiency is about \( 0.92 \times 0.90 = 83\% \). The Class-E control loop is low-power and does not significantly affect this efficiency. This meets the design objective of 75\%.

From Equation 2.3.3, the average received power is \( P_{OUT} = 189 \text{ mW} \). This power is delivered to the rectifier, which has a typical efficiency of 90\%, so the delivered power to the load is nominally \( 189 \times 0.9 = 170 \text{ mW} \). This is relatively close to the design objective of 200 mW.

The efficiency and power specifications over variations in coupling and load are shown to be met in the simulation results in the next section.

### 4.9 Simulation Results

This section aims to show that the power amplifier control method derived in Section 4.7 indeed improves the efficiency and balances the delivered power of a wireless charging system over changing conditions.

The simulation circuit is the same as the one used in Section 4.1. To implement the control method of Section 4.7, the tunable parameters are the Class-E power amplifier resonant tank inductance \( L_1 \) (which includes \( L_{EX} \)) and shunt capacitance \( C_P \) (including the discrete capacitance \( C_3 \) and about 100 pF of parasitic switch and diode output capacitance). The coupling and load parameters that are changed are the coupling coefficient \( k \) and the output voltage \( V_{OUT} \), respectively.

Two simulations are performed. First, simulations show that changing coupling and load both affect the switch drain voltage at switch turn-on, and that compensating the power amplifier brings the power amplifier back towards the ZVS condition. One representative case is shown, and other cases are presented in Appendix A. Second, simulations show that the control
method used causes the overall system to have generally higher efficiency and more balanced
delivered power over changing conditions than two fixed systems tested.

4.9.1 Effects of Changing Conditions and Compensation on ZVS

![Switch drain voltage for L1 = 5 uH, CP = 80 pF system, k = 0.2, VOUT = 3.8 V conditions](image)

Figure 4.10 shows the switch drain voltage waveform for the circuit in Figure 4.1. The power amplifier is in the ZVS condition.

![Switch drain voltage for L1 = 5 uH, CP = 80 pF system, k = 0.15, VOUT = 3.8 V conditions](image)

Figure 4.11: Switch drain voltage for \( L_1 = 5 \) uH, \( CP = 80 \) pF system, \( k = 0.15 \), \( V_{OUT} = 3.8 \) V conditions

With the same power amplifier parameters, \( k \) is decreased to 0.15. This affects the ZVS condition as shown in Figure 4.11. Mathematically, \( R_{EQ} \) is decreased (the primary side sees less loading), so the resonant current phase shift \(|\phi|\) is increased. In this case, the dynamic system compensates by decreasing \( L_1 \) and increasing \( C_P \).
Figure 4.12: Switch drain voltage for $L_1 = 4.9$ uH, $C_P = 160$ pF system, $k = 0.15$, $V_{OUT} = 3.8$ V conditions

Figure 4.12 shows the compensated system, which is now much closer to the ZVS condition. The feedback and compensation method maintains ZVS over a variety of other coupling and load changes, as shown in Appendix A.
4.9.2 End-to-End Efficiency and Delivered Power of Dynamic vs. Fixed Systems

Figure 4.13: Simulated end-to-end efficiency and delivered power of dynamic vs. fixed systems

Figure 4.13 shows the simulation results for the end-to-end efficiency and delivered power over coupling and load variations of three systems: the dynamic system using the derived power amplifier control method and two fixed systems. The fixed systems have $L_1$ and $C_P$ values that result in zero-voltage switching at a particular coupling and loading condition. In particular,
one system has $L_1 = 4.9$ uH and $C_P = 88$ pF, while the other system has $L_1 = 5.1$ uH and $C_P = 48$ pF. The delivered power levels are those seen at the load, and they are represented by the color of each data point in the plot.

The dynamic system has clear advantages over the fixed systems. First, for low coupling and high output voltage conditions, the efficiency and delivered power of the 5.1 uH fixed system drops drastically due to the power factor of the circuit being increasingly less than one, while the dynamic system follows the 4.9 uH fixed system curves. Second, for high coupling conditions, the efficiency of the 4.9 uH fixed system drops significantly due to losses caused by output capacitor discharge at switch turn-on, while the dynamic system follows the 5.1 uH fixed system curve. Third, for relatively low coupling and relatively high output voltage conditions, the delivered power of the 4.9 uH fixed system increases drastically due to the relatively low shunt capacitance and high Q, while the dynamic system delivered power is balanced between the 4.9 uH and 5.1 uH fixed system delivered power levels. Essentially, the power amplifier control method adjusts the system to match the performance of the best fixed system for a given coupling and load condition.

4.10 Summary of Theoretical Analysis of Design for Efficiency and Power

The last three chapters present a theoretical analysis of the power amplifier control loop for the wireless charging system in this work. The design enables the system to achieve over 75% nominal end-to-end efficiency transferring $\sim 170$ mW over 9 mm. The design also enables the system to maintain this high efficiency and regulate the delivered power as the coupling coefficient and output voltage conditions change. The system does this by actively monitoring the Class-E power amplifier switch drain voltage at the time of switch turn-on to check if the power amplifier is in the zero-voltage switching condition. If not, the system compensates by adjusting the primary coil inductance and shunt capacitance to bring the power amplifier back towards the zero-voltage switching condition while maintaining the delivered power level.

The following chapters present the design and implementation of this enhanced wireless charging system, starting with IC design and continuing with PCB system design.
Chapter 5

Adaptive Wireless Charging Integrated Circuit Design

The adaptive wireless charging system implements the control technique derived in the previous chapter using a custom controller microchip fabricated in 0.18 μm HVCMOS. This chapter focuses on the integrated circuit design.

5.1 Wireless Charging Controller Microchip Block Design

![Diagram](image)

Figure 5.1: Wireless charging controller microchip block diagram
Figure 5.1 shows a block diagram of the wireless charging controller microchip. The chip is broken up into five main parts: the power amplifier switch drive buffer, the pulse generator block, the analog block, the digital block, and the switch block.

The power amplifier switch drive buffer takes in an external clock signal and buffers it on chip to drive the external power amplifier switch gate.

The pulse generator block takes in the same external clock signal and extracts a single pulse about every millisecond to trigger the analog and digital blocks. Thus, the adaptive control activates about every millisecond. This is fast enough to allow the system to react to changing conditions rapidly and also slow enough to ensure stability.

The input signal, which is the power amplifier switch drain voltage, flows through the analog, digital, and switch blocks in series to the output. The analog block determines if the drain voltage at switch turn-on is sufficiently close to zero. Then the digital block determines if compensation of the power amplifier is needed. If so, the switch block handles adjusting the tunable power amplifier shunt capacitance and resonant tank inductance.

The following sections detail the design of each of the blocks on the chip.

### 5.2 Microchip Power Amplifier Switch Drive Buffer Design

The power amplifier switch drive buffer drives the gate of the external power amplifier switch. The input is an externally generated 3.3 V 6.78 MHz clock that is level shifted to 1.8 V on chip. This level-shifted fast clock signal $FCLK$ is buffered through an inverter chain to drive the $\sim 100$ pF input capacitance of the power amplifier switch. $FCLK$ is also routed as an input to the pulse generator block.

Due to the inverter chain delay, the power amplifier switch drive signal $DRIVE$ is a delayed version of $FCLK$. In simulation, the delay from the rising edge of $FCLK$ to the moment $DRIVE$ starts to rise is 2.1 ns in the fastest case. The delay from the rising edge of $FCLK$ to the point $DRIVE$ is 67% of its final value is 8.2 ns in the fastest case.
5.3 Microchip Pulse Generator Block Design

The pulse generator block supplies the pulse signal that activates the analog and digital blocks to sample and process the power amplifier drain voltage. The input is the externally generated level-shifted fast clock signal $FCLK$. The pulse generator block extracts exactly one pulse from this clock every interval according to a slow clock signal generated by an on-chip 1.8 V $\sim$ 1 kHz ring oscillator. The designs of the ring oscillator and pulse extractor are described in the following subsections.

5.3.1 Ring Oscillator Design

Figure 5.2: Ring oscillator schematic diagram

Figure 5.2 shows the schematic of the on-chip 1.8 V $\sim$ 1 kHz ring oscillator, which is implemented as a ring of three inverters with complementary inputs and outputs operating in sub-threshold. One inverter output is buffered to generate the slow clock $SCLK$ that is used by the pulse extractor.

Figure 5.3: Leakage-based inverter schematic diagram

Figure 5.3 shows the schematic of the inverter. This inverter is based on the design in [32]. In essence, the inverter uses leakage currents instead of active currents to switch states in one
direction. Compared to other inverter and delay circuits, such as standard inverter chains and current-starved inverters, this leakage-based inverter consumes very low power.

The inverter has complementary inputs and outputs $R/RB$ and $Q/QB$, respectively. The low-to-high transition of $R$ pulls the outputs $Q$ and $QB$ to zero and $VDD$, respectively. The high-to-low transition of $R$ causes leakage current through M4 to slowly discharge $QB$ and leakage current through M3 to slowly charge $Q$. Positive feedback between these two transistors eventually drives $Q$ and $QB$ to $VDD$ and zero, respectively.

In simulation, the ring oscillator has widely varying frequencies around 1 kHz depending on process corner, due to the large variation in leakage currents across corners. The variation is tolerable, however, because the ring oscillator frequency is not critical. In particular, the ring oscillator frequency does not impact the overall system efficiency because this frequency is much lower than the operating frequency. The output $SCLK$ of the ring oscillator goes into the pulse extractor.

### 5.3.2 Pulse Extractor Design

![Figure 5.4: Pulse extractor schematic diagram](image)

Figure 5.4 shows the schematic for the pulse extractor. The output $CLK$ is a single pulse at the $SCLK$ frequency that goes high aligned with the rising edge of $FCLK$ two $FCLK$ edges after the rising edge of $SCLK$, and goes low aligned with the following rising edge of $FCLK$. The flip-flop chain synchronizes $SCLK$ and $FCLK$. The output $CLK$ pulse triggers the analog and digital blocks of the chip.

In simulation, the delay between the rising edge of $FCLK$ and the rising edge of $CLK$ is less than 1 ns across process corners.
5.4 Microchip Analog Block Design

The analog block in the wireless charging control chip consists of two micropower clocked comparators that sample the power amplifier switch drain voltage at the switch turn-on time to determine if this voltage is sufficiently close to zero, specifically within a specified range from zero to $V_{REF}$.

Both comparators are clocked with the pulse $CLK$ (and its complement $CLKB$) generated by the pulse generator block. Note that the rising edge of $CLK$ is aligned with the rising edge of the power amplifier gate drive signal that turns on the switch.

5.4.1 Zero-Volt Comparator Design

![Zero-volt comparator circuit design](image)

Figure 5.5: Zero-volt comparator circuit design

Figure 5.5 shows the circuit design for the comparator that compares the input voltage to zero volts. This circuit is derived from the design in [33]. On the rising edge of $CLK$, the circuit compares the current between the leftmost two branches, with positive feedback to decrease the decision time. For example, if the input voltage $IN$ is less than zero, then M3 in the right branch draws more current than M2, pulling node $Z$ to ground and driving the outputs $ABV0$ and $BLW0$ to zero and $VDD$, respectively. Positive feedback among the outputs, M5, and M6 decrease the decision time. Bias currents are minimized to lower power consumption.

In simulation, for input differential voltages down to 30 mV, the output is within 100 mV of the correct value within 3 ns across process corners. The output signals feed into the digital block logic, as described in a later section.
5.4.2 $V_{REF}$ Comparator Design

![Comparator Circuit Diagram]

Figure 5.6: $V_{REF}$ comparator circuit design

Figure 5.6 shows the schematic for the comparator that compares the input voltage to a positive voltage $V_{REF}$. The topology is a standard Strong-Arm comparator topology. On the rising edge of $CLK$, the circuit essentially compares the current in the two branches of the circuit, with positive feedback to decrease the decision time.

In simulation, for input differential voltages down to 30 mV, the output is within 100 mV of the correct value within 3 ns across process corners. The power consumption of the comparator is also very low since it is clocked. The output signals feed into the digital block logic, as described in the next section.

5.5 Microchip Digital Block Design

The digital block of the wireless charging control chip consists of digital logic that processes the outputs of the dual comparator to determine whether compensation of the power amplifier is required. There are two main logic sections. The first section determines whether the power amplifier switch drain voltage is within, above, or below the specified range of zero to $V_{REF}$. The second section determines based on the previous settings of the tunable shunt capacitor bank and tunable multi-tapped inductor what the new settings should be.

The digital logic sections are clocked by a signal derived from $CLK$ used to clock the comparators in the analog block. The first section is clocked by $CLKD$, which is a delayed
version of $CLK$ to account for the analog block propagation delay. The second section is clocked by $CLKD2$, which is a delayed version of $CLKD$ to account for the propagation delay of the first digital section.

5.5.1 Range-Determination Digital Logic Design

Figure 5.7: Range-determination digital logic schematic

Figure 5.7 shows the schematic for the first digital logic section that takes as input the outputs of the analog block and determines whether the power amplifier switch voltage is within or outside the specified range. The inputs are clocked in by $CLKD$. The output signals $IN\_RNG$, $ABV\_RNG$, and $BLW\_RNG$ are inverted indications of whether the switch voltage is in range, above range, or below range, respectively.

5.5.2 Compensation-Determination Digital Logic Design

Figure 5.8: Compensation-determination digital logic schematic for setting A
Figure 5.8 shows the schematic for a part of the second digital logic section that takes as input the outputs of the first digital logic section as well as the previous settings of both the shunt capacitor and multi-tapped inductor, and determines how to update these settings. As detailed in Section 4.7, there are five settings corresponding to five combinations of power amplifier shunt capacitance and resonant tank inductance, as shown in Table 5.1 from highest to lowest (highest to lowest capacitance). At each rising edge of $CLKD2$, exactly one setting out of five is chosen.

<table>
<thead>
<tr>
<th>Compensation Setting</th>
<th>Excess Inductance $L_{EX}$</th>
<th>Shunt Capacitance $C_P$</th>
</tr>
</thead>
<tbody>
<tr>
<td>D</td>
<td>0.7 uH</td>
<td>240 pF</td>
</tr>
<tr>
<td>C</td>
<td>0.75 uH</td>
<td>190 pF</td>
</tr>
<tr>
<td>B</td>
<td>0.8 uH</td>
<td>160 pF</td>
</tr>
<tr>
<td>A</td>
<td>0.85 uH</td>
<td>140 pF</td>
</tr>
<tr>
<td>Z</td>
<td>0.9 uH</td>
<td>130 pF</td>
</tr>
</tbody>
</table>

Table 5.1: Tunable excess inductance and shunt capacitance values used in design

The schematic in Figure 5.8 shows the circuit to choose setting A. The inputs to this circuit are $IN\_RNG$, $ABV\_RNG$, and $BLW\_RNG$ from the first digital logic block, and $Z$, $A$, and $B$, with information on whether the previous setting was $Z$, $A$, or $B$.

In general, this digital logic chooses a given setting if one of the following conditions is true:

- The setting was the previous setting and the switch voltage is in range
- The previous setting was one setting lower and the switch voltage is below range
- The previous setting was one setting higher and the switch voltage is above range

The logic follows this format for choosing any of the settings from A to D. Note that in the logic for choosing D, the third condition does not exist. Instead, the third condition is that if the switch voltage is below range and the previous setting was D, the new setting is still D.

![Figure 5.9: Compensation-determination digital logic schematic for setting Z](image)
The logic for choosing setting Z is slightly different from that to choose settings A-D. As shown in Figure 5.9, setting Z for the shunt capacitance and resonant tank inductance is chosen if none of other settings A-D is chosen. This is the default compensation setting that is chosen when the microchip is first powered on and during a reset.

5.6 Microchip Timing

Figure 5.10: Drive buffer, pulse generator, analog block, and digital block timing diagram

Figure 5.10 shows an example timing diagram of the signals in the chip. The clocks are derived from a 3.3 V off-chip 6.78 MHz oscillator that feeds into the chip. Around the rising edge of \( CLK \), the first signal to go high is the off-chip oscillator output. After a delay due to the level-shifting circuits, \( FCLK \) goes high. Then, after a delay due to the pulse extraction circuits, \( CLK \) goes high and the comparators sample \( v(wt) \). Then, the power amplifier switch gate goes high and the switch brings \( v(wt) \) to zero. Then, \( CLKD \) goes high and the first logic block output becomes valid. Then, \( CLKD2 \) goes high and the second logic block output becomes valid to select a new compensation setting. In simulation, the delay between an \( FCLK \) transition and
transition of a new setting is about 8 ns.

In this example, on the rising edge of $CLKD$, $v(wt)$ is below range as sampled on the rising edge of $CLK$, so $BLW\_RNG$ goes low. On the rising edge of $CLKD2$, since the previous setting was B, the second logic block adjusts the new setting to C.

One critical timing constraint is that the rising edge of $CLK$ must come before the rising edge of the power switch gate drive. This is so the comparators in the analog block can correctly sample the switch voltage before the switch turns on. The other critical timing constraints are that the delay from $CLK$ to $CLKD$ and the delay from $CLKD$ to $CLKD2$ must be greater than the propagation delay of the analog block and the first digital logic block section, respectively. These constraints are ensured through simulation across corners.

5.7 Microchip Switch Block Design

The switch block circuits implement the setting chosen by the digital logic block by controlling the tuning of the power amplifier shunt capacitance and resonant inductance. The shunt capacitance is made up of a bank of five capacitors while the resonant tank inductance is implemented as a multi-tapped coil with five taps.

\[ A \rightarrow B \rightarrow C \rightarrow D \]

\[ v(wt) \]

Figure 5.11: Tunable power amplifier shunt capacitance schematic diagram
Figures 5.11 and 5.12 show the schematic for the tunable shunt capacitance and resonant inductance. The components below the dashed line in each figure are on chip, while the components above the dashed line are off chip.

The fixed capacitor in Figure 5.11 has capacitance corresponding to the setting Z in Table 5-1, minus the parasitic output capacitance of the power amplifier switch and diode (about 100 pF in simulation). Each of the other switchable capacitors in Figure 5.11 are sized such that when the capacitor for a given setting is switched into the circuit, the total shunt capacitance is the correct value as specified in Table 5.1.

On the chip, four high-voltage power MOSFETs control switching each of the capacitors for settings A-D into and out of the circuit. The switches are sized proportional to the capacitance of the associated off-chip capacitors, so that the off-state capacitance and on-state resistance are proportional to the associated capacitance.

The taps on the multi-tapped resonant inductor, as shown in Figure 5.12, are controlled by off-chip solid-state relays. This design is similar to that reported in [19]. Relays are used instead of MOSFETs due to their ability to block the AC voltage and current present in the inductor. The inductance between each tap is 50 nH, so the chip can tune the resonant inductance in increments of 50 nH between 4.9 µH and 5.1 µH according to Table 5.1. On the chip, five identical high-voltage power MOSFETs turn on and off the control currents for the relays.
5.8 Chapter Summary

This chapter presents the design of an integrated circuit for implementing the power amplifier control loop derived from mathematical analysis. At a rate determined by the pulse generator block, the chip samples the power amplifier switch drain voltage at switch turn-on and compares it to a specified range between 0 and $V_{REF}$ to determine if the power amplifier is in the zero-voltage switching condition. If not, the chip changes the compensation setting, determined by the digital block and implemented by the switch block, to bring the power amplifier back towards ZVS while maintaining power levels. The next chapter details how this chip is used in a full wireless charging system.
Chapter 6

Adaptive Wireless Charging System Design

This chapter describes the PCB design incorporating the microchip from Chapter 5 into a complete adaptive portable-to-portable resonant inductive wireless charging system operating at 6.78 MHz.

![Full wireless charging system diagram](image)

Figure 6.1 shows the full wireless charging system including the wireless power circuit of Chapter 2, the wireless charging control chip of Chapter 5, and accessory circuits for the chip.
Energy is transferred from the 3.3 V source in the transmitter device to the ultra-capacitor in the receiver device, which powers the load. The following sections describe the implementation of each of the blocks in the wireless charging system, with focus on the power loss of each of the components.

6.1 Wireless Power Circuit Implementation

6.1.1 Class-E Power Amplifier Implementation

The Class-E power amplifier components $L_{CHOKE}$, M1, D1, and $C_P$ are selected to maximize nominal efficiency of the power amplifier.

The choke inductor maintains constant current from the battery to the power amplifier, and thus must have high inductance with low conduction loss. $L_{CHOKE}$ is chosen to be 22 $\mu$H with a DC resistance of 100 m$\Omega$. For 200 mW nominal power drawn from the source, the constant current flowing through the choke inductor is $200/3.3 \approx 60$ mA, so the average power loss is $0.06^2 \times 0.1 \approx 0.4$ mW, which is negligible compared to the delivered power.

The MOSFET M1 switches at the 6.78 MHz operating frequency to generate the AC current to drive the primary resonant tank. It must have low input capacitance and low on-resistance for low switching and conduction losses. A parallel combination of two Si1012CR MOSFETs is chosen that optimizes this tradeoff. M1 has a total input capacitance of 100 pF and an on-resistance of 0.4 $\Omega$ (each MOSFET) at a drive strength of 1.8 V. Thus, the combined MOSFET M1 has a switching loss of $100 \times 10^{-12} \times 1.8^2 \times 6.78 \times 10^6 \approx 2.2$ mW. From simulation, M1 has a conduction loss of 3.2 mW.

The purpose of the shunt diode D1 is to clamp the switch drain voltage $v(ut)$ at a voltage very close to zero, to prevent the body diode of M1 from turning on. This avoids loss because the body diode ordinarily has a high forward voltage. The Schottky diode DB2J209, with a forward voltage of 0.28 V at 60 mA forward current, is chosen. Since the forward voltage is small and little current flows through the diode, the power loss due to D1 is negligible.

The shunt capacitors $C_P$ are originally sized according to the design values in Table 5.1, but in testing the capacitances are slightly modified to optimize performance.
Table 6.1: Shunt capacitance values used in testing

<table>
<thead>
<tr>
<th>Capacitor</th>
<th>Capacitance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fixed</td>
<td>27 pF</td>
</tr>
<tr>
<td>Setting A</td>
<td>27 pF</td>
</tr>
<tr>
<td>Setting B</td>
<td>47 pF</td>
</tr>
<tr>
<td>Setting C</td>
<td>82 pF</td>
</tr>
<tr>
<td>Setting D</td>
<td>100 pF</td>
</tr>
</tbody>
</table>

Table 6.1 shows the final capacitance values used in implementation. These capacitors must have low parasitic resistance at the operating frequency to have low loss, so high-Q capacitors are selected. These have negligible loss.

6.1.2 Wireless Link Implementation

The wireless link consists of the primary and secondary resonant tanks $C_1$, $L_1$, $C_2$, and $L_2$. The capacitors in both resonant tanks must have low parasitic resistance at the operating frequency to have low loss, so high-Q capacitors are selected. In testing, a 127 pF primary-side capacitance and 116 pF secondary-side capacitance are selected. These have negligible loss.

The primary-side and secondary-side inductors are implemented as transmit and receive coils, respectively.

![Figure 6.2: Transmit and receive coil design](image)

Both coils are designed and simulated in the EM simulator software IE3D. The nominal inductance of each coil is designed to be 5 $\mu$H. The trace width and inter-trace spacing of each
coil are varied to determine their effects on coil inductance, which is found to be small. The trace width and trace thickness have effects on the coil ESR, however, which influences the power loss of the coils. Both coils are designed with a 0.3 mm trace width, 2-oz copper trace thickness, and 0.2 mm inter-trace spacing. This results in coil ESR of about 1.5 Ω each. In simulation the primary-side coil power loss is about 10 mW and the secondary-side coil power loss is about 5 mW. Both of these losses are major contributors to the nominal efficiency and are considered in the nominal efficiency calculation in Section 4.8.

The secondary coil is a fixed inductance and is designed with 17 turns for a nominal inductance of 5 µH. In testing the inductance is found to be 4.8 µH. The primary coil is a five-tap coil and is designed with 11.5 turns for a nominal inductance of 5 µH for the middle tap and 50 nH inductance difference between taps. In testing the nominal inductance is found to be 5.1 µH.

The relays to switch in the taps must have low on-resistance and off-capacitance for low power loss when on and low parasitic effects when off. The G3VM-21LR11 relay is chosen to optimize this RxC tradeoff. Each relay has an on-resistance of 0.27 Ω at 1.7 mA drive current and an off-capacitance of 40 pF. The on-resistance is much lower than the ESR of the in-series primary-side coil, so the power loss when on is negligible. The off-capacitance is much lower than the capacitance of the in-series resonant capacitance, so the off-state parasitic effects are negligible.

As shown in Figure 5.12, the control current for each relay is set by a resistor to 1.7 mA to balance the on-resistance of the relay with the power dissipation of the control circuit. This power dissipation is 0.0017 × 1.8 ≈ 3 mW. A 10 uF capacitor in parallel with the resistor allows high in-rush current when the relay turns on so that it may turn on faster. The turn-on time of each relay is 20 µs while the turn-off time is 100 µs. Note that the turn-on time is less than the turn-off time so that there is never a time when no relays are on (in that case, the current through the primary-side inductor would not have a path to flow).

6.1.3 Rectifier Implementation

The full-wave rectifier consists of four discrete Schottky diodes. These diodes must have low forward voltage for low power loss when on. The average diode current for an output power of
200 mW and a typical output voltage of 3.8 V is about 53 mA. The PMEG2005AELD diode is chosen, with a forward voltage of 0.18 V at this current. Thus, the average power loss for each diode is $0.053 \times 0.18 \approx 10$ mW. Since two diodes are on at any given time, the total power loss is about 20 mW, for a rectifier efficiency of about 90%.

Lower power loss may be achieved by using a CMOS rectifier, by replacing the discrete diode devices with MOSFETs with lower conduction voltage drops. This is not explored in the tested system but is the subject of extensive research [26, 27, 28, 29, 30].

### 6.1.4 Load Implementation

The rectifier is loaded by an ultra-capacitor that charges up from 2.5 V to 5 V as it receives power. The ultra-capacitor must have low ESR for low power loss when charging. For testing, a 3 F 5 V ultra-capacitor is selected with an ESR of 75 mΩ. The power loss due to such a low ESR is negligible.

### 6.1.5 Wireless Power Circuit Implementation Summary

<table>
<thead>
<tr>
<th>Component</th>
<th>Implementation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Choke Inductor</td>
<td>744053220 power inductor</td>
</tr>
<tr>
<td>M1</td>
<td>Si1012CR NMOS (2 in parallel)</td>
</tr>
<tr>
<td>D1</td>
<td>DB2J209 Schottky diode</td>
</tr>
<tr>
<td>Shunt Capacitors</td>
<td>Johanson Technology S-Series high-Q capacitors</td>
</tr>
<tr>
<td>Resonant Capacitors</td>
<td>Johanson Technology S-Series high-Q capacitors</td>
</tr>
<tr>
<td>Rectifier Diodes</td>
<td>PMEG2005AELD Schottky diodes</td>
</tr>
<tr>
<td>Output Storage Capacitor</td>
<td>PHV-5R4H305-R ultra-capacitor</td>
</tr>
</tbody>
</table>

Table 6.2: Wireless power circuit components used in testing

Table 6.2 summarizes the components used in the wireless power circuit.

### 6.2 Wireless Charging Control Chip Implementation

The wireless charging control chip described in Chapter 5 is packaged in a 48-lead TQFP package. The supplies, inputs, and outputs of the packaged chip are described in the following subsections. A chip pinout is included in Appendix B.
6.2.1 Chip Supplies

The voltage supplies for the chip include a high-voltage 3.3 V supply $V_{CC}$ for driving the power switches of the switch block and a low-voltage 1.8 V supply $V_{DD}$ for the on-chip analog and digital logic.

6.2.2 Chip Inputs

The three main signal inputs to the chip are the analog input $IN$ for the switch voltage $v(ut)$, the reference voltage $V_{REF}$, and the external 6.78 MHz oscillator clock $OSC$. The analog input is generated from $v(ut)$ by passing this voltage through a protection MOSFET M2 and diode D2. These two components are designed to restrict the analog input voltage to a maximum of $V_{CC} = 3.3$ V, since the on-chip input transistors for the analog input are rated for 3.3 V. The MOSFET must have short turn-on and turn-off times and the diode must have low forward drop for effective clamping. The Si1062X MOSFET is chosen to be M2, with a typical turn-on and turn-off time of 2 and 16 ns, respectively. The PMEG2005AE LD Schottky diode is chosen to be D2, with less than a 50 mV forward voltage drop at the microamp-level input currents of the analog input.

The reference voltage $V_{REF}$ is generated by an external resistor divider sourced from $V_{CC}$. The clock signal is generated from an external silicon oscillator.

6.2.3 Chip Outputs

The chip drives the main power amplifier switch M1 with a 1.8 V clock signal at 6.78 MHz. The chip also includes several control outputs for adjusting the power amplifier, four for switching the $C_P$ shunt capacitor bank and five for switching the $L_1$ transmit coil tap relays.

6.3 Accessory Circuit Implementation

6.3.1 1.8 V $V_{DD}$ Buck Converter Implementation

A 3.3 V-to-1.8 V buck converter is chosen to supply to the on-chip analog and digital circuits and the relay control circuits. Together, these circuits draw about 1.7 mA on average (dominated
by the relay control current). The buck converter must be efficient at this output current to minimize power loss. The TPS62202 converter is chosen, with 92% efficiency.

### 6.3.2 6.78 MHz Oscillator Implementation

The on-board 6.78 MHz oscillator supplies the clock signal for the wireless charging controller chip. The oscillator must be efficient at the 3.3 V supply voltage $V_{CC}$. The LTC6900 is chosen, with a supply current of 0.9 mA.

It is important that the oscillator starts up only after one of the relays in the wireless power circuit is on. Otherwise, the power amplifier builds up current in the choke inductor that has no path to flow when the power amplifier switch is off. Since the oscillator runs off $V_{CC}$ while the relays run off $V_{DD}$, which is derived from $V_{CC}$ through a buck converter, there is a period of time when $V_{CC}$ is stable but $V_{DD}$ is not stable. To avoid the associated race condition, the AP2280-2 power switch is added between $V_{CC}$ and the supply input of the oscillator to turn on the oscillator only after $V_{DD}$ is stable for 1 ms. This allows a relay to fully turn on before the oscillator starts up.

### 6.4 Board Design Considerations

In this work, wireless charging systems are designed with two different transmitters. The first transmitter is for testing purposes, with all signals routed to test points, focusing on ease of measurement. The 3.3 V supply $V_{CC}$ for the test transmitter is sourced from an external power supply. The second transmitter is the prototype, focusing on a minimized form factor. The prototype transmitter board includes a 30-pin connector to plug into an iPhone, and $V_{CC}$ for the transmitter is sourced from the cell phone battery (specifically, on pin 18 of the 30-pin connector).

For all transmitter and receiver board designs, a number of layout considerations are important for proper operation. These considerations are described in the following subsections.
6.4.1 Isolation to Transmit and Receive Coils

It is important for the transmit and receive coils to be physically isolated from surrounding planes, traces, and circuitry. This is because a significant magnetic field is generated or received by these coils, and this field can couple to other metal structures. As a result, the operation of these structures can be impacted, or these structures can shield the magnetic field.

Metal planes, which are used in other areas of the PCB, are absent underneath the transmit and receive coils. Traces supplying the control current to the relays for the taps on the transmit coil are located as far away from the coil as possible. Sensitive circuitry such as the custom chip and the oscillator are located as far away from the transmit coil as possible.

6.4.2 Undesired Inductance of Transmit and Receive Coils

The primary-side transmit and secondary-side receive coils are designed in simulation to have a specific inductance. A specific series capacitance is matched to each coil so that each coil resonates at a desired frequency. These resonant frequencies are significantly impacted by undesired inductances of the traces leading to these coils. Thus, these trace inductances should be minimized. To do this, the traces leading to the coils are made as short and wide as possible, since inductance is proportional to trace length and inversely proportional to trace width.

Another source of undesired inductance is a wide open current loop. The main high-current loops in the transmitter and receiver contain the currents going through the transmit and receive coils. These loops are made as small as possible by decreasing trace lengths in these loops through the use of ground planes, and by designing return current paths in a different layer just underneath forward current paths.

A third source of undesired inductance is the shape of the routing to the transmit coil. The routing to each of the taps of this multi-tapped coil affects the flux linkage. To keep the inductance differential consistent between consecutive taps, a consistent radially symmetric routing shape is implemented.
Figure 6.3 shows the layout of the trace routing to the primary-side transmit coil.

6.5 Chapter Summary

This chapter presents the design of a complete wireless charging system using the microchip described in Chapter 5. The standard wireless power topology of Chapter 2 is used with the microchip implementing closed-loop compensation of the Class-E power amplifier. The next chapter presents measured performance results of this system with respect to the key metrics of efficiency and power.
Chapter 7

Implementation and Measurement

Results

This chapter shows implementation details and reports measurement results of the portable-to-portable wireless charging system.

7.1 Chip Implementation

Figure 7.1: Chip die photo
Figure 7.1 shows a die photo of the wireless charger controller chip. The chip is fabricated in a TSMC 0.18 μm HV process, and the die measures 1.6 mm by 1.7 mm. Much of the chip area is consumed by the nine power switches controlling the tunable power amplifier components. The analog front-end includes the other blocks of the chip, comprising the pulse generator block, analog block, digital logic block, and power amplifier MOSFET driver block.

### 7.2 Test System Implementation

![System test board](image)

Figure 7.2: System test board

Figure 7.2 shows the test wireless charging system incorporating all the components of the circuit design described in Chapter 6. The wireless charging controller chip is in the socket at the left. The primary-side transmitter board is on the left, while the secondary-side receiver board is on the right, positioned above the transmitter board by a fixed distance determined by nylon spacers. The secondary receiver coil is positioned above the primary transmitter coil.

For the system test board, instead of from a cell phone battery the power to the transmitter comes from a power supply hooked to a BNC connector on the board.
7.3 Power Amplifier Control Loop Operation

![Figure 7.3: Dynamic outputs of test system chip](image)

The wireless charging controller chip is tested by observing the operation of the power amplifier control loop as coupling and load conditions change. Figure 7.3 shows the control outputs of the chip as the receiver board is moved relative to the transmitter board, thereby changing coupling conditions. The four traces indicate the control signals for settings A (bottom trace) through D (top trace) chosen by the chip. When a setting is selected by the chip, the corresponding trace goes high and the corresponding set of power amplifier shunt capacitance and resonant tank inductance values are switched into the circuit.

The plot shows the chip outputs as the coupling is decreased and then increased. In the beginning, the coupling is maximum, so none of the traces are high (setting Z is chosen). As the coupling decreases, settings A, then B, then C, then D are chosen. As the coupling increases, setting C, then B, then A, then Z are chosen. The chip adapts to changing conditions within 1 ms.

Figures 7-4 through 7-6 show how the wireless charging controller chip dynamically maintains the zero-voltage switching condition of the power amplifier as coupling changes.
The receiver board distance from the transmitter board is increased in 1 mm increments from 7 mm to 14 mm, and the power amplifier switch voltage waveform is shown for 8 mm, 10 mm, and 13 mm distances. The critical point in each plot is the switch voltage as it is decaying close to zero, specifically at the moment the switch turns on. In all cases from 8 mm to 13 mm distances, the chip adjusts the power amplifier shunt capacitance and resonant tank inductance.
to maintain about zero switch drain voltage at switch turn-on. For shorter distances than 8 mm, the chip is no longer able to compensate and the switch turn-on voltage starts to become more positive. For longer distances than 13 mm, the chip is no longer able to compensate and the switch turn-on voltage starts to become more negative.

7.4 Efficiency and Power Measurement Results and Discussion

![Efficiency and Power Measurement Results and Discussion](image)

Figure 7.7: Measured end-to-end efficiency results vs. other systems
Figure 7.7 and Figure 7.8 show the measurement results for the wireless charging test system. The separation between the transmitter and receiver coils (the charging distance) is varied from 2.4 mm to 14.3 mm. In all tests, the output ultra-capacitor is charged from 2.5 V to 5 V. Efficiency is determined by measuring and deriving the total output energy over the charging cycle and dividing by the total input energy. Note that this efficiency is end-to-end, from input supply to output load. Delivered power is determined by measuring and deriving the average output power over the charging cycle. The charging time reported in Figure 7.8 is the time to deliver 20 J of energy to the output (for example, to charge a 1.5 F capacitor from 1.9 V to 5.5 V).

From Figure 7.7, the maximum end-to-end efficiency of the wireless charging test system in this work is 78% at a charging distance of about 6-7 mm. The efficiency is above 70% over the range of distances between 3 mm and 12 mm. From Figure 7.8, the delivered power varies around the nominal power of 200 mW. Over the charging distance range of 7-12 mm, the range of variation is 80 mW, or $\frac{80}{200} = 40\%$. All of these measurements meet the design objectives in Section 3.1.

The adaptive wireless charging system in this work has significant advantages over other
systems. The adaptive system has higher efficiency and an acceptable delivered power at a wider range of charging distances than a fixed system. Note that the fixed system has the primary transmit inductance $L_1$ nominally fixed to 5 $\mu$H (compensation setting B) and the shunt capacitance $C_P$ fixed to 92 pF, and the efficiency of the fixed system is calculated assuming no control overhead. In particular, the efficiency of the fixed system drops off at short charging distances due to increased switching losses and drops off at longer charging distances due to lower power factor, while the efficiency of the adaptive system stays higher in both cases. The delivered power of the fixed system drops off at longer charging distances due to decreased power factor, while the delivered power of the adaptive system stays at an acceptable level. The adaptive wireless charging system also has higher end-to-end efficiency than three previously published works described in Chapter 3 [3, 2, 4].

The measurement results show that the delivered power varies significantly over charging distance. This is due to the changing Q and power factor of the power amplifier. These variations are partially compensated by the dynamically tunable shunt capacitance adjusted by the chip, but it is not enough. This amount of power variation is tolerable by the application, however, because since the charging times of the receivers in this application are only a couple minutes, such a power variation will not change the charging times by many minutes.

Figure 7.9: Measured efficiency breakdown
Figure 7.9 shows the efficiency breakdown of the test system and gives a clear sense of the reasons behind efficiency variations with charging distance. For very short charging distances, the power amplifier efficiency is relatively low because the high coupling causes the transmitter power amplifier to go out of the zero-voltage switching condition, which causes output capacitor losses each switching cycle. For longer charging distances, the efficiency is limited by the wireless link efficiency, much of which is determined by the physics. For mid-range charging distances where neither of these losses is dominant, the efficiency reaches a peak.

Figure 7.10: Power consumption breakdown at 7 mm charging distance

Figure 7.10 shows the power consumption breakdown of the system at a charging distance of 7 mm. The system is able to achieve high efficiencies in many areas, including the power amplifier and control overhead. The power amplifier efficiency is improved by the dynamic adjustment of the wireless charging control chip to keep the amplifier in the zero-voltage switching condition. The control overhead of the chip is very low. The 1.8 V analog and digital logic and the 3.3 V digital logic consume just 35.28 $\mu$W and 9.9 $\mu$W of power, respectively. The control overhead is dominated by the relay control current and the oscillator supply current.

The two main contributors to the end-to-end efficiency from the breakdown are the wireless link efficiency and the rectifier efficiency. The wireless link efficiency is determined by losses in the parasitic coil resistances and the physical magnetic coupling loss. The wireless charging control chip decreases parasitic resistance losses by keeping the transmitter power amplifier in the zero-voltage switching condition and thus limiting harmonic energy in the coils. For longer charging distances, however, the physical magnetic coupling loss dominates this efficiency. The rectifier efficiency is limited by the voltage drop loss across the discrete diodes. An integrated
CMOS rectifier would increase this efficiency.

### 7.5 Comparison to Previous Work

<table>
<thead>
<tr>
<th>ACADEMIC WORK</th>
<th>THIS WORK</th>
<th>Lee et al.</th>
<th>Lo et al.</th>
<th>Baker, Sarapeshkar</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Authors</strong></td>
<td>Medical / Sensor Devices</td>
<td>ISSCC 2014</td>
<td>IEEE BCS 2013</td>
<td>IEEE BCS 2007</td>
</tr>
<tr>
<td><strong>Application</strong></td>
<td>30 mm / 25 mm</td>
<td>Deep-Brain Stimulation</td>
<td>Retinal Stimulation</td>
<td>Bionic Systems</td>
</tr>
<tr>
<td><strong>Resonant Frequency</strong></td>
<td>6.78MHz</td>
<td>40mm / 10mm</td>
<td>10mm / 10mm</td>
<td>30mm / 30mm</td>
</tr>
<tr>
<td><strong>Input Voltage</strong></td>
<td>3.3V</td>
<td>Fixed ±2V</td>
<td>Fixed 1.8V, ±12V</td>
<td>3.4V</td>
</tr>
<tr>
<td><strong>Output Voltage</strong></td>
<td>Variable (2.5V-5V)</td>
<td>15mm</td>
<td>10mm - 30mm</td>
<td>1mm - 10mm</td>
</tr>
<tr>
<td><strong>Charging Distance</strong></td>
<td>2.4mm - 14.3mm</td>
<td>0.24mW</td>
<td>100mW</td>
<td>10mW</td>
</tr>
<tr>
<td><strong>Efficiency (End-End)</strong></td>
<td>78% @ 7mm, 73% @ 10mm</td>
<td>24% @ 10mm</td>
<td>60% @ 7mm, 54% @ 10mm</td>
<td>96%</td>
</tr>
<tr>
<td><strong>Efficiency (Rectifier)</strong></td>
<td>92%</td>
<td>72%</td>
<td>63% - 82%</td>
<td>72%</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>ACADEMIC WORK</th>
<th>THIS WORK</th>
<th>Kendir et al.</th>
<th>Ghovanloo, Atkouri</th>
<th>Catryse et al.</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Application</strong></td>
<td>Retinal Stimulation</td>
<td>Implants</td>
<td>Implants</td>
<td>Implants</td>
</tr>
<tr>
<td><strong>Coil Diameter</strong></td>
<td>40mm / 22mm</td>
<td>36mm / 20mm</td>
<td>22mm / 13mm</td>
<td>60mm / 20mm</td>
</tr>
<tr>
<td><strong>Resonant Frequency</strong></td>
<td>1MHz</td>
<td>1MHz</td>
<td>500kHz</td>
<td>700kHz</td>
</tr>
<tr>
<td><strong>Input Voltage</strong></td>
<td>5V</td>
<td>Fixed 16V</td>
<td>Fixed 4.83V</td>
<td>2V</td>
</tr>
<tr>
<td><strong>Output Voltage</strong></td>
<td>Variable (3V-6V)</td>
<td>7mm - 15mm</td>
<td>10mm</td>
<td>Fixed 5V</td>
</tr>
<tr>
<td><strong>Charging Distance</strong></td>
<td>7mm - 15mm</td>
<td>250mW</td>
<td>250mW</td>
<td>30mW</td>
</tr>
<tr>
<td><strong>Power</strong></td>
<td>250mW</td>
<td>250mW</td>
<td>250mW</td>
<td>50mW</td>
</tr>
<tr>
<td><strong>Efficiency (End-End)</strong></td>
<td>65.8% @ 7mm, 36.3% @ 15mm</td>
<td>65% @ 2mm, 67% @ 7mm, 51% @ 15mm</td>
<td>29.9% (with LD0 at output)</td>
<td>36% (with zener regulator at output)</td>
</tr>
<tr>
<td><strong>Efficiency (Rectifier)</strong></td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>INDUSTRY WORK</th>
<th>THIS WORK</th>
<th>WiTricity</th>
<th>Qualcomm</th>
<th>Energiel</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Group</strong></td>
<td>Medical / Sensor Devices</td>
<td>Power 2.0</td>
<td>4V</td>
<td>Qi</td>
</tr>
<tr>
<td><strong>Application</strong></td>
<td>30 mm / 25 mm</td>
<td>Mobile Devices on Charging Pad</td>
<td>Mobile Devices on Charging Pad</td>
<td>Mobile Devices on Charging Pad</td>
</tr>
<tr>
<td><strong>Coil Diameter</strong></td>
<td>30mm / 75mm</td>
<td>Mobile Devices on Charging Pad</td>
<td>Mobile Devices on Charging Pad</td>
<td>Mobile Devices on Charging Pad</td>
</tr>
<tr>
<td><strong>Resonant Frequency</strong></td>
<td>6.78MHz</td>
<td>6.78MHz</td>
<td>6.78MHz</td>
<td>4V (lithium-ion battery)</td>
</tr>
<tr>
<td><strong>Output Voltage</strong></td>
<td>Variable (2.5V-5V)</td>
<td>~4V (lithium-ion battery)</td>
<td>~4V (lithium-ion battery)</td>
<td>~4V (lithium-ion battery)</td>
</tr>
<tr>
<td><strong>Charging Distance</strong></td>
<td>4mm - 13mm</td>
<td>~10nm</td>
<td>~10nm</td>
<td>~10nm</td>
</tr>
<tr>
<td><strong>Power</strong></td>
<td>200mW</td>
<td>~1-10W</td>
<td>~1-10W</td>
<td>~1-10W</td>
</tr>
</tbody>
</table>

Figure 7.11: Specification comparison to previous published work and commercial products

Figure 7.11 compares critical specifications of this work to previously published work and commercial products currently available [34, 1, 3, 4, 2, 35, 36, 12, 37, 13]. In the academic arena, many other publications have demonstrated wireless power systems for medical applications, such as deep-brain or retinal stimulation. These systems are all based on resonant inductive power transfer like in this work, and all use similarly sized coils to transfer similar power levels over similar distances.

There are two main differences between this work and previously published academic work. The first is that previous systems are designed for a constant output voltage. This implies an application in which the transmitter is continuously supplying a receiver device with power to maintain a supply voltage in the receiver. In contrast, this work is designed for a variable output voltage. This imposes more variability on the transmitter-side loading. This is also
more suitable for the application of portable-to-portable wireless charging, in which it is more convenient for the user to quickly charge a portable device for a long battery life as opposed to constantly wirelessly powering this portable device.

The second difference is in the measured end-to-end efficiency of this system versus previous systems. The peak efficiency is higher than previous work, and the efficiency is higher over a range of charging distances. This is due to the demonstrated power amplifier control loop as well as the system design, and it has advantages for the portable-to-portable wireless charging application in which the transmitter is energy-constrained and efficiency is key.

In the industrial arena, many products and standards exist with similar specifications compared with this work. These standards all support resonant inductive transfer of wireless power over similar distances. The major difference is the application. Consumer products on the market today focus on pad-based charging of higher-power consumer electronics such as cell phones. This means these products have larger coils and transfer higher power. This work focuses on the application of charging lower-power consumer electronics devices from higher-power ones. As a result, the system designed in this work has smaller coils and lower power. This new application space is largely unexplored by industry and has potential because of the vast number of low-power portable devices people own and because of the ubiquity of cell phones to charge these portable devices.

In summary, the unique design of the system in this work allows the system to achieve high efficiency in the dynamically changing conditions of portable-to-portable wireless charging, which is a key enabler. This promising application is currently underdeveloped in industry.

### 7.6 Prototype System Implementation

A prototype system is designed and implemented that draws energy from a cell phone battery to wirelessly charge low-power portable devices. The design of this prototype system focuses on a small form factor and high charging efficiency.
Figure 7.12 shows the implemented prototype system transmitter. The connector on the left in Figure 7.12 is a 30-pin connector used in many iPhone products. Using this connector, the prototype system draws power for the 3.3 V supply from the cell phone battery. The transmitter board measures 2.1 in by 1.35 in, about the size of a matchbox. This prototype system transmitter behaves in the same way as the test system described in the previous sections.

Using this prototype system transmitter, a number of receivers for different applications are designed. These applications are described in Chapter 8.

7.7 Chip and System Limitations

There are a three limitations of the chip and system that may impact operation. First, as discussed in Section 7.3, the wireless charging system is designed to maintain zero-voltage switching of the power amplifier between the charging distances of 8 mm and 13 mm at 4 V output voltage. This is limited by the number and granularity of the shunt capacitor and resonant inductor values. A different design can be derived according to the equations in Chapter 4 that uses different tunable values to support different charging distance ranges. The tradeoff is between the size and cost of the tunable components versus the range of charging distances supported. If a greater number of tunable values is desired, the chip silicon must be
redesigned to incorporate more compensation settings (more than the current number of five).

Also, ideally, care should be taken to not vary the tunable components too much between consecutive settings. If consecutive settings are very different values of the tunable components, then for certain charging conditions the chip may not be able to determine a satisfactory setting with respect to the switch turn-on voltage being within the specified range (set by 0 and $V_{REF}$ at the analog block of the chip). In this case, the chip will oscillate between two consecutive settings. This is not ideal but entirely acceptable, since it has little effect on the operation or efficiency of the wireless charging system (beyond the small efficiency loss due to the outputs that switch at the $\sim 1$ kHz control logic clock frequency). $V_{REF}$ may be increased to increase this range and prevent such oscillations, at the cost of decreased control on the switch turn-on voltage values that are considered satisfactory by the chip. $V_{REF}$ must not be increased beyond 3.3 V, which is the voltage rating of the corresponding input transistor in the analog block.

A second limitation relates to chip operation as the load and coupling conditions are varied very slowly. When the switch turn-on voltage is very close to the endpoints of the specified range (0 or $V_{REF}$), the comparators in the analog block of the chip have significantly longer decision times. In some cases, this delay may be longer than the delay before the comparator outputs are clocked into the logic in the digital block of the chip. The result of this is that two comparator outputs may be high, and two consecutive settings may be chosen by the chip at the same time. After this, as the chip performs further compensation, these two settings are adjusted independently. The presence of multiple settings is eliminated when an extreme setting is chosen by the chip (Z or D) or on reset. The presence of multiple settings has only a small effect on the wireless charging system operation. The setting with a larger shunt capacitance and smaller resonant inductance dominates the system operation, and any other selected settings have little effect. The efficiency is degraded slightly due mostly to the power dissipation of additional relay control currents. For every additional relay that turns on, there is an additional 1.7 mA of control current from a 1.8 V supply. This corresponds to $1.7 \times 1.8 \approx 3$ mW of additional power dissipation, or 1.5% of the nominal 200 mW delivered power.

A third limitation relates to system operation at power-on, especially when there is no receiver present. In all cases, the chip starts at the lowest compensation setting Z at power-on,
and in the case with no receiver, the chip quickly increases the setting to A, then B, then C, then D. As the settings are switched, one relay for the transmitter multi-tapped coil must turn off and another must turn on. The relays are designed such that the turn-on time is much shorter than the turn-off time. If the 1.8 V supply for the relay control currents is unstable during this time, however, such as at power-on, the relays may take an unusually long time to turn-on. If the turn-on time exceeds the turn-off time, the current flowing through the resonant tank will have no path to flow, and this will cause the current to go completely out of sync with the power amplifier switch driving this current. The result is very limited power transfer at poor efficiency. This problem is especially prevalent when hot-plugging a prototype system into a cell phone. Although this is a serious problem, it is an easy one to solve as long as the 1.8 V supply rail is allowed to stabilize for a sufficient amount of time (a couple milliseconds) before the power amplifier switch starts switching at power-on, and as long as the relay control circuits are carefully designed such that the turn-on time is much shorter than the turn-off time.

7.8 Summary of Wireless Charging Integrated Circuit and System Design and Implementation

The last three chapters present a complete wireless charging system design and implementation consisting of a resonant inductive wireless power circuit with dynamic adjustment of the primary side controlled by an ultra low-power microchip. All discrete components in the system are chosen or designed carefully for maximum efficiency. The system achieves higher end-to-end efficiency than previously published results over varying charging conditions. The next chapter highlights some of the applications of this system in practice.
Chapter 8

Portable-to-Portable Wireless Charging Applications

The implemented portable-to-portable wireless charging system with an integrated feedback and compensation chip has diverse applications. A number of commercial low-power portable devices and research devices benefit from portable wireless charging using a cell phone. These applications range from medical devices to wireless accessories to niche devices. Seven demonstration applications are designed and built to showcase this vast diversity, and this chapter describes each application.

![General receiver circuit for applications](image)

**Figure 8.1: General receiver circuit for applications**

For each application, a receiver circuit is designed according to Figure 8.1. This is a basic resonant inductive receiver as described in Chapter 2 and designed in Chapter 6. The energy stored on the output ultra-capacitor $C_{OUT}$ is passed to a DC-DC converter to generate the necessary voltage rails for the particular application device.

One important aspect of the wireless charging system in this work is that the secondary-side
receiver circuit is extremely simple. The entire system is designed so that virtually all of the complexity is on the transmitter side because in practice, there are many available receivers (a given user has many portable devices) that can be charged from a single transmitter (a given user has one cell phone). Thus, moving complexity to the transmitter side saves overall cost and complexity. Simple receiver circuitry also makes adoption in the marketplace easy. Since there are countless low-power portable devices manufactured by many companies today, it is imperative that the changes required by wireless charging receivers be as simple and noninvasive as possible. This way, individual manufacturers can implement wireless charging receivers easily using reference designs and off-the-shelf devices. The wireless charging transmitter can be implemented by a single manufacturer as a mobile device accessory or adopted by the handful of mobile device manufacturers.

8.1 Toys - Toy Light

Figure 8.2: Wireless charging system for a toy light display

Figure 8.2 shows an application using the prototype wireless power transmitter to charge and power a toy light display using energy from a cell phone. The light display includes 31 LEDs arranged in a floral pattern. The center white LED is always on, while red, green, and blue
LEDs turn on in sequence around the floral pattern. The LEDs together draw around 180 mW of power and turn on whenever the cell phone transmitter is in range of the receiver.

This toy light demonstration represents a number of flashy low-power toys or displays that may be powered through wireless power transfer. Other potential examples include miniature train sets with trains that run off wireless power, interactive dolls that are wirelessly recharged using a cell phone, and electronic Lego blocks that receive wireless power.

8.2 Wearable Devices - Fitbit Fitness Tracker

Figure 8.3: Wireless charging system for a Fitbit fitness tracker

Figure 8.3 shows an application using the prototype wireless transmitter to charge a Fitbit fitness tracker. The Fitbit is a smart wristband that measures and logs a person’s movements throughout the day. The Fitbit, along with many other similar fitness trackers, is charged through a USB cable plugged into a computer port (no USB charger is provided for charging
from the wall outlet). The internal battery lasts about five days.

The Fitbit is modified to include a receiver coil around the band of the device. The coil is made from eight turns of Litz wire. At one section in each turn of the coil, the wire is cut and then rejoined through a flat-flexible connector. This allows the coil to be disconnected and the band to come apart when putting on and taking off the wristband. The receiver circuitry is unlike the circuitry in Figure 8.1 in that an ultra-capacitor and DC-DC converter are not used at the output. Instead, the rectifier feeds directly into the battery-charging circuitry of the Fitbit, and the internal battery is recharged using wirelessly received power. The charging voltage at the output of the rectifier is limited by a shunt zener diode. The Fitbit designs and specifications used for modifying the device are given in [38, 39].

The Fitbit can be recharged by placing the device on top of the cell phone transmitter, as shown in Figure 8.3. Charging takes about 15 minutes, after which the Fitbit lasts an entire day. A typical use case for this application is for a user to take off the device every day before taking a shower, place the device on a cell phone, and charge the device in the amount of time it takes for a shower, for a full day’s use. This is more convenient than turning on a computer to charge the device through a USB cable.

The Fitbit represents a number of wearable devices that may be quickly recharged for a full day’s use. Other examples are the vast number of other fitness trackers on the market today, including the Nike Fuelband and the Jawbone Up, as well as other wearables in research, such as smart watches.
8.3 Medical Devices - Cochlear Implant

Figure 8.4: Wireless charging receiver for a cochlear implant

Figure 8.4 shows the wireless charging receiver for a research cochlear implant. A cochlear implant is a medical device that is surgically implanted inside the ear with electrodes to stimulate neurons in the cochlea to give the sensation of hearing. The chip on the top of the board in the figure is a sound processor that converts input sound to the stimulation waveforms for the electrodes. The chip is designed for a fully-implantable cochlear implant prototype in which there are no devices external to the ear and the entire implant is invisible from the perspective of another person. The detailed design for the chip is reported in [40]. All other components required for operation of the cochlear implant are on the board, including the wireless power receiver, power converters, and clock generators. The entire system is the same size as a conventional cochlear implant.

Wireless power is crucial for implanted devices such as the cochlear implant. The wireless power receiver is located on the bottom of the board in the figure. The basic receiver circuit in Figure 8.1 is used with DC-DC converters to generate 0.6 V, 1.5 V, 3.3 V, and 8.2 V supplies for the chip. The entire system can be charged in about 2 minutes (to charge a 1.5 F ultra-capacitor) for a 7.5-hour battery life. This is convenient for users, since only a couple quick charge sessions using a cell phone are needed per day.

The cochlear implant is a representative of a number of medical devices, implanted and
non-implanted, that would benefit from portable-to-portable wireless charging. Other examples include pacemakers [41], retinal implants [4], deep-brain stimulators [34], EEG and ECG sensors [42], and gastrointestinal sensors [43]. Most of these devices are currently in research and development stages, and having fast and efficient portable wireless charging technology would facilitate this design process.

8.4 Media Devices - iPod Shuffle

Figure 8.5 shows the wireless charging receiver for an iPod Shuffle in an armband. Many people use the portable MP3 player to listen to music while exercising, in which case the device is oftentimes placed in an armband worn around the upper arm.

The design and specifications of the MP3 player used in modifying it are reported in [44]. The wireless power receiver is based on the circuit in Figure 8.1 and generates a 3.5 V supply for the portable device. The circuitry is split up into two parts, the first attached to the armband as seen in the right of the figure and the second embedded into the device itself or located behind the device. The first part includes the wireless power receiver coil and rectifier while the second part includes the output ultra-capacitor and DC-DC converter. In a typical use case, the user quickly charges the device just before an exercise routine by placing the transmitter cell phone close to the receiver coil in the armband. Charging takes about 2 minutes (to charge a 3 F ultra-capacitor). Afterwards, the user can operate the device for 30 minutes, or about the
length of a typical exercise routine. This is more convenient for users than the usual charging process, which involves turning on a computer and charging the device through a special dock.

The iPod Shuffle is a representative of a range of portable media-related devices that may be wirelessly recharged using a cell phone. Other devices include a variety of other MP3 music players, as well as low-power cameras.

8.5 Wireless Accessories - Wireless Keyboard

Figure 8.6 shows the wireless charging receiver for a wireless keyboard. People often use these Bluetooth-connected wireless keyboards to reduce clutter or increase positional freedom when using a computer, laptop, or tablet.

The wireless charging receiver is based on Figure 8.1 and generates a 2.8 V supply for the portable device. As shown in the figure, the entire receiver circuitry including the receiver coil, receiver electronics, and storage ultra-capacitor is designed to fit within the battery compartment of the portable device. In typical use cases, the user may quickly charge the wireless keyboard before using it or the higher-power device that the keyboard is being used with may charge the keyboard during operation. Charging takes about 2 minutes (to charge a 3 F ultra-capacitor) for a 3-hour battery life. Being able to wirelessly charge the keyboard when needed is more convenient for users than having to expend a large effort to buy and replace batteries in the device when they become depleted.
The wireless keyboard is only one example of the many wireless accessories that may be outfitted with wireless receiver circuitry. Other devices include wireless mice and Bluetooth headsets.

8.6 AA / AAA-Battery Operated Devices - Graphing Calculator

Figure 8.7 shows the wireless charging receiver for a TI-84 graphing calculator. The disassembly guidelines of the calculator used in modifying it are reported in [45]. The wireless charging receiver is based on Figure 8.1 and generates a 5.5 V supply for the portable device. As shown in the figure, the entire receiver circuitry including the receiver coil, receiver electronics, and storage ultra-capacitor is designed to fit within the battery compartment of the portable device. In a typical use case, the user may quickly charge the calculator before using it. Charging takes about 5 minutes (to charge a 5 F ultra-capacitor) for a 1-hour battery life. This is convenient
for users because batteries never have to be replaced or taken out and charged in a battery charger.

The calculator is only one example of the many AA or AAA battery-operated devices that may be wirelessly recharged instead of powered by conventional batteries. Other devices include low-power stereos, flashlights, and toys.

## 8.7 Coin Cell Battery-Operated Devices - Bicycle Light

Figure 8.8 shows the wireless charging receiver for a bicycle light. The wireless charging receiver is based on Figure 8.1 and generates a 2.4 V supply for the portable device. As shown in the figure, the entire receiver circuitry including the receiver coil, receiver electronics, and storage ultra-capacitor is designed to fit inside the normal enclosure of the bicycle light. In particular, the receiver coil is built into the back casing of the device. In a typical use case, the user may quickly charge the light before using it. Charging takes about 2 minutes (to charge a 3 F ultra-capacitor) for a 25-minute battery life. This is convenient for users because batteries never have to be replaced, and users never have to worry about having a discharged bicycle light battery because they can charge the device whenever they need it.

The bicycle light is only one example of the many coin cell battery-operated devices that
may be wirelessly recharged instead of powered by conventional batteries. Other devices include watches, hearing aids, and toys.

8.8 Chapter Summary

<table>
<thead>
<tr>
<th>Category</th>
<th>Device</th>
<th>Charging Time</th>
<th>Usage Time</th>
</tr>
</thead>
<tbody>
<tr>
<td>Toys</td>
<td>Toy Light</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>Wearable Devices</td>
<td>Fitbit Fitness Tracker</td>
<td>15 Minutes</td>
<td>1 Day</td>
</tr>
<tr>
<td>Medical Devices</td>
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<td>2 Minutes</td>
<td>7.5 Hours</td>
</tr>
<tr>
<td>Media Devices</td>
<td>iPod Shuffle</td>
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<td>Wireless Accessories</td>
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<tr>
<td>AA / AAA Battery Devices</td>
<td>Graphing Calculator</td>
<td>5 Minutes</td>
<td>1 Hour</td>
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<tr>
<td>Coin Cell Battery Devices</td>
<td>Bicycle Light</td>
<td>2 Minutes</td>
<td>25 Minutes</td>
</tr>
</tbody>
</table>

Table 8.1: Summary of wireless charging receiver applications

Table 8.1 shows a summary of the diverse applications for the portable-to-portable wireless charging system. This shows the vast number of portable devices that would benefit from this system, in many different areas. In all cases, a person can charge a portable device using a cell phone in a few minutes and subsequently use the device for a typical day. Note that as portable devices are developed to use less power, these devices will be able to be charged in a shorter amount of time and last on that charge for a longer amount of time.
Chapter 9

Conclusions and Future Directions

9.1 Summary of Contributions

Today’s world is dominated by portable electronics. These devices facilitate people’s activities in a number of areas, from entertainment to health to work. Many of the ways low-power portable devices influence people’s lives are highlighted in Chapter 8. In addition, researchers and companies are actively exploring new portable devices, in particular devices that integrate more functions and devices that sense more of the world around us.

The fundamental requirement of all electronics is the supply of power. With the pervasiveness of low-power portable devices, the question of how to supply the unique power demands of each device is becoming an escalating issue. Most portables on the market today answer this question through wired charging cables. But as the number of portable devices increases further, increasing the number of charging cables in proportion will lead to greater user inconvenience, increased costs, and decreased environmental sustainability.

This work attempts to solve this issue by proposing a common wireless power interface to recharge all these low-power portable devices using higher-power portable devices. Higher-power portable devices such as cell phones are ubiquitous today and are excellent sources of portable energy. Since many portable devices people use are generally closer to people’s cell phones than to electrical outlets, in many cases it is more convenient to recharge these portable devices using cell phones than using wired chargers. The use case model of quick wireless charging of portable devices on demand is an attractive one in many applications, as outlined
in Chapter 8. A common wireless charging interface also avoids the issue of the number of charging cables scaling in proportion with the number of portable devices, and this facilitates increased environmental sustainability and lower cost.

The portable-to-portable wireless charging system designed in this work attempts to solve many of the challenges associated with portable wireless power transfer. These challenges stem from the use case model, in which the transmitter portable device is by definition energy constrained and the charging process naturally entails dynamically changing coupling and load conditions. This means that maintaining high efficiency while delivering consistent power levels over changing conditions is key. This work addresses these challenges through derivation of a new power amplifier control technique, a complete wireless charging integrated circuit and system design, and a series of demonstration applications.

Mathematical analysis presented in Chapters 2-4 develops the theory behind the standard wireless power topology and shows how this design reacts to changing conditions with respect to key metrics. A power amplifier control method is derived that improves upon the standard wireless power topology to improve efficiency and balance power as coupling and load conditions change. Specifically, the control loop includes feedback from the transmitter-side power amplifier to inform compensation of the power amplifier shunt capacitance and primary resonant tank inductance to maintain zero-voltage switching while regulating power.

Integrated circuit and system design presented in Chapters 5-7 implement the power amplifier control loop in an adaptive resonant inductive wireless charging system. A microchip is fabricated in a 0.18 μm HVCMOS process that dynamically compensates the power amplifier based on changing conditions. A complete wireless charging system is built around this microchip to wirelessly transfer energy from a cell phone battery to a low-power portable device. A maximum end-to-end efficiency of 78% is achieved transferring about 200 mW over a 7 mm distance between 25-30 mm diameter coils. Efficiencies over 70% are maintained for a range of distances from 4-12 mm and delivered power levels vary less than 40% for a range of distances from 7-12 mm.

As detailed in Chapter 8, seven demonstration applications are designed and implemented that showcase the applicability of the portable-to-portable wireless charging system to recharge a diverse set of portable devices, from wearable devices to medical implants to wireless accessories.
Each portable can be charged in 2-15 minutes for a typical day’s use.

These applications represent only a small set of use cases for portable-to-portable wireless charging. The following section describes how further work on the technology may be done to expand its applicability even more.

9.2 Towards a Power Sharing Vision

Chapter 1 presents a vision for portable devices, termed power sharing, in which power is shared among devices similar to the way data is currently shared. In this vision, higher-power devices robustly replenish energy stores of lower-power devices to optimize some overall metric conveniently and efficiently.

Figure 9.1: Body area network for data and power transfer

One embodiment of this vision is presented in Figure 9.1, which shows a body area network of many sensor and other wearable portable devices that both transfer data with and receive power from a cell phone. Body area networks for data communication are extensively studied [46, 47, 48, 49, 50, 51], and a wireless power framework on top of this network would facilitate further development. This is an area of active research [52, 53, 54, 55].

To realize this power sharing vision, there are a number of key directions for continued research into wireless power transfer. These include:
• **Power transfer over longer distances**: A major technology required by the power sharing vision is the ability to wirelessly transmit energy over longer distances, on the order of a meter. The main impediment to achieving this is the extremely low coupling between a transmitter and receiver separated by a long distance. This leads to system issues such as low end-to-end efficiency, as explored in [56, 57]. A number of potential research avenues exist to overcome this obstacle, however, and systems exist that transfer power over longer distances [58]. First, low efficiency may be tolerable in some applications with low or non-critical power requirements. Second, different transmitter/receiver designs or frequencies can be investigated to decrease the effect of low coupling [59, 60, 24, 61]. Third, intermediate or relay nodes may be added to a wireless power system to decrease the effective coupling distance [11].

• **Power transfer among multiple devices**: A number of considerations arise with power transfer among multiple devices as compared to power transfer between a single transmitter and a single receiver. These include the differing power sourcing and sinking specifications of each device, the differing amounts of energy stored in each device, the differing coupling and load conditions present between each pair of devices, and the methods and fairness of these methods of transferring power from a single transmitter to multiple receivers or from multiple transmitters to a single receiver. Some of these considerations are included in [62, 63]. Metrics can be defined that consider the overall performance of the power network with respect to efficiency or charging time, and systems can be designed that achieve an overall optimum.

• **User interaction with wireless power transfer**: As the number of wireless power-enabled devices increases, an interesting question arises regarding exactly what involvement users should have with the sharing of energy among devices. In current wired charging use case models and in the wireless charging use case model proposed in this work, the user consciously enables the transfer of energy from a transmitter to receiver. In the future, a better use case model might be automatic transfer of energy, in which many devices stay charged without any user effort. Other issues worth considering are which devices are allowed to share power, and how this is determined on the device level.
and on the user level.

- **Security with wireless power transfer**: As with any technology in which operation uses a public medium, power transfer over the air is by default vulnerable to security breaches. Possible problems include undesired transfer of energy from a user device to an unapproved device, which raises issues with user satisfaction, and transfer of an undesirably high amount of power from a user device, which raises issues with user safety. Security layers need to be designed over the power electronics to authenticate users and protect against adversaries. Security is well explored in the field of wireless data communication [64, 65] and these principles can be applied to wireless power transfer.

- **Multi-modal operation with energy harvesting**: As portable devices are developed with lower and lower power requirements, energy harvesting becomes viable for the energy requirements of more devices, at least in certain modes of operation [66]. Certain applications can benefit from energy harvesting when power demands are low and wireless power transfer when power demands are high. Devices with energy harvesting mechanisms can also self-replenish their stores of energy to transfer to other devices [67].
Appendix A

Details of Class-E Power Amplifier Operation

Appendix A presents details of the Class-E power amplifier operation across changing coupling and load conditions as described in Chapter 4. For reference, the Class-E power amplifier is shown in Figure A.1. This appendix first presents the effect of changing conditions on the primary-side equivalent resistance $R_{EQ}$. This is to show the effect of changing $R_{EQ}$ on power amplifier zero-voltage switching (ZVS). ZVS is considered in two cases: full ZVS (both switch drain voltage and slope are zero at switch turn-on) and general ZVS (only voltage is zero). In each case, tuning the power amplifier shunt capacitance $C_P$ and resonant tank excess inductance $L_{EX}$ maintains ZVS across changing conditions. Simulation results at the end show that the power amplifier control method of tuning $C_P$ and $L_{EX}$ maintains general ZVS. As shown in
Chapter 4, this improves efficiency and balances delivered power across changing conditions.

A.1 Effect of Changing Coupling and Output Voltage on Equivalent Resistance

The effect of changing coupling and output voltage on primary-side equivalent resistance is found by considering the output power.

From Subsection 2.3.1, the output power is delivered only by the component of the rectifier voltage that is at the operating frequency. This component has amplitude $\frac{4}{\pi}V_{OUT}$. Thus, the average output power is

$$P_{OUT} = \frac{1}{2} \left( \frac{4}{\pi} \right)^2 \frac{V_{OUT}^2}{R_L} \quad (A.1.1)$$

From Equation 2.3.3, the output power may also be written as

$$P_{OUT} = V_{S,rms}^2 \left( \frac{k^2 L_1 C_2 R_L}{(k^2 L_1 + C_2 R_{PRI}(R_L + R_2))^2} \right) \quad (A.1.2)$$

Equating the two expressions and solving for $R_L$ gives

$$V_{S,rms}^2 \left( \frac{k^2 L_1 C_2 R_L}{(k^2 L_1 + C_2 R_{PRI}(R_L + R_2))^2} \right) = \frac{1}{2} \left( \frac{4}{\pi} \right)^2 \frac{V_{OUT}^2}{R_L} \quad (A.1.3)$$

$$\Rightarrow \sqrt{V_{S}^2} \left( \frac{k^2 L_1 C_2 R_L}{(k^2 L_1 + C_2 R_{PRI}(R_L + R_2))^2} \right) = \sqrt{\left( \frac{4}{\pi} \right)^2 \frac{V_{OUT}^2}{R_L}} \quad (A.1.4)$$

$$\Rightarrow R_L = \frac{4}{\pi} \frac{V_{OUT}}{|V_{S}|} \left( \frac{k^2 L_1 + C_2 R_{PRI}R_2}{k \sqrt{L_1 C_2} - C_2 R_{PRI} \frac{4 V_{OUT}}{|V_{S}|}} \right) \quad (A.1.5)$$

$$\Rightarrow R_L = \frac{k^2 L_1 L_2 w^2 + R_2 R_{PRI}}{k \sqrt{L_1 L_2 w^2} \frac{|V_{S}|}{V_{OUT} - R_{PRI}}} \quad (A.1.6)$$

where the last equation comes from the fact that $w = \frac{1}{\sqrt{L_2 C_2}}$.

Then from Equation 2.3.2, the reflected primary-side impedance at resonance is

$$R_{EQ} = \frac{k^2 L_1}{C_2(R_2 + R_L)} \quad (A.1.7)$$
\[
\Rightarrow R_{EQ} = \frac{k\sqrt{L_1L_2w} \left( \frac{\pi |V_S|}{4 V_{OUT}} k\sqrt{L_1L_2w} - R_{PRI} \right)}{\frac{\pi |V_S|}{4 V_{OUT}} R_2 + k\sqrt{L_1L_2w}} \quad (A.1.8)
\]

A.2 Effect of Changing Equivalent Resistance on Power Amplifier Operation in Full ZVS

In angular terms, given that the switch turns on at \( wt = 2\pi N \) and turns off at \( wt = 2\pi N + \pi \), for \( NeZ \), full ZVS means that two statements are assumed for \( wt = 2\pi N \):

\[
v(wt) = 0 \quad (A.2.1)
\]

\[
\frac{d}{d(wt)} v(wt) = 0 \quad (A.2.2)
\]

where \( v(wt) \) is the switch drain voltage.

![Switch voltage waveform with full ZVS condition](image)

Figure A.2: Switch voltage waveform with full ZVS condition

Figure A.2 shows a power amplifier switch voltage waveform that satisfies the two ZVS assumptions. The following analysis determines the required values of the power amplifier parameters \( C_P \) and \( L_{EX} \) and shows how changing \( R_{EQ} \) might affect these requirements. The analysis consists of three parts. In the first part, an expression for \( v(wt) \) is derived given full ZVS. In the second and third parts, this expression is used to derive the required values of \( L_{EX} \) and \( C_P \), respectively.
A.2.1 Derivation of $v(\omega t)$ in Full ZVS Condition

Due to the ZVS assumptions, at $\omega t = 2\pi$,

$$i_C(2\pi) = wC_P \frac{d}{d(\omega t)} v(\omega t)|_{\omega t=2\pi} = 0$$  \hspace{1cm} (A.2.3)

Since the switch has just begun to conduct current at this time,

$$i_S(2\pi) = I_O + i_R(2\pi) = 0$$  \hspace{1cm} (A.2.4)

Let $i_R(\omega t) = I_R \sin(\omega t + \phi)$. Then,

$$I_O = -i_R(2\pi) = -I_R \sin(2\pi + \phi) = -I_R \sin(\phi)$$  \hspace{1cm} (A.2.5)

For $\pi < \omega t < 2\pi$, in other words when the switch is off,

$$i_C(\omega t) = I_O + I_R \sin(\omega t + \phi) = -I_R (\sin(\phi) - \sin(\omega t + \phi))$$  \hspace{1cm} (A.2.6)

$$\Rightarrow v(\omega t) = \frac{1}{wC_P} \int_{\pi}^{\omega t} i_C(\omega t) d(\omega t) = -I_R X_P [\cos(\omega t + \phi) + \cos(\phi) + \sin(\phi)(\omega t - \pi)]$$  \hspace{1cm} (A.2.7)

where $X_P = \frac{1}{wC_P}$. At $\omega t = 2\pi$, $v(\omega t) = 0$, so

$$v(2\pi) = 2 \cos(\phi) + \pi \sin(\phi) = 0$$  \hspace{1cm} (A.2.8)

$$\Rightarrow \phi = \tan^{-1}\left(-\frac{2}{\pi}\right) \approx -32.48^\circ$$  \hspace{1cm} (A.2.9)

$$\Rightarrow i_R(\omega t) = I_R \sin(\omega t - 32.48^\circ)$$  \hspace{1cm} (A.2.10)

$$\Rightarrow v(\omega t) = I_O X_P \left[ \omega t - \frac{3\pi}{2} - \frac{\pi}{2} \cos(\omega t) - \sin(\omega t) \right]$$  \hspace{1cm} (A.2.11)

This shows that in ZVS, the resonant current is phase shifted from the switch drive voltage by $\phi = -32.48^\circ$. This is observed in the Class-E power amplifier waveforms in Chapter 2, in Figure 2.7.
A.2.2 Derivation of $L_{EX}$ in Full ZVS Condition

The power delivered to $R_{PRI} + R_{EQ}$ comes only from voltage that is in phase with the resonant current. In particular, considering the frequency components of $v(wt)$ at the fundamental operating frequency, only the component that is in phase with $i_R(wt)$ delivers power to the load. The out-of-phase component is dropped across the excess inductance $L_{EX}$. Using a Fourier series expansion, the in-phase component magnitude is,

$$V_S = \frac{1}{\pi} \int_0^{2\pi} v(wt) \sin(wt + \phi) d(wt) = \frac{1}{\pi} I_O X_P \left[ 2\sin(\phi) - \frac{\pi^2}{4}\sin(\phi) - \frac{\pi}{2}\cos(\phi) \right]$$  \hspace{1cm} (A.2.12)

The out-of-phase component magnitude is,

$$V_{EX} = \frac{1}{\pi} \int_0^{2\pi} v(wt) \cos(wt + \phi) d(wt) = \frac{1}{\pi} I_O X_P \left[ 2\cos(\phi) - \frac{\pi^2}{4}\cos(\phi) + \frac{\pi}{2}\sin(\phi) \right]$$  \hspace{1cm} (A.2.13)

Since voltage drops are proportional to impedances,

$$\frac{V_{EX}}{V_S} = \frac{wL_{EX}}{R_{PRI} + R_{EQ}} = \frac{\left( 2 - \frac{\pi^2}{4} \right) \cos(\phi) + \frac{\pi}{2} \sin(\phi)}{\left( 2 - \frac{\pi^2}{4} \right) \sin(\phi) - \frac{\pi}{2} \cos(\phi)} = \frac{\pi^3}{16} - \frac{\pi}{4} \approx 1.152$$  \hspace{1cm} (A.2.14)

This shows that in ZVS, the excess inductance $L_{EX}$ must be related to the reflected resistance and primary-side resistance $R_{PRI} + R_{EQ}$ by a constant factor. In particular, when $R_{EQ}$ changes, ZVS is not maintained unless $L_{EX}$ is adjusted accordingly. Note that given a value for $R_{EQ}$, there is a unique value of $L_{EX}$.

A.2.3 Derivation of $C_P$ in Full ZVS Condition

Power enters the system through the choke inductor $L_{CHOKE}$. Ideally, it should be sized as large as possible to keep the input current $I_O$ as constant as possible. Given this, the choke
inductor cannot have any average voltage across it, so \( v(wt) = V_{CC} \), or

\[
V_{CC} = \frac{1}{2\pi} \int_{\pi}^{2\pi} v(wt)d(wt) = \frac{1}{\pi} I_O X_P
\]  
(A.2.15)

Plugging this into Equation A.2.12 gives,

\[
V_S = V_{CC} \left[ 2\sin(\phi) - \frac{\pi^2}{4} \sin(\phi) - \frac{\pi}{2} \cos(\phi) \right] = -\frac{4}{\sqrt{\pi^2 + 4}} V_{CC} \approx -1.07 V_{CC}
\]  
(A.2.16)

Since \( I_O = -I_R \sin(\phi) = I_R \frac{2}{\sqrt{\pi^2 + 4}} \), plugging this into Equation A.2.15 gives,

\[
V_{CC} = \frac{2}{\pi \sqrt{\pi^2 + 4}} I_R X_P
\]  
(A.2.17)

Since \( I_R = -\frac{V_S}{R_{PRI} + R_{EQ}} \) (the negative sign comes from the direction of the currents in Figure 4.6), plugging this and Equation A.2.16 into the previous equation gives,

\[
V_{CC} = \frac{8}{\pi(\pi^2 + 4)} \frac{V_{CC}}{wC_P} \frac{1}{R_{PRI} + R_{EQ}}
\]  
(A.2.18)

\[
\Rightarrow wC_P = \frac{8}{\pi(\pi^2 + 4)} \frac{1}{R_{PRI} + R_{EQ}}
\]  
(A.2.19)

This shows that in ZVS, the capacitance \( C_P \) must be related to the reflected resistance and primary-side resistance \( R_{PRI} + R_{EQ} \) by a constant factor. In particular, when \( R_{EQ} \) changes, ZVS is not maintained unless \( C_P \) is adjusted accordingly. Note that given a value for \( R_{EQ} \), there is a unique value of \( C_P \).

### A.3 Effect of Changing Equivalent Resistance on Power Amplifier Operation in General ZVS

In a more general case, a partial zero-voltage switching condition can be met in which only one of the assumptions is true, namely that at switch turn-on, only the switch drain voltage is zero.
That is, for $wt = 2\pi N$:

$$v(wt) = 0$$  \hspace{1cm} (A.3.1)

In other words, the slope of the switch voltage at turn-on is not necessarily zero. Define this non-zero slope to be $S$.

\[ \text{Figure A.3: Switch voltage waveform with general ZVS condition} \]

Figure A.3 shows a power amplifier switch voltage waveform that satisfies the general ZVS assumption. In this condition, the effect of changing equivalent resistance on the requirements of the power amplifier parameters is analyzed. In this condition, there are seven unknowns: $\phi$ (resonant tank current phase shift), $I_R$ (resonant tank current magnitude), $S$ (switch turn-on slope), $V_{EX}$ (voltage amplitude dropped across resonant tank excess inductance), $V_S$ (voltage amplitude dropped across resonant tank resistance), $L_{EX}$ (excess inductance), and $C_P$ (shunt capacitance). Six simultaneous equations are derived in the analysis. The format of the analysis is similar to that from the previous section: an equation for $v(wt)$ is derived, then the equation is decomposed into $V_S$ and $V_{EX}$ components, and then $v(wt)$ is considered.

**A.3.1 Analysis of $v(wt)$ in General ZVS Condition**

At $wt = 2\pi$,

$$i_C(2\pi) = wC_P \left( \frac{d}{d(wt)} v(wt) \right)_{wt=2\pi} = wC_PS \neq 0$$  \hspace{1cm} (A.3.2)

Since the switch has just begun to conduct current at this time,

$$i_S(2\pi) = I_O + i_R(2\pi) - wC_PS = 0$$  \hspace{1cm} (A.3.3)
Let \( i_R(wt) = I_R \sin(wt + \phi) \). Then,

\[
I_O = -i_R(2\pi) + wC_PS = -I_R \sin(2\pi + \phi) + wC_PS = -I_R \sin(\phi) + wC_PS \quad (A.3.4)
\]

For \( \pi < wt < 2\pi \), in other words when the switch is off,

\[
i_C(wt) = I_O + I_R \sin(wt + \phi) = -I_R(\sin(\phi) - \sin(wt + \phi)) + wC_PS \quad (A.3.5)
\]

\[
\Rightarrow v(wt) = \frac{1}{wC_P} \int_{\pi}^{wt} i_C(wt) d(wt) \quad (A.3.6)
\]

\[
\Rightarrow v(wt) = -I_RX_P [\cos(wt + \phi) + \cos(\phi) + \sin(\phi)(wt - \pi)] + S(wt - \pi) \quad (A.3.7)
\]

where \( X_P = \frac{1}{wC_P} \). At \( wt = 2\pi \), \( v(wt) = 0 \), so

\[
v(2\pi) = -I_RX_P [2 \cos(\phi) + \pi \sin(\phi)] + \pi S = 0 \quad (A.3.8)
\]

\[
\Rightarrow \boxed{I_RX_P [\pi \sin(\phi) + 2 \cos(\phi)] = \pi S} \quad (A.3.9)
\]

### A.3.2 Decomposition of \( v(wt) \) in General ZVS Condition

From the previous subsection, \( v(wt) = -I_RX_P [\cos(wt + \phi) + \cos(\phi) + \sin(\phi)(wt - \pi)] + S(wt - \pi) \). The component of this voltage in phase and at the same frequency as the current \( i_R \) is dropped across the resistance \( R_{PRI} + R_{EQ} \), while the out-of-phase component is dropped across the inductance \( L_{EX} \). Thus,

\[
V_S = \frac{1}{\pi} \int_{\pi}^{2\pi} v(wt) \sin(wt + \phi) d(wt) \quad (A.3.10)
\]

Solving this integral gives

\[
V_S = \frac{I_RX_P}{2} \sin(2\phi) + \frac{2I_RX_P}{\pi} \cos(2\phi) - S \cos(\phi) + \frac{2S}{\pi} \sin(\phi) \quad (A.3.11)
\]
Setting $S = 0$ gives the same result as derived in Subsection A.2.2. The out-of-phase component magnitude is,

$$V_{EX} = \frac{1}{\pi} \int v(wt) \cos(wt + \phi) d(wt)$$  \hspace{1cm} (A.3.12)

Solving this integral gives

$$V_{EX} = \frac{I_R X_P}{2} \cos(2\phi) - \frac{2I_R X_P}{\pi} \sin(2\phi) + S \sin(\phi) + \frac{2S}{\pi} \cos(\phi) - I_R X_P$$  \hspace{1cm} (A.3.13)

Setting $S = 0$ gives the same result as derived in Subsection A.2.2.

### A.3.3 Analysis of $\overline{v(wt)}$ in General ZVS Condition

No average voltage is dropped across the choke inductor $L_{CHOKE}$, so $\overline{v(wt)} = V_{CC}$, or

$$V_{CC} = \frac{1}{2\pi} \int v(wt) d(wt)$$  \hspace{1cm} (A.3.14)

$$\Rightarrow V_{CC} = -I_R X_P \left( \frac{\pi}{4} + \frac{1}{\pi} \right) \sin(\phi) - \frac{I_R X_P}{2} \cos(\phi) + \frac{\pi}{4} S$$  \hspace{1cm} (A.3.15)

Setting $S = 0$ gives the same result as derived in Subsection A.2.3.

### A.3.4 Overall Analysis of General ZVS Condition

The last three subsections present four equations relating the seven unknown parameters of the power amplifier. A fifth equation relates the resonant tank current to the voltage dropped across the resonant tank resistance,

$$I_R = -\frac{V_S}{R_{PRI} + R_{EQ}}$$  \hspace{1cm} (A.3.16)

(The negative sign comes from the direction of the currents in Figure 4.6.) A sixth equation relates the relative voltages across the resonant tank excess inductance and resistance to their
relative impedances,
\[
\frac{V_{EX}}{V_{S}} = \frac{wL_{EX}}{R_{PRI} + R_{EQ}}
\] (A.3.17)

In summary, the six equations are

\[
\begin{align*}
\pi S &= I_{RXP} [\pi \sin(\phi) + 2 \cos(\phi)] \\
V_{S} &= \frac{I_{RXP}}{2} \sin(2\phi) + \frac{2I_{RXP}}{\pi} \cos(2\phi) - S \cos(\phi) + \frac{2S}{\pi} \sin(\phi) \\
V_{EX} &= \frac{I_{RXP}}{2} \cos(2\phi) - \frac{2I_{RXP}}{\pi} \sin(2\phi) + S \sin(\phi) + \frac{2S}{\pi} \cos(\phi) - I_{RXP} \\
\frac{V_{EX}}{V_{S}} &= \frac{wL_{EX}}{R_{PRI} + R_{EQ}} \\
I_{R} &= -\frac{V_{S}}{2} \frac{R_{PRI} + R_{EQ}}{2} \\
V_{CC} &= -I_{RXP} \left(\frac{\pi}{4} + \frac{1}{\pi} \right) \sin(\phi) - \frac{I_{RXP}}{2} \cos(\phi) + \frac{S}{\pi} 
\end{align*}
\] (A.3.18)

These equations can be used to solve for the values of the unknown variables. There are six equations and seven unknowns, so one of the variables is independent, let it be \(S\). The equations can be decoupled by the following manipulations (assuming \(S \neq 0\)).

First, rearrange the 1st equation as
\[
I_{RXP} = \frac{\pi S}{\pi \sin(\phi) + 2 \cos(\phi)}
\] (A.3.19)

Then plug it into the 6th equation to eliminate the \(I_{R}\) and \(X_{P}\) unknowns, obtaining
\[
\phi = \tan^{-1}\left(\frac{-2V_{CC}}{\pi V_{CC} + S}\right)
\] (A.3.20)

Next, plug Equation A.3.19 into the 2nd equation, obtaining
\[
V_{S} = \frac{\frac{2}{\pi} S \sin(2\phi)}{\pi \sin(\phi) + 2 \cos(\phi)}
\] (A.3.21)

Next, plug Equation A.3.19 into the 3rd equation, obtaining
\[
V_{EX} = \frac{\frac{2}{\pi} S \cos(2\phi) + (\frac{2}{\pi} - \frac{S}{\pi}) S}{\pi \sin(\phi) + 2 \cos(\phi)}
\] (A.3.22)
Dividing the last two equations,

\[
\frac{V_{EX}}{V_S} = \cot(2\phi) + \left(1 - \frac{\pi^2}{4}\right) \csc(2\phi)
\]  

(A.3.23)

which replaces the 4th equation. In the full ZVS case when \( \phi = \tan^{-1}\left(-\frac{2}{\pi}\right) = -32.48^\circ \), \( \frac{V_{EX}}{V_S} = \frac{\pi^3}{16} - \frac{\pi}{4} \approx 1.152 \), as derived in Subsection A.2.2. Using these new equations, the original set of equations is modified as

\[
\begin{align*}
\phi &= \tan^{-1}\left(\frac{-2V_{CC}}{\pi V_{CC} + S}\right) \\
V_S &= \frac{\frac{\pi}{2} S \sin(2\phi)}{\pi \sin(\phi) + 2 \cos(\phi)} \\
V_{EX} &= \frac{\frac{\pi}{2} S \cos(2\phi) + \left(\frac{\pi}{2} - \phi\right) S}{\pi \sin(\phi) + 2 \cos(\phi)} \\
\frac{V_{EX}}{V_S} &= \frac{\pi L_{EX}}{R_{PRI} + R_{EQ}} = \cot(2\phi) + \left(1 - \frac{\pi^2}{4}\right) \csc(2\phi) \\
I_R &= -\frac{V_S}{R_{PRI} + R_{EQ}} \\
X_P &= \frac{\pi S}{I_R (\pi \sin(\phi) + 2 \cos(\phi))}
\end{align*}
\]  

(A.3.24)

These equations may be used to determine the values of \( L_{EX} \) and \( C_P \), the Class-E power amplifier design parameters, given some \( R_{EQ} \). The general way to do this is as follows. To find \( L_{EX} \) and \( C_P \) given \( R_{EQ} \), first choose an \( S \). Then, use the 1st equation to determine \( \phi \). Next, use the 4th equation with knowledge of \( R_{EQ} \) to determine \( L_{EX} \). Next, determine \( V_S \) from the 2nd equation. Next, determine \( I_R \) from the 5th equation. Finally, determine \( X_P \) and thus \( C_P \) from the 6th equation. Note that since \( S \) is an independent variable, one of \( L_{EX} \) or \( C_P \) can have multiple values while the entire power amplifier still satisfies the general ZVS condition. This is an important degree of freedom that is exploited by the power amplifier control method to both optimize efficiency and balance delivered power.

To obtain a deeper understanding of these equations, the following graphs show specific power amplifier parameters with respect to the independent parameter \( S \).
Figure A.4 is a plot of the 1st equation, showing how the phase shift of the resonant tank current varies with the switch voltage slope at turn-on. The indicated point is when $\phi = \phi_{ZVS} = -32.48^\circ$, in other words, the phase shift determined in the previous section at which both ZVS assumptions are met (both switch voltage and slope are zero at switch turn-on). For negative slopes, the phase shift becomes more negative. Positive slopes do not exist because the switch voltage would have had to cross zero before the switch turn-on time.
Figure A.5: $|V_S|$ and $|V_{EX}|$ vs. $S$ in general ZVS

Figure A.5 is a plot of the 2nd and 3rd equations, showing how the components of the switch voltage that are dropped across the resonant tank excess inductance and the resistance vary with the turn-on slope $S$. The indicated points are when $\phi = \phi_{ZVS}$. At this point, $|V_S| = \frac{1}{\sqrt{\pi^2 + 4}} V_{CC} \approx 3.544$ V, which is consistent with analysis in the previous section. Also, $\frac{V_{EX}}{V_S} = \frac{\pi^2}{8} - \frac{\pi}{4} \approx 1.152$, which is consistent with previous analysis. For negative slopes, the resonant tank voltage becomes more out of phase with the resonant current, so more voltage is dropped across the excess inductance, hence the lower $V_S$ and higher $V_{EX}$ in the plot.
Figure A.6: $\frac{X_{EX}}{R_{PRI}+R_{EQ}}$ vs. $S$ in general ZVS

Figure A.6 is a plot of $\frac{X_{EX}}{R_{PRI}+R_{EQ}} = \frac{V_{EX}}{V_S}$ vs. $S$, where $X_{EX} = wL_{EX}$. It is evident that given a value for $R_{EQ}$, picking different values of $S$ results in different ratios of $V_{EX}$ to $V_S$ and in different values of $L_{EX}$ that satisfy the general ZVS condition.

Figure A.7: $\frac{X_P}{R_{PRI}+R_{EQ}}$ vs. $S$ in general ZVS

Figure A.7 is a plot of $\frac{X_P}{R_{PRI}+R_{EQ}} = -\frac{\pi S}{V_S(\pi \sin(\phi)+2\cos(\phi))}$ vs. $S$. It is evident that given a
value for $R_{EQ}$, picking different values of $S$ results in different values of $C_P$ that satisfy the general ZVS condition.

### A.4 Simulation Results

The simulation results in this section give more examples showing how the proposed feedback and compensation control method maintains zero power amplifier switch drain voltage at switch turn-on over various coupling and output voltage conditions. Figure 4.1 shows the circuit and parameters used for simulation. $L_1$ and $C_P$ (sum of $C_3$ and parasitic switch and diode output capacitances) are the tunable parameters and $k$ are $V_{OUT}$ are the parameters representing the changing conditions.

![Figure A.8: Switch drain voltage for $L_1 = 5$ uH, $C_P = 80$ pF system, $k = 0.25$, $V_{OUT} = 3.8$ V conditions](image)

With the original parameters, $k$ is increased to 0.25. This affects the ZVS condition as shown in Figure A.8. Mathematically, $R_{EQ}$ is increased (the primary side sees more loading), so the resonant current phase shift $|\phi|$ is decreased. In this case, the dynamic system compensates by increasing $L_1$ and decreasing $C_P$.

![Figure A.9: Switch drain voltage for $L_1 = 5.1$ uH, $C_P = 50$ pF system, $k = 0.25$, $V_{OUT} = 3.8$ V conditions](image)

Figure A.9 shows the compensated system, which is now much closer to the ZVS condition.
Figure A.10: Switch drain voltage for $L_1 = 5\ \text{uH},\ C_P = 80\ \text{pF}$ system, $k = 0.2$, $V_{OUT} = 5\ \text{V}$ conditions

With the original parameters, $V_{OUT}$ is increased to 5 V. This affects the ZVS condition as shown in Figure A.10. Mathematically, $R_{EQ}$ is decreased, so the resonant current phase shift $|\phi|$ is increased. In this case, the dynamic system compensates by decreasing $L_1$ and increasing $C_P$.

Figure A.11: Switch drain voltage for $L_1 = 4.9\ \text{uH},\ C_P = 160\ \text{pF}$ system, $k = 0.2$, $V_{OUT} = 5\ \text{V}$ conditions

Figure A.11 shows the compensated system, which is now much closer to the ZVS condition.

Figure A.12: Switch drain voltage for $L_1 = 5\ \text{uH},\ C_P = 80\ \text{pF}$ system, $k = 0.2$, $V_{OUT} = 3\ \text{V}$ conditions

With the original parameters, $V_{OUT}$ is decreased to 3 V. This affects the ZVS condition as shown in Figure A.12. Mathematically, $R_{EQ}$ is increased, so the resonant current phase shift
$|\phi|$ is decreased. In this case, the dynamic system compensates by increasing $L_1$ and decreasing $C_P$.

![Graph showing switch drain voltage](image)

Figure A.13: Switch drain voltage for $L_1 = 5.1$ uH, $C_P = 50$ pF system, $k = 0.2$, $V_{OUT} = 3$ V conditions

Figure A.13 shows the compensated system, which is now much closer to the ZVS condition.
Appendix B

Wireless Charging Controller Chip

Pinout

Figure B.1: Chip pinout diagram

Figure B.1 shows the pinout of the wireless charging controller chip. Table B.1 is a description of the pin functionality of the microchip.
<table>
<thead>
<tr>
<th>Pin</th>
<th>Description</th>
<th>Type</th>
<th># Leads</th>
</tr>
</thead>
<tbody>
<tr>
<td>VCC</td>
<td>3.3 V supply</td>
<td>Supply</td>
<td>4</td>
</tr>
<tr>
<td>VDD</td>
<td>1.8 V supply</td>
<td>Supply</td>
<td>5</td>
</tr>
<tr>
<td>GND</td>
<td>Ground</td>
<td>Supply</td>
<td>11</td>
</tr>
<tr>
<td>IN</td>
<td>Input to analog block</td>
<td>Analog input (3.3 V)</td>
<td>1</td>
</tr>
<tr>
<td>VREF</td>
<td>Voltage reference for analog block</td>
<td>Analog input (3.3 V)</td>
<td>1</td>
</tr>
<tr>
<td>OSC</td>
<td>6.78 MHz clock</td>
<td>Digital input (3.3 V)</td>
<td>1</td>
</tr>
<tr>
<td>RSTB</td>
<td>Active-low chip reset</td>
<td>Analog input (1.8 V)</td>
<td>1</td>
</tr>
<tr>
<td>DRIVE</td>
<td>Power amplifier switch drive</td>
<td>Digital input (3.3 V)</td>
<td>1</td>
</tr>
<tr>
<td>Z, A, B, C, D</td>
<td>Control compensation setting</td>
<td>Digital output (1.8 V)</td>
<td>5</td>
</tr>
<tr>
<td>ABV0</td>
<td>Input voltage is above zero</td>
<td>Digital output (1.8 V)</td>
<td>1</td>
</tr>
<tr>
<td>BLW3</td>
<td>Input voltage is below VREF</td>
<td>Digital output (1.8 V)</td>
<td>1</td>
</tr>
<tr>
<td>CLK</td>
<td>Analog block clock</td>
<td>Digital output (1.8 V)</td>
<td>1</td>
</tr>
<tr>
<td>A/B/C/D Cap</td>
<td>Control pins for switched capacitors</td>
<td>I/O (high voltage)</td>
<td>4</td>
</tr>
<tr>
<td>Z/A/B/C/D Tap</td>
<td>Control pins for multi-tap coil relays</td>
<td>I/O (high voltage)</td>
<td>5</td>
</tr>
<tr>
<td>NC</td>
<td>No connect</td>
<td></td>
<td>6</td>
</tr>
</tbody>
</table>

Table B.1: Description of wireless charging controller pinout
Bibliography


[37] Alliance For Wireless Power.

[38] iFixit. Fitbit Flex teardown, 2013.


