THREE DIMENSIONAL WIRELESS CHARGING SYSTEM WITH FLEXIBLE RECEIVER ALIGNMENT

By

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To my family
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<tr>
<td>3-D</td>
<td>Three-Dimensional</td>
</tr>
<tr>
<td>A4WP</td>
<td>Alliance for Wireless Power</td>
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<tr>
<td>PTU</td>
<td>Power Transmitting Unit</td>
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<td>PRU</td>
<td>Power Receiving Unit</td>
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<tr>
<td>PTE</td>
<td>Power Transfer Efficiency</td>
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<td>RX</td>
<td>Receiving</td>
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<td>TX</td>
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<td>VNA</td>
<td>Vector Network Analyzer</td>
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<td>WPT</td>
<td>Wireless Power Transfer</td>
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THREE DIMENSIONAL WIRELESS CHARGING SYSTEM WITH FLEXIBLE RECEIVER ALIGNMENT

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Chair: Jenshan Lin
Major: Electrical and Computer Engineering

This dissertation begins with the description and performance analysis of a traditional planar wireless charging system for a single consumer electronic device. To overcome the limitations of such planar structures, three-dimensional (3-D) transmitting coil arrangements suitable for different applications are proposed and tested both numerically and experimentally at 6.78 MHz, the industrial, scientific and medical (ISM) frequency band used in AirFuel specification.

The first 3-D arrangement allows free positioning of a wearable device about a 360° axis of rotation and simultaneous charging of a mobile phone. The 3-D transmitting coil structure provides a cylindrical charging surface with a uniform magnetic field with less than 0.02 coupling variation. The prototype is able to deliver 1 W to a wearable device and 5 W to a mobile phone simultaneously, achieving a power transfer efficiency of 48%, and up to 79.5% coil-to-coil efficiency.

The second 3-D coil arrangement is designed as part of a charger system with a cup-shaped structure. This configuration provides free alignment for portable devices by maximizing the magnetic flux perpendicular to each internal surface of the charger. Besides, a flexible ferrite sheet surrounding is wrapped around the exterior walls to
prevent the leakage of magnetic flux and comply with the electromagnetic compatibility (EMC) requirements.

Both systems described above are a significant improvement to traditional planar wireless chargers. However, they still may present some limitations depending on the positioning of the receiving coil. Therefore, a third 3-D arrangement is proposed in this dissertation. This third charging system significantly improve flexibility in the positioning of the receiving coil by independently adjusting the phase and the amplitude of the input signals, eliminating the blind spots within a cylindrical volume. Moreover, the symmetric arrangement of the transmitting coils yields a uniform magnetic field around the longitudinal axis. A class-E power amplifier with an arrangement of parallel matching networks is also proposed to provide constant current over a wide range of load variations.

Finally, conclusions obtained from experimental and simulation results for the three wireless charging systems are presented. Ideas and suggestions for future applications and/or improvements are also discussed in the last chapter.
1.1 History of Wireless Power Transfer

Over the last decade, the interest in wireless power transfer has risen dramatically. However, the history of scientists trying to remove the power cord can be traced back to the 18th century [1]. Starting with the first experimental studies that lead to discovery of Ampere’s Law [2] and Biot-Savart Law in the 1820s [3], Faraday’s Law in the 1830s [4]; followed by the unification of the electromagnetic theory through Maxwell’s equations in the 1860s [5]. It was Heinrich Rudolf Hertz’s who first proved Maxwell’s hypothesis using an induction coil connected to an oscillator to spread electromagnetic waves wirelessly through series of sparks [6]. This first attempt of wireless transmission was only across a tiny air gap and occurred when Hertz turned on the oscillator. Later on, Nikola Tesla, the pioneer of modern electricity and the proponent of wireless power transfer [7], performed many experiments at his Colorado Springs laboratory. Fig. 1-1 shows one of his famous demonstrations, which consisted in using three-meter metal balls to transmit electrical sparks between sixty-feet distance [7]. Nevertheless, due to the poor power transfer efficiency, the concept was abandoned until the 1950’s. Thanks to the development of microwave theory, experiments of wireless power transfer in far field started to be carried out once more. It was not until 2001, Boys and Green at the University of Auckland proposed an inductive power transfer (IPT) system where components can be identified and improved separately [8][9]. The arrangement of a general IPT system, is shown in Fig.1-2. However, the
operating frequency is limited below 10 kHz in their work. The major breakthrough took place in MIT in 2006, when the research group led by Prof. Marin Soljačić [10]. They successfully achieved the wireless transmission of 60 W of power using a pair of coils resonating around 9 MHz. Since then, WPT technology has been widely applied to industry, including portable consumer electronics and electrical vehicles. Meanwhile, several specification standards have been proposed and followed by many companies [11-15]. Among them, the largest and most influential alliances are Qi [11], Wireless Power Consortium (WPC) [12], Power Matters Alliance (PMA) [13] and Alliance for Wireless Power (A4WP) [14], which are synonyms of the two main techniques used in near-field wireless power transfer: Inductive coupling and magnetic resonance, respectively. In 2015, with the rise requirement of user experience, the last two organizations merged to form a new Alliance, AirFuel Alliance [15], which can
support dual-mode charging. The standards and accordingly specifications are listed in Table 1-1.

![System Block Diagram]

**Figure 1-2.** First near-field wireless power transfer experiment using inductive coupling [8].

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<td><strong>Mode</strong></td>
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<td><strong>Frequency</strong></td>
</tr>
<tr>
<td><strong>Support Alliances</strong></td>
</tr>
<tr>
<td><strong>Advantages</strong></td>
</tr>
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### 1.2 System Block Diagram

A near-field wireless power transfer system usually contains three main parts:

1. PTU and PRU coils form a loosely or strongly coupled transformer
2. PTU converts DC power to AC power and drive coils to generate magnetic field
3. PRU receives AC power and converts it to a stable DC output voltage to charge batteries.

![Diagram of a typical wireless charging system](image)

Figure 1-3. A typical wireless charging system [16]

Fig. 1-3 shows a typical block diagram of the whole system, which includes: transmitting coil, power amplifier, and matching networks in PTU; receiving coil, matching network, rectifier, and low-dropout voltage regulator in PRU [16]. Unlike far-field wireless power transfer, near-field charging is highly sensitive to the size and placement of receiving devices as well as the charging status of the battery. More detailed techniques of how to overcome the misalignment and environmental issues while supporting multiple devices simultaneously will be proposed and analyzed in chapter 2, 3, and 5. In the next few sections, discussion will be focused on theories and conventional structures of important blocks.

1.3 Transformer

When it goes to consumer electronics charging, a larger received power (>100 mW) is usually required. Therefore, in order to meet FCC and RF safety regulations [17], near-field wireless power transfer is a more practical choice without causing
interference to other devices. As mentioned before, near-field WPT terminologies can be divided into two categories based on its coupling and transmission distance.

1.3.1 Inductive Coupled Transformer

In this configuration, power is transferred through a pair of tightly coupled coils. Which means, TX and RX coils are usually close match in size and shape, and the transfer distance between TX and RX is much shorter than the diameter of coils itself, as shown in Figure 1-4. Therefore, inductive coupling requires precise alignment to achieve competitive power transfer efficiency. For consumer electronics design, industry usually use magnets or physical constraints to maintain precise alignment, for instance, a smart watch charger or a tooth brush charger [18] [19].

![Tightly coupled wireless power transfer](image_url)
1.3.2 Magnetic Resonant Transformer

On the contrary, in a magnetic resonant wireless charging system, TX coils are designed to operate at the resonant frequency of receiving coil with much longer distance apart. The magnetic field density is enhanced and concentrated within RX coil area through the resonator, as shown in Figure 1-5. As a result, this solution provides the freedom of devices positioning. Moreover, due to the spatial freedom and stronger magnetic field, this method can charge multiple devices simultaneously with higher power transferred. Because of the longer transmission distance and lower coupling coefficient required, this method also called as loosely-coupled transformer.

Figure 1-5. Loosely coupled wireless power transfer
1.3.3 Comparison of Tightly Coupled and Loosely Coupled WPT

Figure 1-6 shows the coil-to-coil efficiency versus the distance between TX and RX [20]. When the distance is less than 10 mm, both methods can achieve higher than 90% efficiency. However, when the distance increases, the decrease of efficiency is much slower with the loosely coupled resonant no matter where the placement is. This makes loosely coupled WPT a better candidate when we want to charge multiple receivers from a single primary coil. That is also why magnetic resonant system becomes more favorable recently when users are expecting to charge more than just a smart phone, as shown in Figure 1-7 [21]. Nevertheless, due to higher electromagnetic emission, loosely coupled system is less suitable for the application with strict EMI or

![Simulated efficiency comparison of tightly coupled inductive and loosely coupled resonant coil to coil system. [20]](image)

Figure 1-6. Simulated efficiency comparison of tightly coupled inductive and loosely coupled resonant coil to coil system. [20]
EMF requirements, for instance, in-vehicle wireless charging. More detailed design considerations will be discussed in Chapter 3.

1.4 **Switched Power Amplifier**

Power transfer efficiency is crucial for wireless power transfer system. Because of that, a switched power amplifier (PA) is more suitable than a linear amplifier to provide the required output power while minimizing power dissipation [22-24]. Among all types of switched power amplifier, class-D and class-E are most common architectures adopted for wireless charging system, especially for consumer electronics application. The theory and comparisons of these two typical structures are discussed in the following.

1.4.1 **Class-D Power Amplifier**

Fig. 1-8 (a) shows an ideal single-ended class-D power amplifier [25]. The series inductor-capacitor (LC) tank is connected as a filter that only allows fundamental current
flowing into the load. Besides, two transistors are used to realize the square waveform. Therefore, the power dissipation is zero if the switches are assumed as ideal.

Figure 1-8. Class-D power amplifier. (a) Half-bridged (single-ended) structure, (b) ideal voltage and current waveform [25]
However, in real case scenario, two parasitic capacitances of the switch, $C_{ds1}$ and $C_{ds2}$, limits the efficiency of a class-D power amplifier. The power dissipation due to this capacitance is

$$P_{dis} = \frac{fV_{DC}^2C_{ds}}{2}$$

where $f$ is the operational frequency.

Since this power loss is proportional to the frequency, this structure is not suitable for higher frequency application. Moreover, comparing to class-E power amplifier, class-D structure needs one more switch, which makes it harder to synchronize when the switching frequency is too high.

1.4.2 Class-E Power Amplifier

A typical single-ended Class-E power amplifier is shown in Fig. 1-9 (a) [26], where the transistor is treated as an ideal switch and Cds is added in parallel. When the power transistor is turned on, the drain voltage of the transistor and the derivative of the voltage are forced to zero with zero slope to obtain non-overlapped drain voltage and current waveforms, as illustrated in Fig. 1-9 (b) [26]. Therefore, by removing the drain capacitance losses in Class-D power amplifier, Class-E circuit can achieve 100% efficiency theoretically [27]. Key design concepts are introduced in [22-23, 25-27], where maximum 94% efficiency can be achieved for a 13.56 MHz WPT system by optimizing the load network [22]. However, the main challenge of this structure is the high sensitivity to the output impedance, which means that the zero-voltage switching condition occurs only at certain load resonance value [28][29]. When a transmitting system supports more than one receiving device, the load variation may easily exceed
its tolerance, which will cause hard-switching phenomenon and the power consumption increases.

Figure 1-9. Class-E power amplifier. (a) Single-ended structure, (b) ideal voltage and current waveforms. [26]
1.4.3 Comparison of Class-E and Class-D Power Amplifier

Since magnetic resonant structure operates under weak coupling coefficient region, to compensate the loss between the transformers, class-E power amplifier is used in this dissertation due to its higher power transfer efficiency. Although an extra RF chock inductor is required, it can be easily integrated with transmitting circuit board without causing further area. Table 1-2 shows the comparison of two structures [30].

<table>
<thead>
<tr>
<th>Criterion</th>
<th>Class-D</th>
<th>Class-E</th>
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<tr>
<td>Maximum Output Power (Ideally)</td>
<td>$\frac{2V_{in}^2}{\pi^2 R}$</td>
<td>$\frac{8V_{in}^2}{(\pi^2 + 4)R}$</td>
</tr>
<tr>
<td>Drain Voltage Tolerance</td>
<td>$V_{in}$</td>
<td>$\geq 2.5V_{in}$</td>
</tr>
<tr>
<td>Gate Switch Losses</td>
<td>$2C_gV_{in}^2 f$</td>
<td>$C_gV_{in}^2 f$</td>
</tr>
<tr>
<td>Ideal Switch Behavior</td>
<td>zero current switching</td>
<td>zero voltage switching</td>
</tr>
<tr>
<td>Minimum active components</td>
<td>2 transistors</td>
<td>1 transistor</td>
</tr>
<tr>
<td>(Single-ended)</td>
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1.5 Challenges and Motivations for Consumer Electronics Applications

Small, battery-powered electronic devices that require less than 1 W to operate have proliferated in recent years. Wearable devices, the “Internet of Things”, and implantable medical devices all belong to this category. A common feature of these devices is that it can be impractical or impossible to connect a dc voltage supply to recharge the battery. Therefore, there is significant interest in non-contact charging methods such as resonant wireless power transfer (WPT). WPT has been
demonstrated for implantable devices, remote sensors, and wearable devices [31-35]. The ideal WPT system would provide a charging rate similar to that which can be obtained using a conventional dc supply, with high power transfer efficiency. In addition, it would provide a high degree of spatial freedom, so as to provide the user with an effortless charging experience. In practice, it is difficult to achieve both objectives. Systems that provide high efficiency typically require precise alignment of charger and device, while systems that offer free positioning tend to suffer from low charging rate and/or low efficiency. Furthermore, much of the research in the mobile electronics space has focused on planar devices such as mobile phones and tablets [36-41], whereas wearable devices often have unusual shapes that may be awkward for the user to place on a planar charging pad.

As a result, some studies with non-planar WPT system with multiple transmitting coils has been studied to improve the freedom of positioning [42-44]. In [42], a three crossed-dipole coil receiver with open-ended structure is demonstrated inside a larger TX structure. The design concept of two orthogonal transmitter coils are similar to RX coils but connected in series and driven with the same ac current. The analysis showed that the receiver is successfully powered no matter what the orientation is. However, this alignment-insensitive characteristic can be achieved when the receiver has three orthogonal coils, which is against the feature of most consumer electronic products. In [43], a 3-D TX coils structure is proposed to generate omni-directional magnetic field. With similar design concept as it is presented in [42], the TX coils in this paper are also constructed with orthogonal orientation. By activating ac current with different phases, magnetic field can be generated in all directions within the enclosed glass charging ball.
Nevertheless, this enclosed system structure is against the user intuition, which makes it unsuitable for practical application. In [44], a custom-designed inductive power transfer (IPT) box system is discussed. By feeding non-identical signals into the cubical-shaped primary TX coils, a rotating magnetic flux flow is generated. Though the traveling magnetic field can produce omni-directional flux vectors, the distribution of magnetic field strength is varied greatly at different position within the created power zone. Hence, the overall power transfer efficiency becomes worse when the size and the required power lever of supported receiving devices are with different order.

On the other hand, some simpler nonplanar charging system with a single transmitting coil are proposed [45-46]. In [45], a helix-shaped primary coil is proposed to wirelessly power and collect biological data from a sensor chip. Though this work investigated the received power of a telemetric device located on a freely moving rodent, the magnetic field vector in horizon direction is much weaker than it is in vertical direction. Which restricted the alignment of the secondary coil. In [46], a bowl-shaped transmitting coil is proposed for hearing aids. With slightly tilt along the sidewall, this work improved the disadvantage in [45] and generated uniform magnetic field in $\rho$ and $z$ direction simultaneously, which is perpendicular to possible alignments of a receiving coil. Therefore, the overall power transfer efficiency of this work is increased. However, the bowl-shaped TX coil also created a dead-zone near the center of the structure. Even though the blind area is narrow and unlikely to happen for its application in [46], the efficiency becomes worse when charging multiple devices simultaneously for other applications. More importantly, prototypes demonstrated in [45] and [46] are designed
for the RFID applications that only require micro-watt power level. It is not adequate for the devices with higher battery power, such as a cell phone or a smart watch.

Therefore, three 3-D TX structures are proposed and verified in the following chapters. Primary coils with cylindrical shape successfully generated uniform magnetic field around the charging area. Due to different applications, the EMI and EMC requirements are also considered and improved. All the charging systems are demonstrated at 6.78 MHz, the industrial, scientific and medical (ISM) frequency band used in AirFuel magnetic resonance standard [15].
2.1 Limitation of Traditional Structure

Since most commercialized chargers are designed in a planar shape, to understand its limitation for charging smaller devices, an experiment is conducted using a certificated planar coil [47]. The experiment setup is shown in Fig. 2-1 (a). A smart watch is placed on the planar charger and scanned over the whole charging area. However, the experiment results show that the magnetic field is non-uniform for a wearable receiving coil, as plotted in Fig.2-1 (b). There are some null-points within the charging area. Another issue is power transfer efficiency. In Fig. 2-1 (a), the leakage of magnetic flux becomes more serious because of the smaller receiving device. According to preliminary measurement results, power transfer efficiency for a single wearable device on a single-phone charging pad is less than 25%.

Another experiment was done with a cell phone and a wearable device simultaneously using the same certificated planar coil, as shown in Fig. 2-2. A cell phone was put on the transmitting coil and charged with a 5 W battery, while the wearable device was scanned over the rest charging area. Although the wearable device can receive more than 0.4 W in most area, the EM field distribution was still non-uniform. Besides, input port of the charging pad, which is a SMA connector in this setup, highly affected the placement of the wearable device.

Therefore, to overcome above limitations, a fully-customer designed 3-D structure is proposed to charge a wearable device and a cell phone.
Figure 2-1. Measurement of a wearable receiving coil vs. a planar wireless charging system. (a) The measurement setup. (b) The coupling coefficient vs. difference positions.
Figure 2-2. Use a planar charging pad to charge a cell-phone and a wearable device.

2.2 Coil Design

2.2.1 Transmitting Coil

In order to support simultaneous charging of a mobile phone and a smart watch, a three-dimensional (3-D) coil structure is proposed in this dissertation, as shown in Fig. 2-3. The area between the upper turns and base turns along the sidewall of the cylinder is designed for wearable device charging, while the top spiral coil is designed for mobile phone charging. The magnetic flux distribution generated by the coil was simulated using ANSYS HFSS and plotted in Fig. 2-4(a).
According to Ampere’s “right-hand rule”, two parallel wires carrying currents in opposite directions generate a magnetic field in the same direction. In the proposed design, the current in the base turns flows opposite to the current in the upper turns. This enhances the magnetic field in the sidewall area where the wearable device rests. The vertical wire connecting the upper turns to the base turns generates a magnetic field that can cause an asymmetry in the field normal to the charging surface. To minimize the impact of this vertical wire on the charging field, it was placed inwards, towards the central axis of the coil. This makes the asymmetry effect negligible, as shown in Fig. 2-4 (a).

Fig. 2-4 (b) shows an EM simulation result for the top spiral coil. In order to allow simultaneous charging of the wearable device and the phone, the top spiral coil was
Figure 2-4. Magnetic field strength (a) in wearable device charging area, in A/m, (b) in mobile phone charging area, in A/m.

designed to produce the same magnetic field strength as the sidewall area for the same coil current. The magnetic field is highly uniform across the top charging surface,
mimicking the characteristics of an ideal solenoid. The coil current for both simulations was 1 A. In normal operation, coil current is adjustable in a range close to this value.

The transmitter coil was fabricated by winding 16-gauge wire around a 3-D printed plastic coil form. The measured inductance and quality factor of the transmitter coil were 1.30 $\mu$H and 250, respectively. A ferrite sheet was placed underneath the coil to shield the electronics in the base of the charger.

### 2.2.2 Receiving Coil for a Wearable Device

The design constraints for the receiving coil were more restrictive, as space was limited within the smart watch construction. A planar coil structure with a size of 29x31 mm$^2$ was used, as shown in Fig. 2-5. The receiver coil was optimized to provide a large enough mutual inductance to keep the transmitter coil current reasonable, while

![Layout of the wearable device receiving coil.](image)

Figure 2-5. Layout of the wearable device receiving coil.
maintaining a high quality factor to minimize receiver-coil losses. A 2-oz copper PCB spiral coil with 9 turns was fabricated using 300 μm traces, spaced 275 μm apart. The coil achieved a quality factor of 84 with an inductance of 4.99 μH. The self-inductance and the resonant frequency of the receiving coil is simulated in Fig. 2-6.

### 2.2.3 Coupling Coefficient

The coupling coefficient of the coil pair was extracted from two-port S-parameter measurements on a vector network analyzer. Fig. 2-7 shows the scan pattern for our coupling measurements in addition to the measured data. The measured coupling coefficient is between 0.07 and 0.09 throughout the charging area. For comparison, we also scanned the wearable receiving coil (RX) in a serpentine pattern across a planar spiral transmitting coil. The resulting scan of coupling coefficients, also shown in Fig. 2-7, illustrates the advantageous coupling uniformity of the 3-D transmitter coil.

![Inductance Graph](image)

Figure 2-6. Resonant frequency of receiving coil is three times away from the operating frequency.
Figure 2-7. Coupling coefficient measured as the wearable receiving coil (RX) was scanned across the proposed 3-D transmitter coil and a planar transmitter coil for 36 different positions.

2.3 System Design

A Power Transmitting Unit (PTU) and Power Receiving Units (PRU) were designed to demonstrate the operation of the coil pair. The PTU used a full-bridge Class-E amplifier operating in a zero-voltage-switching mode followed by an LCL-π
network, as shown in Fig. 2-8. The LCL-π matching network serves to hold the transmitter coil current roughly constant with respect to PRU placement and load conditions [48]. More explains on how to provide constant current will be discussed in Chapter 5. The PRUs used the multi-mode wireless power IC described in previous work [16].

![Diagram of LCL-π matching network]

Figure 2-8. Resonant frequency of receiving coil is three times away from the operating frequency.

In the case of the wearable device PRU, this IC was paired with the wearable device coil and operated in the 6.78 MHz mode. The smart phone PRU used the multi-mode coil described in previous work [16]. Both PRUs communicate to the PTU using load modulation. As in a Rezence system, the PTU monitors status messages from the PRUs and adjusts coil current to optimize power transfer for both devices.

### 2.4 Implementation and Assembly of Housing

For a heavy wire such as a thickness of 16-gauge wire using in our design, it is not easy to wind it along a cylinder and fix it on a 3-D structure. Therefore, a special
housing with notches on the surface is designed and printed by a 3-D printer for our application, as shown in Fig. 2-9.

In Fig. 2-9 (a), grooves on the top plane and the sidewall are designed to fit in a 16-guage copper wire. The depth and the width of grooves are calculated and adapted for the wire with the adjustable tolerance. The notch on the top plane is used for regulating the wire across two different dimensions. In Fig. 2-9 (b), the cavity in the base of the housing is saved for the printed circuit board and other power electronics. The input port of the power line is designed to connect the TX coil in the shortest path to minimize the undesired power loss. In addition to the PCB and electronic components, the cavity is also designed for placing the ferrite sheets to shield the components from electromagnetic interferences caused by above cooper wires. Fig. 2-9 (c) shows the cover of housing outside of the transmitter coil. The tilt and the ridge are designed for the intuitive positioning of the wearable devices. When a wearable device is placed on
the charger, the tilt make the device naturally attach to the sidewall to enhance the power transfer efficiency. Furthermore, the thickness of the case is not only used to cover the copper wire, but also to prevent the interference from two resonators. Notably, to avoid any unwanted impedance variation caused by the environment, a plastic printing material is used for the proposed housing with the resistance from water or temperature damage.

A photograph of the WPT system with a smart watch and a mobile phone is shown in Fig. 2-10. The power transfer efficiency of the system including amplifier, rectifiers, regulators and coil losses is 48% for the combination of the mobile phone and smart watch charging at a rate of 5 W and 1.15 W, respectively. When charging the watch alone, the power efficiency is relatively low because of fixed losses in the transmitter amplifier, which is designed to support up to 10 W power delivery to load devices. Coil-to-coil efficiency for charging the smart watch is estimated to be 79.5%.

Figure 2-10. Resonant frequency of receiving coil is three times away from the operating frequency. Photo courtesy of Patrick Riehl.
A comparison of the performance of this system with previously published work in terms of PTE, power delivered to loads (PDL) and spatial freedom is given in Table 2-1. Typically, efficiency is quoted as “DC-to-DC”, with input power measured at the PTU amplifier input and output power measured at the PRU rectifier output. In this work, output power was measured at the regulated 5 V output.

### 2.5 Summary

A resonant WPT system that supports simultaneous charging of a smart watch and a mobile phone is demonstrated. A unique 3-D structure provides a uniform 360° charging surface, enabling free positioning of the smart watch. The power transfer efficiency when simultaneously charging a smart watch and a mobile phone is 48%.

<table>
<thead>
<tr>
<th>Device type</th>
<th>Max PDL</th>
<th>Max PTE</th>
<th>PTE terminals</th>
<th>Spatial Freedom</th>
</tr>
</thead>
<tbody>
<tr>
<td>Implantable sensor</td>
<td>2.4 mW</td>
<td>24%</td>
<td>DC-to-DC</td>
<td>20 cm distance, 30 deg tilt</td>
</tr>
<tr>
<td>Hearing aid</td>
<td>28 mW</td>
<td>28%</td>
<td>DC-to-DC</td>
<td>360 deg, inside bowl</td>
</tr>
<tr>
<td>Test structures, open air</td>
<td>275 mW</td>
<td>37%</td>
<td>coil-to-coil</td>
<td>Axially aligned, 1.1 coil diameter spacing</td>
</tr>
<tr>
<td>Mobile phone</td>
<td>3 W</td>
<td>60%</td>
<td>DC-to-DC</td>
<td>Planar charging surface, 32x22 cm²</td>
</tr>
<tr>
<td>Smart watch + Mobile phone</td>
<td>1W + 5W</td>
<td>48%</td>
<td>DC-to-reg. 5V</td>
<td>360 deg cylindrical charging surface</td>
</tr>
</tbody>
</table>

Table 2-1. Comparison of WPT Systems with PDL < 10 W and Increased Spatial Freedom
CHAPTER 3
IN-VEHICLES WIRELESS CHARGING SYSTEM FOR PORTABLE DEVICES

3.1 Introduction

In-vehicles wireless charging is another growing application, as data shows in Fig. 3-1, that the market size will achieve 106 million in 2020 [52]. Unlike the in-door charging, the challenge of in-vehicles charging is the serious misalignment of the receiving device. Furthermore, to avoid any interference with other in-car communication, EMI and EMC issues of this application need to be considered as well. Therefore, an adapted structure from Chapter 2 is discussed in the following.

Figure 3-1. Automobile companies provide vehicles with wireless charging function [52].
3.2  Coil Design

3.2.1  Transmitting Coil

In order to support simultaneous charging multiple devices with freedom of positioning and user-friendly experiences, a three-dimensional (3-D) transmitting coil is presented in this chapter, as shown in Fig. 3-2. Any in-vehicle charger must account for the possibility that the position of the charging device may shift due to the acceleration or deceleration of the vehicle. In order to avoid the null points inside the charger, it is essential to generate the magnetic flux perpendicular to every charging plane. Therefore, the receiving device could easily capture sufficient magnetic flux to maintain competitive power transfer efficiency. The area between the top turns and base turns along the sidewall of the cylinder is designed for charging ring-shaped devices, for instance, a smart watch or a fitness band. On the other hand, the base spiral coil is

Figure 3-2. The 3-D structure of cup-shaped wireless charger with the dimension of 65x85 mm².
designed to generate the magnetic flux normal to the bottom plane and to support the device either standing with an angle from the sidewall or lying down at the bottom, as shown in Fig. 3-3 (a)-(e). The magnetic flux distribution generated by the coil was simulated using ANSYS HFSS and plotted in Fig. 3-3. Three possible alignments will be discussed in the following sections.

### 3.2.2 Simulation Results

Figs. 3-3(a)(b) show the simulation result for the bottom spiral coil. In order to allow simultaneous charging of relatively small receiving devices on the bottom surface,
the central turns of the base spiral coil were removed to produce the similar magnetic flux density over the whole charging area. Therefore, the magnetic field becomes uniform without any weaker area in it, which is suitable for further applications with even smaller receiving coils, such as Bluetooth headsets or hearing aids.

According to Ampere’s “right-hand rule”, two parallel wires carrying currents in opposite directions generate a magnetic field in the same direction [2]. In the proposed design, the current in the base turns flows opposite to the current in the upper turns. This not only provides the magnetic flux toward the center, but also enhances the magnetic field between two opposite coil turns, as shown in Fig. 3-3 (c)(d). By adjusting the pitch and the turns of the transmitting coil, it gives the tolerance of placement and angle inside the structure. In order to reduce the asymmetry effect caused by the vertical wire and also to fit it into the usual cup holder in vehicles, the vertical wire connecting the top turns to the base turns was placed outwards, away from the charging plane. This makes the asymmetry effect negligible.

Figs. 3-3 (e)(f) illustrate another possibility of the placement of the receiving device, which is standing with an angle \( \theta \) between the device plane and the bottom plane. From the simulation results, though the magnetic field is not as uniform as the first two cases, it could capture sufficient magnetic flux with relatively larger coils. The receiving coil used in the simulation is similar to the wearable RX in [48], which has an area of 29x31 mm\(^2\) in a square shape.
3.2.3 Shielding Ferrite for EMI Suppression

Due to stringent restrictions on electromagnetic interference (EMI) in vehicles, a piece of flexible ferrite sheet is used to surround the cup-shaped charger, as shows in Fig. 3-4. The leakage of magnetic flux is reduced down to $2.29 \times 10^{-3}$ A/m outside of the structure. It can also improve the electromagnetic interference from other electronic systems in vehicles. Moreover, this ferrite sheet protects the charging system from unexpected impedance changes caused by other metallic objects in the vehicle.

3.3 Coupling Coefficient and Discussion

The coupling coefficient is measured with the scanning pattern along the whole charging area, as shown in Fig. 3-5. For the sidewall charging, the average coupling coefficient of 0.075 is comparable to the coefficient of 0.07-0.09 in [48], which achieved
a coil-to-coil efficiency of 79\% using the same RX coil. In Fig. 3-5, it could also be observed that the coupling coefficient at the center of the charger decreases when the angle of elevation increases. Meanwhile, it also shows that the coupling coefficient is sufficient to support larger devices like mobile phones when the angle is less than 40°. Notably, the worst coupling coefficient occurs when the elevation angle \( \theta \) equals to 90°, which is standing vertically at the center of the 3-D charger. Fortunately,
the worst case scenario is unlikely to happen while the vehicle is moving. Therefore, it could be excluded from design consideration.

3.4 Summary

An in-vehicle WPT system that supports simultaneous charging for portable devices is proposed and studied. A 3-D structure provides a uniform 360° charging surface, enabling free positioning of the wearable devices. Sufficient magnetic flux perpendicular to each charging surface improves the tolerance to the random movements caused by the acceleration and deceleration of the vehicle.
CHAPTER 4
3-D WIRELESS CHARGING SYSTEM WITH FLEXIBLE RECEIVER COIL ALIGNMENT

4.1 Introduction

In previous chapters, the structures are designed specifically for a cell phone and a smart watch. However, in addition to these two devices, there are other consumer electronics with various shapes and sizes in our daily life. These devices can be as large as a tablet or a virtual real headset, or as small as a hearing aid. Therefore, a 3-D universal wireless charging system is proposed in this chapter. By applying the techniques discussed before, blind spots can be eliminated in the clinical shaped container. Besides, a new estimation method for multi-coils system is also discussed.

4.2 System Design

4.2.1 Coil Configuration

To generate a sufficient and uniform magnetic field density inside a volume, the shape and the arrangement of the transmitter (TX) coil needs to be considered. In a conventional two-coil system, the magnitude of the electromagnetic field will drop rapidly when the charging distance is increased, as described by Ampere’s law. Though the resonant coupling within a bowl-shaped wireless charger is improved in [46], there is still a dead zone near the center of the system when the receiving coil is placed in the perpendicular arrangement.

To overcome the above limitations, a structure with multiple transmitting coils aligned equally along the cylindrical wall is proposed, as shown in Fig. 4-1. The number of the transmitting coils, n, is a tradeoff between the size of the charger, the number and the type of receiving devices. According to the rule of the thumb of the coil size ratio in
[53], the dimension of each transmitting coil is designed to be around 1.5-2 times larger than the wearable receiving coil to improve the leakage of the magnetic flux. A piece of flexible ferrite sheet is used to surround the charger to channel the magnetic flux back to the TX coil. It can also reduce the electromagnetic interference to other electronic systems in the environment. By considering a smartphone and a wearable device as receiving devices, the size of this cylinder is designed to have 10 cm height and 12 cm diameter, while the thickness of the sidewall is 4 mm. Each TX coil cell is fabricated by winding 16-gauge wire with 4 mm spacing. The measured inductance value and quality factor of each cell is 1.08 \( \mu \text{H} \) and 270, respectively.

![Diagram](image)

Figure 4-1. Illustration of 3-D coil-array WPT system (a) Side view (b) Top view
4.2.2 Coil Activation Method

The magnetic field generated by each TX coil can be described by its phase and amplitude as a phasor expression. To create a charging volume with arbitrary spatial freedom of receiver placement, we need to generate not only omni-directional magnetic vectors, but also enhance the amplitude of the magnetic field to have sufficient coupling at center.

Several studies have been conducted to improve power transfer efficiency, such as using phase control to focus the magnetic field onto the receiver [54], applying the convex optimization theory to maximize the received power on both a single coil and multiple coils array systems [55-56]. However, both methods are restricted to 2-D transmitting coil arrays, which can only offer limited freedom in terms of height. Therefore, we expanded the degree of freedom by applying a 3-D coil array amplitude and phase control. As a result, the resulting magnetic field vectors have a rotational effect that enable the receiving coil to receive the power in arbitrary alignment. A six-coil configuration was selected as an example to demonstrate our strategy in this chapter, and the simulation results with different activation patterns depending on the placement of receiver(s) will be discussed.

To simplify the control methodology, we selectively activate TX coils with either 0° or 180° phase to couple the magnetic flux into the RX coil, as shown in Fig. 4-2. In Fig. 4-2 (a), the RX coil is aligned vertically and close to a TX coil (coil 1 in this illustration). In this example, only coil 2, 1, and 6 are activated with the input phase equal to 180°, 0°, 180°, respectively. Coil 1 forces magnetic flux into the center of the RX coil, while TX coils 2 and 6 pull flux out of the sides of the RX coil. Therefore, the opposite input phases of adjacent TX coils establish the forward path and the return
path of the flux, which generates a strong magnetic field distribution between the cylinder sidewall and the RX coil.

However, when the RX coil is placed right at the center of the charger, as shown in Fig. 4-2 (b), coils 1, 2, and 6 are activated with same input phase instead. From Fig. 4-2 (b), it can be observed that the normal vectors of the H field are enhanced towards the RX coil at center position with the enhanced flux density. Moreover, the ferrite sheets attached to the back of the RX coil as well as the ferrite surrounding the charger properly guide the flux back to TX coils. This flux path provides sufficient coupling when the radial distance $\rho$ is increased.

Fig. 4-2 (c) shows the other possible orientation when the RX coil is facing the gap between two adjacent coils. In this example, only coils 1 and 6 are activated with the same input phase to support sufficient coupling. The coil-to-coil efficiency can be improved by minimizing the number of activated TX coils.

Fig. 4-2 (d) illustrates another placement when the RX coil is lying at the bottom of the structure. By adjusting the number of activated TX coils with the same input phase, the RX coil can still receive sufficient coupling. In our demonstration, no planar TX coil is needed at the bottom, which further reduces the complexity of the system control unit.

In summary, the TX coils are selectively activated with 0° or 180° phase depending on the position and orientation of RX coil(s). Constraining the phase to 0° or 180° and constraining the activated coil current to have equal amplitude keeps the charger driving circuit simple. A single amplifier can be used, with a switching network to selectively activate the desired coils.
In summary, the TX coils are selectively activated with 0° or 180° phase depending on the position and orientation of RX coil(s). Constraining the phase to 0° or 180° and constraining the activated coil current to have equal amplitude keeps the

Figure 4-2. Magnetic field is simulated by the proposed cylindrical wireless charging system. Besides, the high permeability material, for instance, a ferrite sheet surrounding the structure is also simulated. Different TX coils are activated when RX coil is at (a) \( \rho=3 \text{ cm}, \phi=0^\circ, z=5 \text{ cm} \). (b) \( \rho=0 \text{ cm}, \phi=0^\circ, z=5 \text{ cm} \). (c) \( \rho=0 \text{ cm}, \phi=30^\circ, z=5 \text{ cm} \). (d) \( \rho=0 \text{ cm}, \phi=0^\circ, z=0 \text{ cm} \).
charger driving circuit simple. A single amplifier can be used, with a switching network
to selectively activate the desired coils.

4.3 Impedance Matrix and Coupling Coefficient $k'$ Estimation

The original definition of the coupling coefficient is derived from the two-port
network, as shown in Fig. 4-3.

Reflected impedance, $Z_{\text{ref}}$, is an equivalent resistance model connected in series
at TX side. The accordingly mutual inductance and reflected impedance are derived as
follows,

$$M_{1,2} = k \sqrt{L_1 \times L_2} \quad (4-1)$$

$$Z_{\text{ref}} = \frac{(\omega M_{1,2})^2}{Z_{\text{oc}}} \quad (4-2)$$

Where $Z_{\text{oc}}$ is the equivalent impedance at the receiver side when the circuit is open.

From the equation (4-2), $Z_{\text{ref}}$ is related to mutual inductance between the pair of coils,
which is also affected by the distance and the angle alignment between coils. That is
because the mutual inductance can be derived as

$$M_{1,2} = \frac{\pi \mu_0 N_1 N_2 a_1^2 b_2^2 |\cos \theta_{12}|}{2 (a_1^2 + d_{12}^2)^{2/3}} \quad (4-3)$$

Where $d_{1,2}$ is the distance between two coils, $\theta_{1,2}$ are the angle between axes of two
coils, $N_1$, $N_2$ are the turns of TX coil and RX coil, $a_1$ and $b_2$ are the radius of TX and RX
coil, respectively.

Therefore, for a single TX and single RX WPT system, the coupling coefficient
estimation is relatively accurate based on above equations. However, when the system
composed of more coils, the derivation of the coupling coefficient needs to be
reconsidered. Among all the analysis [57-62], impedance matrix is a well-adopted method to discuss multi-input and multi-output network.

Assume the system containing m TX coils and n RX coils, then the impedance matrix can be presented as

\[
\begin{bmatrix}
R_1 & j\omega M_{1,2} & \cdots & j\omega M_{1,n-1} & j\omega M_{1,n} \\
 j\omega M_{2,1} & R_2 & \cdots & j\omega M_{2,n-1} & j\omega M_{2,n} \\
\vdots & \vdots & \ddots & \vdots & \vdots \\
 j\omega M_{i,1} & j\omega M_{i,2} & \cdots & j\omega M_{i,n-1} & j\omega M_{i,n} \\
\vdots & \vdots & \ddots & \vdots & \vdots \\
 j\omega M_{n-1,1} & j\omega M_{n-1,2} & \cdots & R_{n-1} & j\omega M_{n-1,n} \\
 j\omega M_{n,1} & j\omega M_{n,2} & \cdots & j\omega M_{n,n-1} & R_n
\end{bmatrix}
\begin{bmatrix}
|I|e^{j\theta_1} \\
|I|e^{j\theta_2} \\
\vdots \\
|I|e^{j\theta_i} \\
|I|e^{j\theta_{n-1}} \\
|I|e^{j\theta_n}
\end{bmatrix}
= 
\begin{bmatrix}
V_1 \\
V_2 \\
\vdots \\
V_i \\
V_{n-1} \\
V_n
\end{bmatrix}
\]

(4-4)

Where the definition of each parameter is listed in Table 4-1.
From Eq. (4-4), the input power generated from the transmitting coil arrays can be obtained as

$$P_{TX} = \sum_i^{m} P_{TX,i} = \sum_i^{m} \sum_i R\{V_i \times |I_j| e^{j\theta_i})\} \quad (4-5)$$

While the output power delivered to the RX coils are

$$P_{RX} = \sum_j^{n} P_{RX,j} = \sum_j^{n} R\{I_j^2 \times R_{Lj})\} \quad (4-6)$$

<table>
<thead>
<tr>
<th>Table 4-1. Parameters Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k_{ij}$</td>
</tr>
<tr>
<td>$\theta_{ij}$</td>
</tr>
<tr>
<td>$M_{ij}$</td>
</tr>
<tr>
<td>$a_i$</td>
</tr>
<tr>
<td>$b_j$</td>
</tr>
<tr>
<td>$d_{ij}$</td>
</tr>
<tr>
<td>$N_i, N_j$</td>
</tr>
<tr>
<td>$\mu_0$</td>
</tr>
<tr>
<td>$r_i$</td>
</tr>
<tr>
<td>$\phi_i$</td>
</tr>
<tr>
<td>$V_i$</td>
</tr>
<tr>
<td>$R_{Lj}$</td>
</tr>
<tr>
<td>$\omega$</td>
</tr>
</tbody>
</table>
Eq. (4-5) (4-6) show the traditional way to evaluate the performance of a multi-ports WPT system. The whole system needs to be fully implemented including the matching networks before calculating the input and output power. Unlike the coupling coefficient for a two port network, there is no such method for a multi-input and multi-out WPT system to get an initial estimation. Therefore, a new parameter $k'$ and the equivalent circuit model are proposed to predict the EM field distribution and transformer performance before designing the whole system.

To simplify the analysis of the system, two cases are discussed separately in the following section.

### 4.3.1 Multi-TX Coil vs. Single RX Coil

Fig. 4-4 shows an illustration of the multi-TX coils and a single RX coil system. Three TX coils (Coil #1, Coil #2, and Coil #3) vs. one RX coil (Coil #4) are used as an example in the following.

Assume each TX coil is identical and the currents flowing through TX coils are the same,

$$|I_{TX1}| = |I_{TX2}| = |I_{TX3}| = |I_{TX}|$$  \hfill (4-7)

$$L_{TX1} = L_{TX3} = L_{TX3} = L_{TX}$$  \hfill (4-8)

$$r_1 = r_2 = r_3 = r$$  \hfill (4-9)
Hence, we will have the same matching network and resonant frequency as

\[ C_{s1} = C_{s2} = C_{s3} = C_s \]  
\[ \omega_0 = \frac{1}{\sqrt{L_{TX1} C_{s1}}} = \frac{1}{\sqrt{L_{TX2} C_{s2}}} = \frac{1}{\sqrt{L_{TX3} C_{s3}}} \]  

Mutual inductance between any two TX and RX coil can also be re-written as

\[ M_{ij} = \frac{\pi \mu_0 N_i N_j a_i^2 b_j^2 |\cos \theta_{ij}|}{2(a_i^2 + d_{ij}^2)^{2/3}} \]

From Eq. (4-12), it can be observed that the transmission distance and the angle misalignment are already considered in the mutual inductance. As a result, the reflected impedance of each TX coil can be derived from Fig. 4-4,
Where the condition $1 \gg j\omega R_i C_{2b} j$ is assumed.

From Fig. 4-4, equivalent input impedance seen by each TX coil #1 can be expressed as

$$Z_1 = \frac{1}{j\omega C_{s1}} + j\omega L_{TX1} + Z_{REF1} + r_1$$  \hspace{1cm} (4-14)$$

Insert Eq. (4-13) into Eq. (4-14), $Z_1$ can be re-written as

$$Z_1 = \frac{1}{j\omega C_{s1}} + j\omega L_{TX1} + r_1 + \frac{\omega^2 k_{14}^2 L_{TX1} L_{RX4}}{j\omega L_{RX4} + j\omega L_{RX4} + r_4 + R_L} = r_1 + \frac{\omega^2 k_{14}^2 L_{TX1} L_{RX4}}{r_4 + R_L}$$  \hspace{1cm} (4-15)$$

Where $\omega = \omega_0$, $j\omega_0 L_{TX1} + \frac{1}{j\omega_0 C_{s1}} = 0$, $j\omega_0 L_{RX} + \frac{1}{j\omega_0 C_{2a}} = 0$

Input impedance of coil #2 and coil #4 can also be calculated as

$$Z_2 = r_2 + \frac{\omega^2 k_{24}^2 L_{TX2} L_{RX4}}{r_4 + R_L}$$  \hspace{1cm} (4-16)$$

$$Z_3 = r_3 + \frac{\omega^2 k_{34}^2 L_{TX3} L_{RX4}}{r_4 + R_L}$$  \hspace{1cm} (4-17)$$

Therefore, the power transfer efficiency can be calculated as

$$\frac{PTE}{\eta_{PA}} = \eta_{coil-to-coil} \times \eta_{RX}$$

$$= \frac{|I_{TX1}|^2 \times (\cos \phi_1 + j \sin \phi_1) \times Z_{REF1} + \cdots + |I_{TX1}|^2 \times (\cos \phi_3 + j \sin \phi_3) \times Z_{REF3}}{|I_{TX1}|^2 \times (\cos \phi_1 + j \sin \phi_1) \times (Z_{REF1} + r_1) + \cdots + |I_{TX1}|^2 \times (\cos \phi_3 + j \sin \phi_3) \times (Z_{REF3} + r_3)}$$

$$\times \frac{R_L}{(r_4 + R_L)}$$
\[
= \frac{(\cos \phi_1 + jsin \phi_1) \times Z_{REF1} + \cdots + (\cos \phi_3 + jsin \phi_3) \times Z_{REF3}}{(\cos \phi_1 + jsin \phi_1) \times (Z_{REF1} + r_1) + \cdots + (\cos \phi_3 + jsin \phi_3) \times (Z_{REF3} + r_3) \times \frac{R_{L4}}{(r_4 + R_{L4})}}
\]
\[
= \frac{(\cos \phi_1 + jsin \phi_1) \times \left(\frac{\omega_0^2 k_{14}^2 L_{TX1} L_{RX4}}{r_4 + R_{L4}} + \cdots + \frac{\omega_0^2 k_{34}^2 L_{TX3} L_{RX4}}{r_4 + R_{L4}}\right)}{(\cos \phi_1 + jsin \phi_1) \times \left(\frac{\omega_0^2 k_{14}^2 L_{TX1} L_{RX4}}{r_4 + R_{L4}} + \cdots + \frac{\omega_0^2 k_{34}^2 L_{TX3} L_{RX4}}{r_4 + R_{L4}}\right) + (\cos \phi_3 + jsin \phi_3) \times \left(\frac{\omega_0^2 k_{13}^2 L_{TX1} L_{RX4}}{r_4 + R_{L4}} + \cdots + \frac{\omega_0^2 k_{33}^2 L_{TX3} L_{RX4}}{r_4 + R_{L4}}\right)}
\]
\[
\times \frac{R_{L4}}{(r_4 + R_{L4})}
\]
\[
= \frac{\omega_0^2 L_{TX} L_{RX4} k r^2}{r_4 + R_{L4}} \times \frac{R_{L4}}{r_4 + R_{L4}}
\]
\[
= \frac{\omega_0^2 L_{TX} L_{RX4} k r^2}{r_4 + R_{L4}} \times \frac{R_{L4}}{r_4 + R_{L4}}
\]
\[
(4-18)
\]

Where \( k' \) is defined as
\[
k' = \sqrt{\frac{k_{14}^2 (\cos \phi_1 + jsin \phi_1) + k_{24}^2 (\cos \phi_2 + jsin \phi_2) + k_{34}^2 (\cos \phi_3 + jsin \phi_3)}{r_4 + R_{L4}}}
\]
\[
(4-19)
\]

From Eq. (4-18) and (4-19), the equivalent circuit model in Fig. 4-4 can be re-draw it as shown in Fig. 4-5. The new equivalent reflected impedance \( Z_{ref'} \), new equivalent parasitic resistances \( r' \) are defined as
\[
Z_{ref} = (\cos \phi_1 + jsin \phi_1) \times Z_{REF1} + (\cos \phi_2 + jsin \phi_2) \times Z_{REF2} + (\cos \phi_3 + jsin \phi_3) \times Z_{REF3}
\]
\[
(4-20)
\]
\[
r' = r_1 (\cos \phi_1 + jsin \phi_1) + r_2 (\cos \phi_2 + jsin \phi_2) + r_3 (\cos \phi_3 + jsin \phi_3)
\]
\[
(4-21)
\]

From the above equations, it can be learned that \( Z_{ref'} \) and \( r' \) and \( k' \) are all related to the input phase of the AC signal. However, since each matching network connects to the transmitting coil independently, the self-inductance of the TX coil
in Fig. 4-5 is actually equal to the original value $L_{TX}$. In this case, not only the coupling coefficient $k'$ can be adjusted with different phase combination, the new equivalent quality factor $Q'$ and loaded quality factor $Q_{\text{loaded}}'$ can be manipulated as well, as shown in Eq. (4-22) and (4-23)

$$Q' = \frac{L_{TX}}{r_1}(\cos \phi_1 + j \sin \phi_1) + r_2(\cos \phi_2 + j \sin \phi_2) + r_3(\cos \phi_3 + j \sin \phi_3)$$ (4-22)

$$Q_{\text{loaded}} = \frac{L_{TX}}{Z_{\text{ref}}'} = \frac{L_{TX}}{(\cos \phi_1 + j \sin \phi_1) \times Z_{\text{REF}_1} + (\cos \phi_2 + j \sin \phi_2) \times Z_{\text{REF}_2} + (\cos \phi_3 + j \sin \phi_3) \times Z_{\text{REF}_3}}$$ (4-23)

### 4.3.2 Multi-TX Coil vs. Multi-RX Coil

To expand the circuit model from a single RX coil to Multi-RX coils, as shown in Fig. 4-6, new equivalent parameters are derived as follows.

Assume $m$ TX coils are activated and $n$ RX coils are placed in the charger,

$$Z_{\text{ref}_1'} = (\cos \phi_1 + j \sin \phi_1) \times Z_{\text{REF}_1,RX_1} + \cdots + (\cos \phi_m + j \sin \phi_m) \times Z_{\text{REF}_m,RX_1}$$ (4-24)
\[ Z_{refm}' = (\cos \phi_1 + j\sin \phi_1) \times Z_{REFm_{RX1}} + \cdots + (\cos \phi_m + j\sin \phi_m) \times Z_{REFm_{RXn}} \quad (4-25) \]

\[ r' = r_1(\cos \phi_1 + j\sin \phi_1) + \cdots + r_m(\cos \phi_m + j\sin \phi_m) \quad (4-26) \]

\[ k_1' = \sqrt{|k_{11}^2(\cos \phi_1 + j\sin \phi_1) + \cdots + k_{m1}^2(\cos \phi_m + j\sin \phi_m)|} \quad (4-27) \]

\[ \vdots \]

\[ k_n' = \sqrt{|k_{1n}^2(\cos \phi_1 + j\sin \phi_1) + \cdots + k_{mn}^2(\cos \phi_m + j\sin \phi_m)|} \quad (4-28) \]

From the Eq. (4-26) and Fig. 4-6, it can be viewed that \( r' \) is independent from receiving side, which reduces the impedance variation range connected to the transmitting coil. Hence, the charging system can be kept stable when supporting multiple RX devices simultaneously.

### 4.3.3 Measured Coupling Coefficient \( k' \)

Since the activation method in this proposed system is relatively simple and the input phases of the AC signal are limited to only 0 and 180 degree, Eq. (4-25), (4-26), and (4-28) can be further simplified as Eq. (4-29), (4-30), and (4-31) respectively.

\[ Z_{refm}' = Z_{REFm_{RX1}} \pm Z_{REFm_{RX2}} \cdots + Z_{REFm_{RXn}} \quad (4-29) \]

\[ r' = r_1 \pm r_2 \cdots \pm r_m \quad (4-30) \]

\[ k_n' = \sqrt{|k_{1n}^2 \pm k_{2n}^2 \cdots \pm k_{mn}^2|} \quad (4-31) \]

Therefore, the following figures of \( k' \) vs. position are calculated and plotted from measured results using Eq. (4-31). Besides, the TX coils placed on the sidewall of the cylinder are all symmetrical, only three TX coils are activated and discussed here.
Case 1) When three TX coils are all activated in-phase, 1 wearable coil is placed randomly inside the charger, as shown in Fig. 4-7 (a):

$$k' = \sqrt{k_{11}^2 + k_{22}^2 + k_{31}^2}$$  \hspace{1cm} (4-32)

The distribution of $k'$ is consistent with the magnetic field simulated in Ansoft HFSS, as shown in Fig. 4-7 (b). It can be observed that when the RX coil is away from the sidewall, both $k'$ and H filed magnitude is decreased. The scanning area is also shown in Fig. 4-7 (b).
Figure 4-7. Results analysis when all TX coils are in-phase. (a) The measured coupling coefficient versus different position of a wearable coil. (b) The simulated H field distribution versus different position of a wearable coil.
(Case 2) When three TX coils are activated with coil #1 and #2 are in-phase and coil #3 is out-phase, 1 wearable coil is placed randomly inside the charger, as shown in Fig. 4-8.

\[ k' = \sqrt{k_{11}^2 + k_{22}^2 - k_{31}^2} \]

(4-33)

In this case, \( k' \) is smaller than the previous case, but the variation range of \( k' \) is also reduced. This input phase combination can help ease the frequency splitting phenomenon which will be discussed later.

![Figure 4-8. The measured coupling coefficient versus different position of a wearable coil while all TX coils are two in-phase, one out of phase.](image)

Case 3) When three TX coils are all activated in-phase, 1 phone coil is placed randomly inside the charger, as shown in Fig. 4-9:

\( k' \) can be calculated using the same equation in case 1, and the results show that \( k' \) is higher than 0.05 within the entire area. It can also be noticed that the distribution of \( k' \) here is smoother that it is in Fig. 4-8 and 4-7. This is because the size of the phone coil is much larger than the wearable coil, so that \( k' \) and the magnetic flux density are less
Case 4) When three TX coils are all activated in-phase, 1 phone coil placed in the center (worst case scenario) and 1 wearable coil scanned over the rest of charging area, as shown in Fig. 4-10. Comparing to the Fig. 4-7 with only one wearable coil charged inside, $k'$ is higher due to the mutual inductance of the phone coil. From Fig. 4-7 to Fig. 4-10, the distribution of $k'$ shows a good agreement with H field distribution simulated using Ansoft HFSS software.

Figure 4-9. The measured coupling coefficient versus different position of a phone coil while all TX coils are all in-phase.
4.4 Measurement Results and Discussion

4.4.1 Measurement Setup and EM Scanning System

The wearable coil used for testing the performance of this system is similar to the coils in [48], which has an area of 29x31 mm$^2$ in a square shape. However, for a smaller receiving device, the variation of mutual inductance caused by different positioning is too subtle to get accurate results. Furthermore, the cable connected to both coils interfere the measurement while moving the receiving device in six degrees of direction. Therefore, the equation in (4-1) and (4-2) are re-derived as follows [63],

$$k^2 = \frac{L_2(L_1 - L_s)}{L_1 \times L_2} = \frac{(L_1 - L_s)}{L_1} = 1 - \frac{L_s}{L_1} \quad (4-34)$$

where $L_s$ equals to the same measurement while $L_2$ is short-connected. By using (4-34), the measurement can be performed in three steps:

Figure 4-10. The measured coupling coefficient versus different position of a phone coil and a wearable coil while all TX coils are all in-phase.
1. Measured the impedance of TX coil by shorting the receiving coil first and scanning it within the whole charging volume.

2. Reset the motor to the origin.

3. Repeated the same scanning pattern in step 1 with opening the two terminals of the receiving coil.

By replacing the data collected from step 1 and 3 into (4-34), the contour figure of coupling coefficients versus positions can be plotted with high resolution.

On the other hand, since it is a 3-D charging volume with six-degrees of freedom, the existing computer numeric control (CNC) stepping machines with only x, y, z translation freedom cannot detect electro-magnetic (EM) field variation accurately. Though traditional electromagnetic interference (EMI) and electromagnetic compatibility (EMC) scanners can achieve higher resolution for 3-D structures by using laser beams, its high cost and metallic parts are not applied to passive EM field measurement and sensitive power transfer efficiency. Therefore, an intelligent EM field scanning system is developed to measure the coupling coefficient and the power transfer efficiency for x,y,z translation and rotation in the cylindrical volume, respectively. This scanning system includes a non-metallic scanner and a LabVIEW system for controlling the movement and the rotation, as shown in Fig. 4-11. According to (4-34), the impedance of TX coil are measured with shorting and opening the receiving coil. After finishing the EM scanning, both measurement results are fed back into the computer to calculate and plot the results, as shown in Fig. 4-12, 4-13, and 4-14. Notably, with the re-position accuracy of less than 0.1 mm, the possibility of mismatch of the open-loop and the closed-loop impedance measurement is negligible.
Figure 4-11. The measurement setup with a 6 degrees of freedom EM scanning system. Photo courtesy of author.
4.4.2 Measured Coupling Coefficient and Power Transfer Efficiency

To demonstrate the spatial freedom of positioning, different orientations and positions are measured. Fig. 4-12 shows the coupling coefficient versus different rotation of the wearable coil. Since the TX coils are arranged symmetrically on the sidewall, only 60° of rotation is presented here. By activating each TX coil independently, an average coupling coefficient larger than 0.03 is achieved for the worst case scenario. Fig. 4-13 shows the corresponding coupling coefficient versus tilt variation when RX coil is at the center of the structure. A coupling coefficient of better than 0.05 is achieved at θ = 0° (no tilt, worst case scenario). Moreover, the coupling coefficient can be seen to vary with the tilt angle.

![Figure 4-12](image.png)

Figure 4-12. The measured coupling coefficient versus different orientation and position of a wearable coil.
coefficient of the smartphone coil was measured to be higher than 0.09 at the center position (worst case scenario). The overall coupling coefficient is increased 0.01 to 0.03 depending on the RX position after adding the flexible ferrite sheet outside the structure.

The power transfer efficiency (PTE) of the system including power amplifier, rectifiers, and coil losses is 39% for supporting a cell phone and a wearable device simultaneously. Fig. 4-14 shows the measured PTE for a single wearable device versus different alignments. The average PTE higher than 8.8% is achieved for the worst case scenario, corresponding to the center position of the cylinder. The maximum coil to coil efficiency for a wearable device is calculated to be 85%. A performance comparison with prior works is listed in Table 4-1.

Figure 4-13. The measured coupling coefficient versus different tilt of a wearable coil at the center of the structure.
4.4.3 Coupling Coefficient Uncertainties Analysis

Propagation of uncertainty is a common way to evaluate an experiment accuracy statistically [64]. In Eq. (4-34), $k$ is defined by two parameters, $L_1$ and $L_S$, which are two independent value sets collected separately. Therefore, fractional uncertainty of $k$ can be expressed as
\[
\frac{\delta k}{|k|} = \sqrt{\left(\frac{\delta L_s}{L_s}\right)^2 + \left(\frac{\delta L_1}{L_1}\right)^2}
\] (4-35)

where \(|k|\) is the best estimate of \(k\), \(\delta L_s\) and \(\delta L_1\) are uncertainties of \(L_s\) and \(L_1\), respectively. According to propagation of uncertainty theory [64], the uncertainties \(\delta L_s\) and \(\delta L_1\) both have the random error and the systematic error, which are decided by the repeatability of measurements and the precision of instruments, individually. Therefore, \(\delta L_s\) and \(\delta L_1\) can be further unfolded into two terms,

\[
\delta L_s = \sqrt{\delta_{LSR}^2 + \delta_{LSS(VNA)}^2}
\] (4-36)

\[
\delta L_s = \sqrt{\delta_{L1R}^2 + \delta_{L1S(VNA)}^2}
\] (4-37)

where \(\delta_{LSR}\) and \(\delta_{L1R}\) are random errors of \(\delta L_s\) and \(\delta L_1\), respectively, while \(\delta_{LSS(VNA)}\) and \(\delta_{L1S(VNA)}\) are systematic errors of \(\delta L_s\) and \(\delta L_1\), respectively. Notably, since both coupling coefficient \(k\) and equivalent coefficient \(k'\) are all calculated using the measured data collected from a vector network analyzer (VNA), the following discussion will be focused on the errors contributed by Keysight (VNA 5061B) and the calibration kit (85033E).

For the middle \(Z\)-range (10 Ω to 100 Ω) defined by Keysight (former Agilent Technologies), the uncertainty of impedance measurement is less than 5% [65]. According to the table in [65], the uncertainty of input impedance \(Z_{11}\) at 6.78 MHz is around 2.5%, which is under an assumption that some reproducible errors has been corrected by using the suitable calibration kit. In this dissertation, all experiments are calibrated and compensated by using 85033E kit for subminiature version A (SMA) connectors. Hence, by multiplying the mean value of \(L_s\) and \(L_1\) with the error percentage
2.5%, $\delta_{LSS(VNA)}$ and $\delta_{L1S(VNA)}$ can be calculated as 1.154 Ω and 1.142 Ω for scanning the wearable coil.

On the other hand, random error is usually caused by unwanted and unpredictable changes in the experiments, such as noise, heat, or unavoidable EM interference from the measurement environment [66]. To take all effects into consideration, $\delta_{LSR}$ and $\delta_{L1R}$ can be presented as

$$\delta_{LSR} = \frac{\sigma_{Ls}}{\sqrt{N}}$$ (4-38)

$$\delta_{L1R} = \frac{\sigma_{L1}}{\sqrt{N}}$$ (4-39)

where $\sigma_{Ls}$ and $\sigma_{L1}$ are the standard deviation of $L_s$ and $L_1$, respectively, and N is the number of measurement samples. From the experimental data measured for wearable coil, $\delta_{LSR}$ and $\delta_{L1R}$ are 0.042 and 0.029, respectively given N=3.

As a result, $\delta k$ can be calculated by replacing the value computed from Eq. (4-35) to (4-39),

$$\delta k = |k| \times \sqrt{\left(\frac{\delta L_s}{L_s}\right)^2 + \left(\frac{\delta L_1}{L_1}\right)^2} = 0.08 \times \sqrt{1.062^2 + 1.057^2} = 0.12$$ (4-40)

4.5 Summary

A 3-D wireless charging structure with the spatial freedom of receiver coil is presented in this chapter. By selectively activating TX coil with either 0° or 180° input phase, the rotating magnetic field is generated and guided towards the RX coil. The symmetrical structure provides sufficient coupling regardless of the placement and
alignment of RX coil, and therefore eliminates the dead-zone within the charger.

Moreover, a new parameter k’ and a new estimation method are proposed and verified.

Table 4-2. Comparison of WPT Systems with PDL < 10 W and Increased Spatial Freedom

<table>
<thead>
<tr>
<th></th>
<th>[49]</th>
<th>[46]</th>
<th>[50]</th>
<th>[51]</th>
<th>[This work]</th>
</tr>
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<tr>
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<td>Implantable  sensor</td>
<td>Hearing aid</td>
<td>Test structures, open air</td>
<td>Mobile phone</td>
<td>Smart watch + Mobile phone</td>
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<td>28 mW</td>
<td>275 mW</td>
<td>3 W</td>
<td>1W + 5W</td>
</tr>
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<td><strong>Max PTE</strong></td>
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<td>28%</td>
<td>37%</td>
<td>60%</td>
<td>39%</td>
</tr>
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<td><strong>PTE terminals</strong></td>
<td>DC-to-DC</td>
<td>DC-to-DC</td>
<td>coil-to-coil</td>
<td>DC-to-DC</td>
<td>DC-to-DC</td>
</tr>
<tr>
<td><strong>Spatial Freedom</strong></td>
<td>20 cm distance, 30 deg tilt</td>
<td>360 deg inside bowl</td>
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<td>Planar charging surface, 32x22 cm²</td>
<td>xyz translation xyz rotation</td>
</tr>
</tbody>
</table>
CHAPTER 5
LOAD-INDEPENDENT CLASS-E POWER AMPLIFIER FOR COIL-ARRAYS WPT SYSTEM

5.1 Introduction

The most general requirement to design a WPT transmitter is to maintain a stable system without sacrificing the ability of charging multiple devices simultaneously at useful charging rate for each device. This concept also reflects the characteristic of the battery charging profile that is typically begins with a constant current mode and then, when the cell charge level reaches a certain value, changing to the constant voltage mode [67]. A common solution is to implement a single passive matching network to generate either constant current or constant voltage flowing through the transmitting (TX) coil independently from a time-varying load [68]. Depends on different topologies and applications, Class D or Class E switching power amplifiers combining with LC matching network are well-discussed [68-70]. However, this limits the design possibilities of wireless transmitter to certain structures, such as a single TX coil or a TX coil array that activates only one coil at a time [71]. While multiple load networks are used in [72], multiple power sources and power transistors are also needed. In this chapter, a load-independent Class E power amplifier is proposed for coil arrays WPT systems. Minimum number of power transistors is utilized in the structure but at the same time to maintain constant transmitting coil current for a transmitting coil array. The circuit is demonstrated with the 3-D coil array system presented in chapter 4, but the concept can be applied to a planar structure as well. As mentioned earlier, to reduce the design complexity, the signal fed into each transmitting coil is with same amplitude and 0° and 180° phase difference.
On the other hand, frequency splitting becomes an interest of topic recently especially when the coupling between two coils becomes stronger. Therefore, when the coupling coefficient exceed the critical value, the waveform of the power gain of the system started to separate two peaks and the received power decreases. The system falls away from the optimized value. The phenomenon is observed and analyzed in the proposed resonant structure, while the solution to suppress the frequency splitting is also discussed.

5.2 LCL-\(\pi\) Matching Network

Matching networks serve in a WPT system to fulfill several purposes, including to achieve maximum power transfer efficiency [73], to deliver nominal power to receiver [74], and/or to provide an active tuning mechanism to suppress frequency splitting [75]. In this proposed system, to supply a fixed current to the load is essential, especially when charging multiple devices simultaneously. Traditionally, in a WPT system, the transmitting current \(I_{TX}\) is variant with the reactance and resistance change due to metal detuning, high permeability materials, addition of receivers, and the time-increasing load, etc. Among above reasons, addition of receivers causes the largest impedance variation. Once \(I_{TX}\) starts changing, the power deliver to the receiver is not constant which will lower the system stability. Therefore, a new matching network is presented to generate a reasonably constant current over a wide range of component and load changes, as shown in Fig. 5-1.

Equivalent Impedance, current, and the components are defined as in Fig. 5-1. Since the main interest here is the controlling current \(I_{TX}\) through the TX coil, equations will be derived step by step in the following. To simplify the derivation, a single TX and
single RX case is discussed first. Hence, a single $Z_{ref}$ is connected in series, as illustrated in Fig. 5-1.

![Figure 5-1. The LCL-π matching network.](image)

Input current $I_{IN}$ can be defined as

$$I_{IN} = \frac{V_{IN}}{Z_{IN}} \quad (5-1)$$

Coil Impedance is given by

$$Z_{TX} = j\omega L_{TX} + \frac{1}{j\omega C_s} + Z_{ref} \quad (5-2)$$

Therefore, input Impedance can be calculated as

$$Z_{IN} = jL_{XM} + \frac{(-jX_{CM}) * Z_{TX}}{Z_{TX} - jX_{CM}} \quad (5-3)$$

Coil current is derived as

$$I_{TX} = I_{IN} * \frac{-jX_{CM}}{Z_{TX} - jX_{CM}} \quad (5-4)$$

Assume

$$X_{LM} = X_{CM} = \omega L_{TX} - \frac{1}{\omega C_s} = X' \quad (5-5)$$
Then TX impedance can be rewritten

\[ Z_{TX} = jX' + Z_{REF} \quad (5-6) \]

Input impedance is recomputed as

\[ Z_{IN} = jX' + \frac{(-jX')^*(jX' + Z_{REF})}{jX' + Z_{REF} - jX'} = \frac{X'I^2}{Z_{REF}} \quad (5-7) \]

From Eq. (5-1) and (5-7), Input current is rewritten as

\[ I_{IN} = \frac{V_{IN} \times Z_{REF}}{X'I^2} \quad (5-8) \]

Insert Eq. (5-3) and (5-8) into (5-4), current flowing into the TX coil can be presented as

\[ I_{TX} = I_{IN} \times \frac{-jX'}{jX' + Z_{REF} - jX'} = \frac{V_{IN} \times Z_{REF}}{X'I^2} \times \frac{-jX'}{Z_{REF}} = \frac{-jV_{IN}}{X'I} \quad (5-9) \]

From Eq. (5-10), it can be observed that TX coil does not depend on the reflected impedance \( Z_{ref} \).

5.3 Class-E Power Amplifier with LCL-\( \pi \) Matching Network

In the last section, a LCL-\( \pi \) matching network is already proved to be independent from \( Z_{ref} \) variation for a single TX coil system. However, a single TX coil structure is not always optimal for a wireless charging system. For instance, it might be advantageous to design a coil array system for charging receiving devices with different size scale. Moreover, a coil array system allowing multiple TX coils to be activated simultaneously can improve the power transfer efficiency and the freedom of positioning [76]. Therefore, two different implementations with the LCL-\( \pi \) matching network are simulated and compared.
Fig. 5-2 illustrates the traditional structure of class-E power amplifier by sharing a single matching network. $Z_{ref,1}, Z_{ref,2}, \ldots \ldots$ to $Z_{ref,n,m}$ represents the reflected impedance induced by receiving devices #1, #2…to receiving device #n, respectively. The 3-way switch allows the TX coil to be activated selectively. Three terminals of the switch are connected to $V +$, $V -$ and ground, respectively. Ideally, a turned on switch

![Diagram of a Class-E power amplifier with a single LCL-π matching network.](image-url)

**Figure 5-2.** A Class-E power amplifier is implemented with a single LCL-π matching network.
acts like a copper wire without loss. However, in the real implementation, a small parasitic resistor, $r_{on}$, exists in the circuit. To reduce the power consumption and avoid the heat generated by $r_{on}$, the power transistor for the switch must be carefully selected. Besides, given the total power delivered by the coil may be up to 25 Watt, the rating of the switch transistor needs to considered as well.

The simulation result with different $Z_{ref}$ is shown in Fig. 5-3. Six coils are activated in phase with $Z_{ref}=0.9\,\Omega, 3\,\Omega, 7\,\Omega, 10\,\Omega, 15\,\Omega, 20\,\Omega$, respectively.

Figure 5-3. Simulated TX coil currents when $V_{IN}=10\,\text{V}$, and $Z_{ref}=0.9\,\Omega, 3\,\Omega, 7\,\Omega, 10\,\Omega, 15\,\Omega, 20\,\Omega$, respectively.
The amplitude variation of $I_{TX}$ is as large as 500% comparing two extreme case scenario. In addition to the amplitude change, a slightly time delay of the current waveform is also introduce even if only resistance variation is simulated. Though it requires less components with this conventional structure, the unstable current behavior is not suitable for coil array system.

Hence, a load independent Class-E power amplifier with multiple load networks is proposed, as shown in Fig. 5-4. The switch in this proposed structure could be an N way switch, which depends on the phase combination of the input signal. In this example, a differential class E power amplifier is used to drive some or all of transmitting coils with 0° or 180° phase relationships, as illustrated in Fig. 5-4. Therefore, a 2-way switch is implemented here. Notably, in order to keep each matching network independent, the switch needs to be placed close to the TX coils.

As a comparison, the circuit is also simulated with $Z_{ref} = 0.9\Omega$, 3\Omega, 7\Omega, 10\Omega, 15\Omega, 20\Omega$, respectively, as depicted in Fig. 5-5. The current $I_{TX}$ variation is less than 5%, as plotted in Fig. 5-5 (b). Moreover, there is nearly no time delay with resistive variation.

Another simulation is performed with the reactance variation from $Z_{ref} = 0.5 \mu H$ to $Z_{ref} = 2 \mu H$, as shown in Fig. 5-6. According to Fig. 5-6 (b), the current $I_{TX}$ variation is less than 8%. Thus, with a stable current behavior over a wide range of resistance and reactance change, the proposed circuit can be applied to the coil array WPT system. By properly adjusting capacitors, $C_{11}$, $C_{12}$..., $C_{2n}$, the zero-voltage switching condition is achieved, as shown in Fig. 5-7.
Figure 5-4. A load independent Class-E power amplifier is implemented with multiple LCL-\(\pi\) matching network.
Figure 5-5. Simulated TX coil currents when VIN=10 V, and $Z_{ref} = 0.9 \, \Omega, 3 \, \Omega, 7 \, \Omega, 10 \, \Omega, 15 \, \Omega, 20 \, \Omega$, respectively. (a) current waveform from 0.93 $\mu$s to 1.15 $\mu$s. (b) The waveform is zoomed-in from 0.99 $\mu$s to 1.0 $\mu$s.
Figure 5-6. Simulated TX coil currents when $V_{IN}=10$ V, and $Z_{ref}$ varies from 0.5 $\mu$H to 2 $\mu$H. (a) current waveform from 0.93 $\mu$s to 1.15 $\mu$s. (b) The waveform is zoomed-in from 0.99 $\mu$s to 1.0 $\mu$s.
Reflected Impedance Variation

In the previous chapter, the reflected impedance is discussed under an assumption when the components are conjugate matched. Nevertheless, under non-ideal matching scenario, where the reactance of $L_{RX}$ is not perfectly cancelled by the matching series capacitor $C_{2a}$, Eq. (4-13) will be re-written as

$$Z_{ref} = \frac{\omega_0^2 k_{TX,RX}^2 L_{TX} L_{RX}}{j\omega_0 L_{RX} + \frac{1}{j\omega_0 C_{2a}} + R_{RX} + \frac{R_L}{j\omega_0 R_L C_{2b} + 1}}$$  \hspace{1cm} (5-10)
When the inductance value of TX coil is relatively smaller than RX coil, remanence of the reactance due to non-ideal mismatch can no longer be ignored. Given that \( Z_{ref} \) is proportional to \( k_{TX,RX}^2 \), even small variations of the coupling coefficient can result in large reactive values of \( Z_{ref} \), which is shown graphically in Fig. 5-8.

Fig. 5-8 illustrate how \( Z_{ref} \) varies with the coupling coefficient and frequency for a wearable device and/or a cell phone, where \( k_{TX,RX} \) is plotted from worst case scenario \( (k_{TX,RX} \approx 0) \) to best case scenario \( (k_{TX,RX} = 1) \). It can be observed that when \( k_{TX,RX} \) is larger than \( 10^{-1} \), reactance of \( Z_{ref} \) becomes either inductive or capacitive. When \( k_{TX,RX} \) rises into the strong-coupled region, which usually defined as the coupling higher than

![Figure 5-8. Reflected impedance variation versus frequency and coupling coefficient.](image-url)
0.15 [77], the variation of $Z_{ref}$ changes dramatically, especially when charging two devices at the same time. Notably, a standard value 50 Ω is used here to estimate equivalent $R_L$. If consider other possible implementation, for instance, when $R_L$ is lower than 50 Ω, the mismatch caused by $Z_{ref}$ becomes more serious.

Fig. 5-9 presents the worst case scenario caused by the $Z_{ref}$ mismatch. The maximum normalized power gain of this proposed system is shifted to around 9 MHz, while $S_{21}$ at operating frequency 6.78 MHz dropped to 0.6.

To avoid this impedance mismatch, $Z_{ref}$ variation needs to consider in advance. Therefore, Eq. (5-6) is modified as

![Figure 5-9. Normalized power gain of the proposed system caused by $Z_{ref}$ mismatch.](image-url)
In this case, $S_{21}$ at 6.78 MHz will remain at maximum value even coupling is stronger than 0.15, as shown in Fig. 5-10.

### 5.5 System Consideration

Since $I_{TX}$ only related to $X'$, Eq. (5-9) can be further written as

$$I_{TX} = \frac{VIN \times \pi}{\sqrt{2} \times \omega L_M}$$

(5-12)

for a differential push pull class-E power amplifier. From (5-12), the initial state of the PTU design is determined. Fig 5-11 shows the flow chart of above consideration.

![Figure 5-10. Normalized power gain of the proposed system when $Z_{ref}$ mismatch is taken into consideration.](image)

$$X_{LM} = X_{CM} = \omega L_{TX} - \frac{1}{\omega C_s} + \text{Imag}\{Z_{ref}\} = X'$$

(5-11)
Figure 5-11. Design flow chart of a wireless charging system.
5.6 Frequency Splitting Discussion

Frequency splitting is a phenomenon happened when the coupling between coils are too strong passing the critical coupled point. According to [77], this critical point usually equals to $k = 0.15-0.2$. Once the response falls into the over-coupled region, the power gain between primary coils and secondary coils changes from one peak curve to multi-peak curves [78]-[82]. As a result, the maximum received power is not at the operation frequency but at two or more adjacent frequencies instead. In this sense, the system is no longer under the optimized condition. This phenomenon has been well-studied and several solutions were proposed, which can be categorized into two groups: active method and passive method.

Active method usually requires a control system to avoid the frequency splitting. Some studies utilize the frequency tracking technique to always operate the system at the optimized frequency instead of a fixed resonant point [78]-[79]. However, this solution is not suitable for some consumer electronics, for instance, AirFuel magnetic resonance standard only works at a single frequency 6.78 MHz. Tunable impedance matching/tracking network is another popular solution to maintain the power transfer efficiency no matter the system is in the strong or weak coupling area [80]. Nevertheless, an extra control circuit is needed to dynamically adjust the impedance. In this case, not only additional components are used to monitor and compensate the corresponding impedance, but also the extra power is consumed.

On the other hand, a lot of researches are conducted regarding passive mechanisms. One solution is to suppress the coupling by simultaneously adjust the position or orientation between the coils [81], which is strongly against the user experience. Another solution is to deliberately design non-identical coils [82]. Though,
slight size difference between transmitting and receiving coils improves the frequency splitting phenomenon, to achieve uniform output power by enlarging the size mismatch is not practical. It will introduce serious magnetic flux leakage and reduce the power transfer efficiency. Furthermore, this technique is not suitable for supporting multiple devices with different dimensions.

From Chapter 4, experiments results show that $k'$ is high than 0.15 when the distance between TX and RX is shorter than 5 mm. Therefore, this phenomenon is investigated in the following section and the solution is also discussed.

5.6.1 Frequency Splitting for a Wearable-Sized Coil

Typical frequency splitting phenomenon appears when the TX coil and RX coil are with the same dimensions and the matching circuits in both sides are also the same, as analyzed in [83]. Assume each TX coil cell is as small as a wearable-sized coil, then numerical simulation results is shown in Fig. 5-12, where $L_{TX} = L_{RX} = 5.85 \, \mu H$.

It can be observed that when the coupling coefficient is higher than 0.15, the response starts to split, the power gain of the wearable device drops very quickly. This phenomenon reflects to real experiments will cause low output voltage. In some worst case scenario, the output voltage is too low to even turn on the next stage.
5.6.2 Suppress the Frequency Splitting for the Proposed System

As mentioned in Chapter 4, in order to reduce the leakage of magnetic flux while supporting multiple devices simultaneously, the size of TX coil cell is selected to be 1.5 to 2 times larger than the wearable coil. This size ratio not only optimizes the power transfer efficiency, but also helps to avoid the frequency splitting in the proposed system. Fig. 5-13 illustrates the overall system simulation result by using the measured value from the real demonstration. From Fig. 5-13, it can be viewed that the response won’t start to split until k=0.35, which means it won’t happen in the proposed system. Another advantage is that the coupling coefficient can be further enhanced for different applications without the risk of frequency splitting.
5.7 Measurement Results and Summary

Impedance box is a standard defined by AirFuel resonance [15]. The measured results with the load-independent Class-E power amplifier shows that the transmitting current remains constant within the certain impedance variation, as shown in Fig. 5-14.

Table 5-1 lists the performance comparison with prior works. The proposed system demonstrate the capability against widest load variation.
Figure 5-14. Measured reflected impedance box comparing to the standard reflected impedance box defined by Airfuel Resonance standard. Coil currents expected to be held constant (< 8%) with this $Z_{ref}$ variation box.

Table 5-1. Comparison of Load-Independent Power Amplifier for WPT System

<table>
<thead>
<tr>
<th>Topology</th>
<th>Voltage-source inverter</th>
<th>Class-D</th>
<th>Class-EF</th>
<th>Class-E</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>38.4 kHz</td>
<td>8 MHz</td>
<td>6.78 MHz</td>
<td>6.78 MHz</td>
</tr>
<tr>
<td>WPT application</td>
<td>N/A</td>
<td>Single TX, single RX</td>
<td>Single TX, single RX</td>
<td>Multi-TXs, multi-RXs</td>
</tr>
<tr>
<td>load variation (Ω)</td>
<td>[Re] 0.5-2</td>
<td>N/A</td>
<td>0-5</td>
<td>0-20</td>
</tr>
<tr>
<td></td>
<td>[Im] N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>-180-+180</td>
</tr>
</tbody>
</table>
CHAPTER 6
CONCLUSION AND FUTURE WORK

6.1 Conclusion

Novel 3-D wireless charging systems are proposed and demonstrated in this dissertation. Design methodology is well discussed for multiple consumer electronic applications. The power transfer efficiency and the user experience are both improved through PTU geometry, matching network, and adaptive control system.

Moreover, a new load independent power amplifier is proposed and demonstrated with supporting multiple receiving devices simultaneously. By selectively activating the transmitting coils, the coupling coefficient can be optimized disregarding the position or orientation of the receiving devices.

6.2 Future Work

In addition to AirFuel resonance standard, which is well-studied in this dissertation, Qi standard using inductive coupling is another popular specification in the market. Therefore, a wireless charging system to transmit and receive for both 100-200 kHz and 6.78 MHz frequency ranges can be explored, which will enable many new research topics.

First, to study dual-mode RX and TX planar coils and to overcome misalignment situations. Techniques discussed in previous chapters including activation and input phase control methods can be applied into the new system. More works can be done on the optimization scheme, such as the coil dimension, EM field design and manipulation, as well as the ferrite layer placement.
Second, new matching networks are required to maintain transmitting currents stable over a wider range of the reflected impedance variation. Besides, how to reduce the loss and jitter noise during the switching can be examined.

Furthermore, more discussions should be entered into regarding the entire system, including the rectifier, switched power amplifier, communication channel, and other sub-systems. Besides, the interference between different frequency bands can be investigated and improved.

Finally, the planar dual-mode wireless charging pad can be further expanded to 3-D multi-mode wireless charging system. The trade-off between the user experience and the power transfer efficiency can be studied.
LIST OF REFERENCES


Ms. Ron-Chi Kuo entered University of Florida in 2011 to pursue her Ph.D. degree in Department of Electrical and Computer Engineering. From May to December 2014, she worked as a graduate research intern at MediaTek Corporation in Woburn, MA. During her Ph.D. study, Ms. Kuo won Best Conference Paper in IEEE Wireless Power Transfer Conference, Boulder, Colorado, 2015 and Best Student Paper Competition Second Place in IEEE Wireless Power Transfer Conference, Aveiro, Portugal, 2016. She was also the winner of Graduate Student Challenge in IEEE International Microwave Symposium (IMS), Montreal, Canada, 2012. Her research interests include near-field and far-field wireless power transfer system as well as electromagnetic design and modeling. Ms. Kuo is currently a student member of IEEE.