

**Systematic Design of a 100 W
6.78 MHz Wireless Charging Station
Covering Multiple Devices and a
Large Charging Area**

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Abstract

Wireless power transfer (WPT) promises to offer safe and convenient charging in consumer electronics applications. The application ranges from sub-watt medical implant device charging to watt-level household charging, further to the kilowatt electric vehicle (EV) and railway charging. At present, two industry standards are widely used to guide the WPT product design. The Qi standard describes an application that requires the receiver to be placed close to the transmitter, called "tightly coupled" WPT. The receiver coil typically has a similar size compared to the transmitter. And the operation frequency is in the kHz range. The Airfuel standard, on the other hand, uses the 6.78 MHz ISM band as the operating frequency, which facilitates high induced voltage when the coupling between the two sides is "loosely coupled". Thus, the Airfuel standard can be adopted in applications where the transmitter size is larger than the receiver for multi-receiver applications. Other benefits for the Airfuel standard include high-Q coil design for high system efficiency, small component size, and less interference with neighboring metals.

Targeting the application described by Airfuel, this work presents the analysis and design of a 100 W, 6.78 MHz WPT station covering 0.5 m x 0.5 m and multiple receivers. The station provides three key features: 1) free-positioning: receivers have constant induced voltage anywhere on the transmitter; 2) free-loading: operation status of one receiver do not affect another receiver; 3) high overall efficiency.

To fulfill 1), an interleaved coil is proposed that has various current distributions in each turn. To fulfill 2), a passive impedance matching network is detailed. To achieve 3), soft-switching of MHz converters is analyzed, and a systematic design method is proposed. Using the resulting system parameters from the systematic design, a prototype is built and tested. The experimental results show that the magnetic field variation on the charging surface is

15.9 %, validating the free-positioning function. The free-loading function is verified by the load-change test. The measured transmitter coil current and the induced voltage on receivers are constant. The measured dc-to-dc efficiency at full load is 92.8 %, which is high compared to other systems in the literature.

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Chapter 1

Introduction

Wireless power transfer has been adopted widely in peoples' lives in the past decade due to its flexibility, convenience, and safety. The application ranges from sub-watt medical implant device charging applications to watt-level household consumer electronics applications, further to the kilowatt-level electric vehicle (EV) and rail-way charging. At present, commercial implementations of wireless power transfer are largely limited to low-power, low-efficiency, and a charger pad where the device must be well aligned to a dedicated region. Though the technology continues to progress, this approach fails to capitalize on the promise of spatial freedom and thought-free ubiquitous charging which makes wireless power transfer attractive. Designing a free-positioning and multi-device wireless charging system design is still challenging.

1.1 Wireless Power Transfer Overview

Wireless power transfer (WPT) refers to the transfer of power from a source to an electric load without any wired connection. The idea of WPT was first carried out by Nikola Tesla in the 1900s, using the Wardencllyffe Tower and through electric-field coupling [75]. The idea of WPT ceased to be popular after Tesla ran out of funding. In the next decades, the promise of wireless power was lost until it was reexamined with the advent of vacuum tubes which enable WPT systems at microwave frequency. WPT has been an active research topic for transcutaneous systems for medical implants since the 1960s [69], and for induction

heaters since the 1970s [43]. For modern short-range applications, the magnetic coupling WPT systems [40] have attracted wide attention since the 1990s, and started to be applied for portable equipment such as toothbrushes since the 2000s [45]. At that time, the WPT application in consumer electronics was largely confined to the watt-level range and small gap. In 2007, researchers from MIT demonstrated that magnetic resonant coupling WPT is capable of transferring 60 W across 2 m air gap with around 40 % efficiency [49], which was a huge leap from the previous system. Since then, researchers have been attracted to resonant WPT study [42]. The power level, efficiency, and transfer distance of the modern WPT prototypes are improving rapidly [42].

Compared to traditional wired charging, WPT has obvious advantages that push this technology forward. WPT removes the need for plugging in the charging cables. The elimination of cable connections increases safety and is convenient because it increases ease of use. Furthermore, WPT has the potential to apply seamless charging, which means customers are able to charge devices wherever near a charging pad. Seamless charging reduces charging intervals, therefore reducing the size and weight of the battery. For some special applications such as implant devices [57], WPT is the only preferred solution. The convenience boosts the wide applications of WPT.

1.1.1 Wireless Power Transfer Technologies

Wireless power transfer technologies are classified into near-field (non-radiative) transfer, and far-field (radiative RF) transfer [38], as shown in Fig. 1.1. Based on the different coupled fields, near-field techniques are categorized into inductive coupling, which uses the coupled magnetic field, and capacitive coupling, which uses the coupled electric field. Inductive coupling can be further divided into magnetic inductive coupling and magnetic resonant coupling.

For capacitive coupling, the achievable power level is highly determined by the coupled capacitance, which is related to the face-to-face area of the receiver and transmitter coupler [61]. In consumer electronics applications, the size of either side is typically small and limits the adoption of captive coupling. An alternative way of increasing power level is building a stronger electric field in the constrained coupled region, which, however, forms a threat to

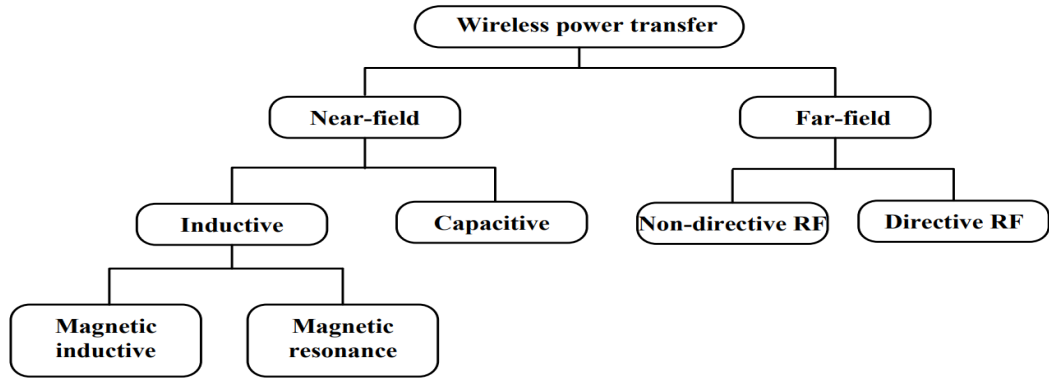


Figure 1.1: Basic principles of WPT.

neighboring devices and humans. Capacitive coupling has been widely demonstrated in EV charging applications [61], where a large capacitance is feasible due to the relatively large coupling area compared to consumer electronics applications. The coupled electric field is typically under the chassis, which isolates the charger from the users [67].

The far-field technique is categorized into direct RF power beamforming and non-direct RF radiation [38]. Lasers, acoustic, and ultrasonics are also used for energy transfer. Though far-field technology can transfer power over a wide air gap, it suffers the limitation of low power level and low-efficiency [38]. Typically, only a small portion of the radiated power is received by the receiver, and the radiated power and field strength is highly limited by safety standards [44]. Far-field WPT is more commonly used in industrial sensor charging, with sub-watt power level and low-efficiency [38].

1.1.2 Inductive Wireless Power Transfer

Though multiple approaches to WPT are being pursued, magnetic inductive coupling power transfer (IPT) has emerged as the most popular approach to WPT over relatively short distances, such as through a table [42].

The initial IPT system introduced to the public uses a similar operating principle to a transformer, where the primary coil and the secondary coil are placed closely and magnetically coupled through the air [42] instead of a ferrite core. An induced voltage will be generated at the receiver coil when the transmitter coil conducts current, even though there is no wired connection between the two sides. The typical application involves low-power consumer electronics such as WPT charging for electric toothbrushes, and WPT charging for cell phones.

A major drawback of the initial IPT is the short transfer distance and the intolerance of position misalignment. The receiver side voltage, which is induced from the coupled magnetic field, is dependent on the field strength. When the receiver side is moved away from the transmitter, or misalignment occurs from the designed position, the coupled magnetic field strength decreases rapidly. Because the transmitter coil current is limited by the leakage inductance, the magnetic field cannot rise to compensate for the rapid decrease of the coupled field, which leads to the decrease of the induced voltage and output power.

To fight the decreased induced voltage, a more complicated converter may be designed in the receiver to maintain the initial output power at the expense of power loss and cost [80], which is not preferred in consumer electronics applications. Another method of increasing the induced voltage is raising the transmitter working frequency, at the expense of higher Tx coil loss [65] and a more complicated control strategy.

An alternative method of maintaining the initial power level is reducing the reactance in the transmitter side, by adding a resonant compensation capacitor, creating a resonant IPT system at the fundamental operating frequency [42]. Under constant input voltage, the maximum achievable Tx coil current increases, which enhances the magnetic field generation ability and increases the transfer distance and power level.

1.1.3 Industrial Standards

The maturity of IPT for consumer electronics was proven by the launch of the Qi standard from the Wireless Power Consortium in 2011 [62]. Qi utilizes two tightly coupled coils and transfers power over millimeters range [62]. The most recent Qi specification, V1.2.3 in 2017, requires the operating frequency of the coupled field between 87 and 205 kHz for the power range from 5 W to 30 W [62]. Qi targets the overall system efficiency to be higher than 70 % [62]. At present, Qi is widely applied by more than 100 companies in their wireless charger design. More than 7000 products have passed Qi certification [12] and the number is increasing rapidly.

Typically, Qi describes a WPT application called “tightly coupled WPT”. In this application, the receiver device must have a similar size as the transmitter pad, and the receiver must be placed at a dedicated place where the two coils are overlapping with each other. Thus, Qi fails to offer positional flexibility, where customers can place their devices anywhere on a large charging mat. It also fails to offer multi-load charging capability due to the similar size of transmitter and receiver. A typical Qi charger is shown in Fig. 1.2a. Recently, a new Qi product that assembles multiple transmitter coils into a large coil array was released [7] for a large charging area and offers multi-load capability, as shown in Fig. 1.2b. Fig. 1.2c is a schematic of the coil array consisting of three small coils. Unfortunately, the multi-receiver product stills require a dedicated space for each receiver



Figure 1.2: Qi-certified IPT product (a) INIU wireless charger, for single receiver application; (b) ZealSound 3-in-1 charger for large charging area and multi-load application; (c) ZealSound 3-in-1 charger coil array schematic.



Figure 1.3: Comparison of two WPT applications (a) Qi charger, and (b) Airfuel charger.

device, failing to capitalize on the promise of spatial freedom and thought-free ubiquitous charging which makes wireless power transfer attractive. Besides, the price of the new transmitter is prohibitively expensive, compared to a simple IPT charging pad for a single receiver.

To foster the advantage of long-distance, high efficiency, and high power level, MHz resonant IPT is developing rapidly which has enhanced induced voltage compared to the kHz Qi IPT. In 2012, Alliance for Wireless Power (A4WP) was launched for MHz resonant IPT. A4WP is intended to create a resonant WPT standard to compete with the existing Qi standard [76]. The standard they proposed is called “Rezence”, which supports power transfer up to 50 watts over distances up to 5 cm. In 2015, A4WP merged with Power Matters Alliance (PMA) forming a new alliance called “AirFuel Alliance” [76]. And the new version of the MHz resonant IPT standard is called “Airfuel”. Compared to the Qi standard, Airfuel raises the working frequency to 6.78 MHz, which is one of the ISM bands. At 6.78 MHz, the magnetic field generates a higher induced voltage on the receiver compared to kHz operation, thus reducing the coupling requirement and expanding the transfer distance and charging area. The low-coupling operation capability also enables multi-receiver charging with one large transmitter coil.

Compared to the Qi standard, Airfuel describes an application called “loosely coupled WPT”. This application models a wireless charging table or desk. In this application, the transmitter and receivers are isolated by an air-gap or table in the centimeter range, and the size of the transmitter could be designed larger than the transmitter pad, expanding charging area. Thus, Airfuel standard potentially supports the promise of spatial freedom and multi-load charging. Fig. 1.3 compares the two different applications described by Qi and Airfuel.

Aside from the two standards for watt-level consumer electronics application, SAE International is working on specifications for high-power IPT application in electric and plug-in hybrid vehicles [10]. In 2013, standard SAE J2954 for EVs was announced. The operating frequency was set to around 85 kHz (range from 81.38-90 kHz). Three power levels are proposed: 1) WPT 1, whose power rating is 3.7 kW; 2) WPT 2, whose power

rating is 11 kW; and 3) WPT 3, whose power rating is 22 kW. At present, EV charging is still being researched.

Key features of the three types of standards are summarized in Table 1.1.

1.1.4 Commercial Product Overview

At present, more than 7000 Qi-certificated commercial products are available in the market [12]. Though the Airfuel certification is still being investigated, there are products available for engineering design purposes targeting 6.78 MHz Airfuel WPT. The products are classified into three categories [41] 1) complete WPT transmitter and receivers systems, 2) discrete coil and power converter modules, and 3) chipsets that integrate power stages and WPT logic into a single chip for designing modules or systems.

For 1), there are Qi-certificated charging pads that integrate the transmission coil, the power converters, and the IPT control logic. Customers can use the system directly to charge their mobile phones, pads, and monitors, and laptops with a built-in WPT module.

For 2), companies are focusing on a specified discrete stage instead of the entire transmitter or receiver system. For example, NuCurrent focuses on designing thin and low-loss PCB coils for both Qi and Airfuel standards. Other major coil manufacture includes Würth Electronics and TDK. The quality factors of the commercial discrete coils are summarized in Table 1.2. Efficient-Power-Conversion (EPC), on the other hand, develops high-efficiency power amplifiers using GaN technology. EPC9083 is one of their class-E amplifiers designed for 6.78 MHz WPT [12]. EPC9083 is one of the class-D modules for WPT. The typical efficiency of their discrete inverter modules ranges from 92 % to 97 %, which varies with products, working frequency, and power level. EPC also delivers receiver boards consisting of rectifiers and the subsequent dc-dc. EPC 9513 for example, is a 5 W, 6.78 MHz receiver power module consists of a Schottky diode rectifier and a SEPIC converter.

In 3), various stages in a WPT system are integrated into a chipset. NXQ1TXH5, for example, is a 5 W Qi-certified transmitter chipset produced by NXP. It integrates a full-bridge inverter, drivers, and logical functions into a 5 mm x 5 mm chip. The reported peak ac-dc efficiency is 75 % when charging a typical 5 W receiver [8]. BQ501210, which is a product of Texas Instruments, integrates the logical functions and drivers into a chipset.

Table 1.1: Comparison of IPT standards.

Standard	Qi	Airfuel	SAE
Technology	Inductive & LF resonant	HF resonant	LF resonant
Frequency	87-205 kHz	6.78 MHz	85 kHz
Multi-receiver	No	Yes	No
Power-level	5-30 W	Up to 50 W	Up to 22 kW
Typical efficiency	65-75 %	75-85 %	90-95 %
Major companies	Apple, Samsung	Dell, EPC	BMW, Ford

Table 1.2: Commercial coil performance review.

Part number	Wire	Frequency	Inductance	Q	Size [cm]	Thickness [mm]
1461794011	PCB	200 kHz	5.6	17	Dia. 2.7	3.7
WT505090	2 mm solid	200 kHz	6.3	66	Dia. 5.1	3.3
760308102144	2 mm Litz	200 kHz	6.3	195	Dia. 5.3	7.0
1461790001	PCB trace	200 kHz	17.3	22	5.0x4.3	0.57
1461794001	PCB trace	200 kHz	8.9	25	6.6x4.8	0.57
1461791001	PCB trace	200 kHz	11.6	22	4.3x3.1	0.53
1461790001	PCB trace	6.78 MHz	2.6	87	Dia. 2.7	0.53
1461794001	PCB trace	6.78 MHz	2.9	148	7.2x3.6	1.15
1461791001	PCB trace	6.78 MHz	4.0	105	Dia. 3.5	0.69

It is designed to be used with BQ500101 power stage [2] to form a two-chip transmitter solution. BQ501210 can support both 5 W receiver and 15 W receiver. The peak ac-dc efficiency is 72 % when charging a typical 5 W receiver, and rises to 82 % for 15 W receiver [1]. Texas Instrument also has a receiver chipset that integrates rectifier, post dc-dc, and the WPT logical function into a single chip. BQ51050B, for example, is a 5 W receiver and the peak ac-dc efficiency is 93 % [3]. Compared to discrete modules, the integrated chips facilitate ease of use.

Fig. 1.4 shows a WPT system fabricated from commercial discrete modules. The efficiency of the commercial Qi systems is summarized in Fig. 1.5. The 5 W systems use the results from NXQ1TXH5 [8] and BQ501210 [2]. The 15 W system uses the result from BQ501210 [2]. The 65 W system uses a reference design by NXP. The discrete systems consist of high-Q transmitter and receiver coils from Nucurrent [9], and high-efficiency power stages from EPC [5]. Typically, the Qi system has an efficiency of around 70 %. The efficiency may increase with power level but is usually less than 85 % even using high-performance discrete modules.

1.2 Summary and Motivation

IPT promises to offer safe and convenient charging in consumer electronics applications. Over the past decade, IPT technology has drawn wide attention from the industry. The application ranges from sub-watt medical implant device charging applications to watt-level household consumer electronics applications, further to the kilowatt-level EV and rail way charging.

In consumer electronics applications, two industry standards are widely used to guide the WPT product design. Qi standard operates within the 205 kHz range and describes the “tightly coupled WPT” application. The receiver device must have a similar size as the transmitter pad, and the receiver must be placed at a dedicated place where the coupling is high. Airfuel standard operates at 6.78 MHz ISM band for increased induced voltage and describes the “loosely coupled WPT” application. The transmitter and receivers are

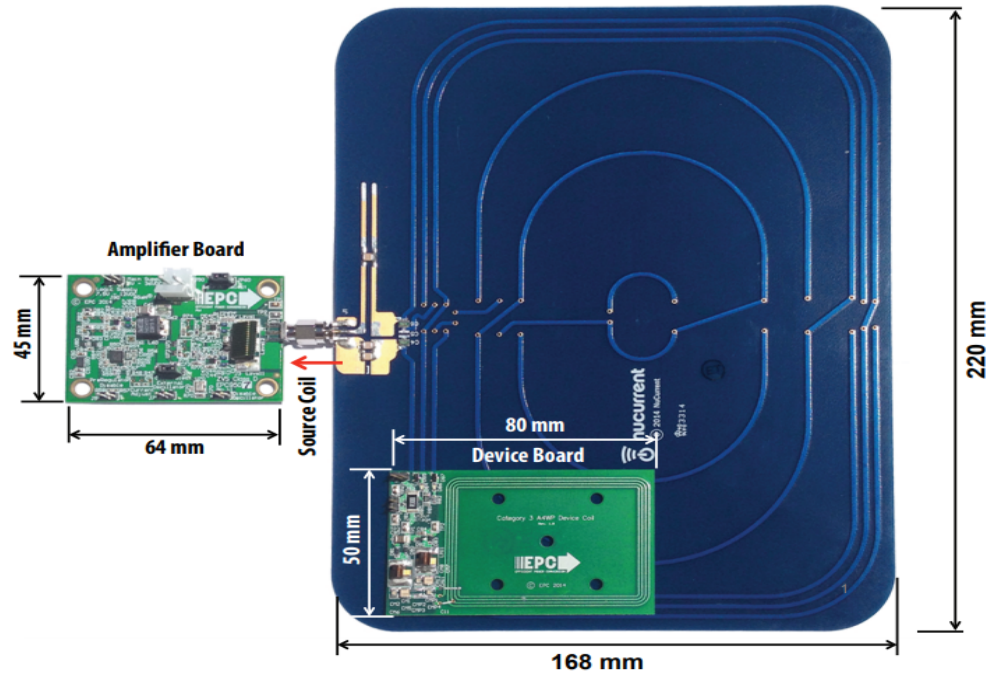


Figure 1.4: IPT system assembled from commercial discrete modules.

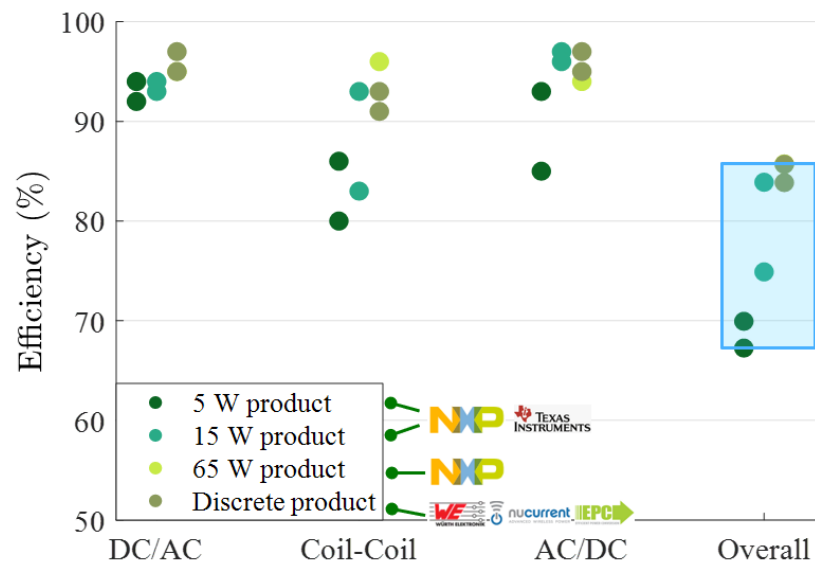


Figure 1.5: Efficiency of commercial Qi products.

isolated by an air-gap or table in the centimeter range, and the size of the transmitter could be designed larger than the transmitter pad expanding charging area.

Over the years, most of the commercial Qi products are largely limited to a charger pad where the device must be well aligned to a dedicated region. The Qi products also suffer low-efficiency issues even using the high-performance discrete modules. If moving the operating frequency to 6.78 MHz, the enhanced induced voltage on the receiver reduces the coupling requirement and expands the charging area, which potentially offers positional flexibility and enables multi-receiver WPT. The high-frequency operation also facilitates a high-Q coil design for high system efficiency, small component size, and less interference with neighboring metals due to increased skin effect and AC resistance to eddy current. With continuous interest in high-efficiency, free-positioning, and free-loading wireless charging, there is strong motivation to apply Airfuel 6.78 MHz WPT and explore the maximum performance with this standard. The desired characteristics for WPT charger in consumer electronics are summarized in Table 1.3.

Despite the promises, designing a 6.78 MHz resonant WPT with free-positioning, free-loading, and high-efficiency has significant challenges to overcome. The goal of this dissertation is to explore the design and performance of the WPT charging station which offers free-positioning, free-loading, and high efficiency.

1.3 Dissertation Outline

This dissertation is organized as follows. Chapter 2 is the literature review of the state-of-art method to achieve the aforementioned three features of free-positioning, free-loading, and high-efficiency. Based on the literature review, Chapter 3 presents the proposed WPT system structure and discusses how it overcomes the disadvantages in the literature. A new systematic design method that co-considers all stages is proposed and discussed. Chapter 4, 5, and 6 discuss the modeling and design of each individual stage. Chapter 7 is a systematic design integrating all individual stages. A 6.78 MHz WPT prototype is presented which supports the system modeling and design method, verifying that the desired features in Table 1.3 have been achieved. Chapter 9 concludes the dissertation and proposes the future

Table 1.3: Desired Characteristics for WPT Charger in Consumer Electronics Applications.

Requirement	Description
Free-positioning	Receiver gets constant induced voltage anywhere on a large area
Free-loading	Multi-load capability, no cross-channel disturbance
High-efficiency	Generally required for any power supply

work, where a novel self-resonant coil is discussed that has a high-Q, low leakage E-field, and compact size.

Chapter 2

Literature Review

In consumer electronics applications, the only commercialized solutions for wireless power have been within the kHz frequency band over the past few years. This is because using alternative, higher frequencies such as 6.78 MHz requires design attention. But with the advancement in power electronics and magnetic design, there is finally hope to realize a 6.78 MHz prototype and explore the limits of its performance. This chapter reviews the fundamentals as well as the state-of-the-art WPT system for consumer electronics applications, specifically focusing on the work to achieve free-positioning, free-loading, and high efficiency.

2.1 IPT System Structure

Among the multiple approaches to WPT, resonant IPT has emerged as the most popular approach to WPT over relatively short distances. The structure of a general resonant IPT system is shown in Fig. 2.1. Aside from the front-end PFC and DC/DC converter, the resonant IPT system consists of an inverter, a transmitter coil with its compensation capacitor, a receiver coil with its compensation capacitor, and a rectifier with the load. Both transmitter and receiver coil are usually tuned to be resonant around the operating frequency to minimize reactive power for high efficiency and power level [42]. In practical applications, a DC/DC converter may be used between the rectifier output and the battery charger IC

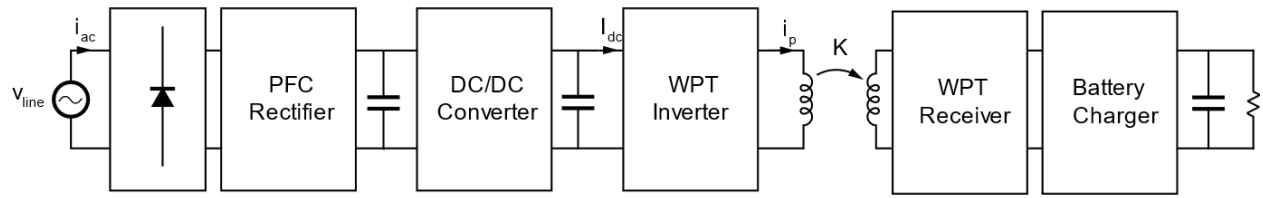


Figure 2.1: Basic structure of a resonant WPT system.

[80], and an impedance matching network may be used between power stages and coils [14] to tune the circuit impedance.

Equivalent circuit model and fundamental harmonic analysis are often used as basics in the IPT system study. A simplified fundamental circuit model of the IPT is shown in Fig. 2.2, which is widely used in kHz WPT systems [42, 53] assuming the converter losses are negligible compared to coil losses. In the transmitter, a voltage source represents the inverter and provides an AC in the primary coil to generate the coupled magnetic field to the receiver. In the receiver, the induced ac voltage on the receiver coil is transferred to the ac load, which represents the rectifier and its load. The resistances R_1 and R_2 model any parasitic resistance in the primary and secondary sides, predominately the resistance of the two coils. Impedances Z_p and Z_s model any additional reactive components inserted into each side. The coupling between two sides is modeled using coefficients k . The induced voltage in the receiver is modeled $V_{ind} = \omega_s k \sqrt{L_p L_s} I_t$.

2.2 Free-positioning Design

WPT systems are desired to offer “free-positioning”, where the receiver gets constant induced voltage anywhere on a large area. Examining the circuit in Fig. 2.2, to achieve free-positioning, receivers should get constant coupling on a large charging surface so that the induced voltage is constant. This requires that the transmitter coil is designed to generate a uniform magnetic field on the surface.

Previous research efforts achieving the uniform magnetic field are classified into three categories: 1) model-based geometric optimization for minimized field variation [28, 55, 19], 2) coil array integrating multiple small coils [31, 78], and 3) coil current manipulating for field shaping [30].

For 1), references [28, 19] model the magnetic field distribution of a rectangular coil using the geometric parameters, i.e. the number of turns and the length and width of each turn. Based on the model, the geometric parameters are carefully designed to achieve a minimum magnetic field variation under the application size limit. The optimized coil shape is presented in Fig. 2.3a [19]. And the corresponding uniform magnetic field is shown in

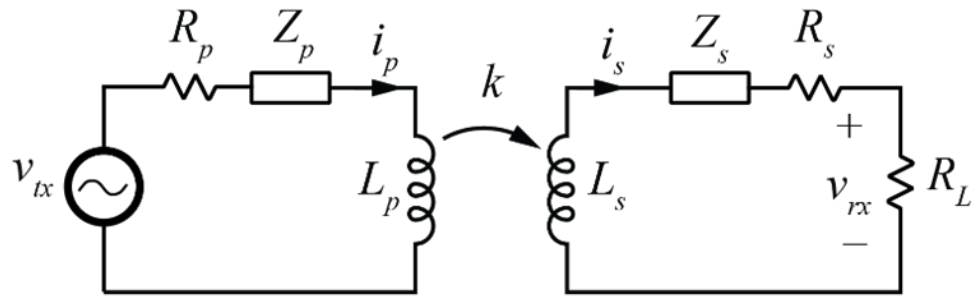
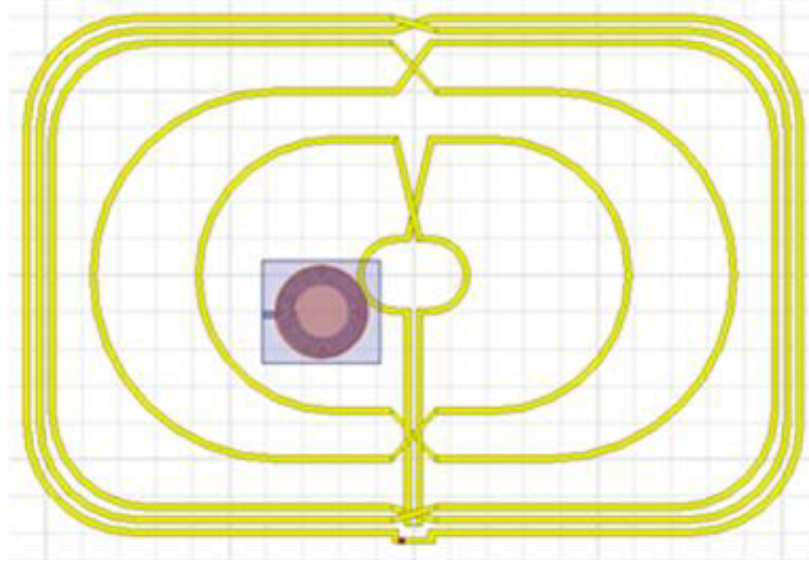
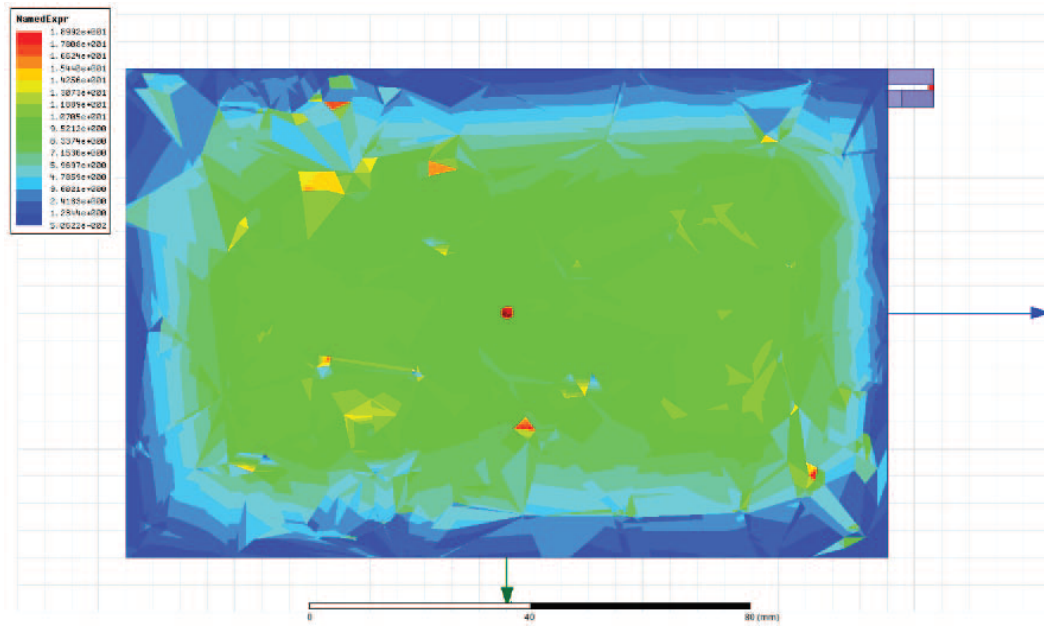


Figure 2.2: Simplified fundamental model of a magnetically-coupled WPT circuit.



(a)



(b)

Figure 2.3: Geometrically optimized coil with uniform magnetic field, coil structure (a) [19] and the field distribution (b) [28].

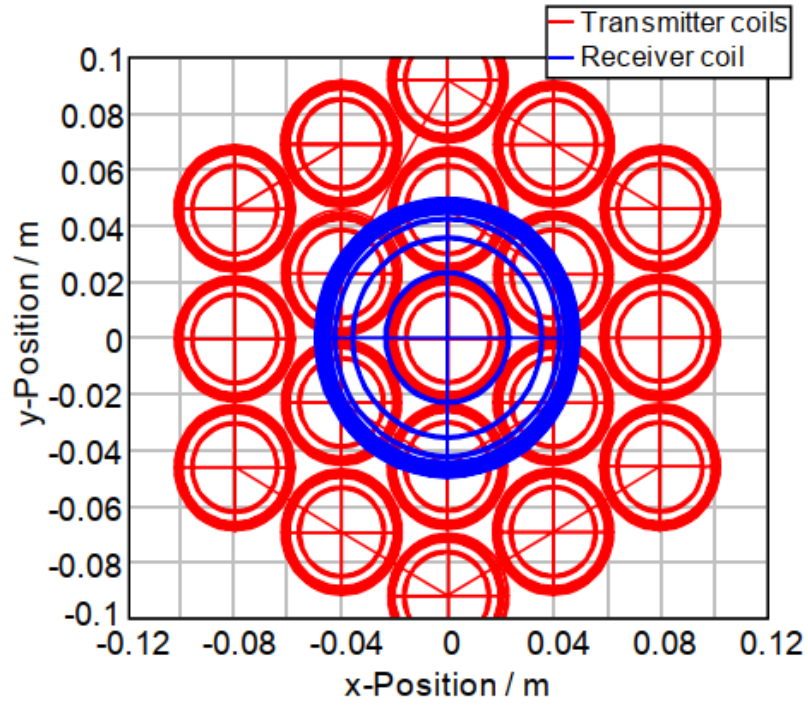
Fig. 2.3b. It should be noted both terminals of this coil 2.3a are connected through the outer side, forming a symmetric structure that exhibits a magnetic field with less variation than traditional coils, which are connected through the inner and outer sides. In reference [55], the geometric optimization was applied to a circular coil and achieves less than 10 % field variation. The geometric optimization method has been verified in the kHz systems [28, 19]. However, this method requires an increased number of turns for finer control of the field distribution to reduce variation, which may lead to a total wire length that potentially approaches $1/4$ wavelength, causing radiation and low-efficiency problems [49]. For a 6.78 MHz wireless charging table with 0.5 m length and 2 m perimeter, the maximum number of turns is around 5 which yields a wire length approaching 11 m, $1/4$ wavelength at this frequency.

For 2), reference [78] presents the design of a coil array that consists of multiple small circular coils. The overall magnetic field is shaped by integrating the small coils into various coil arrangements and controlling the positions of each small coil. The schematic of one coil arrangement is shown in Fig. 2.4a [50]. And the corresponding uniform magnetic field is shown in Fig. 2.4b [50]. However, the overall quality factor of the coil array approximates that of each small coil [50], which suffers a low-Q issue due to the limited size.

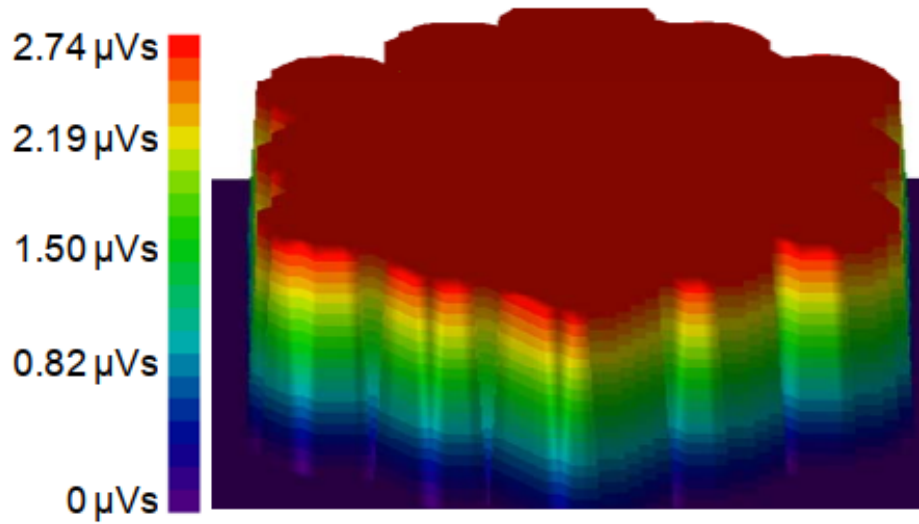
For 3), reference [30] reports a three-dimensional (3D) charging bowl that consists of five discrete coils. The amplitude and phase of the current in the small coils are not identical. Instead, they are separately controlled by a set of digitally controlled inverters. The overall field is manipulated by controlling the current in the small coils. The picture of the 3D bowl together with the multiple converters is shown in Fig. 2.5a [30]. And the schematic of the current manipulation is shown in Fig. 2.5b [30]. This method requires multiple converters to control current in different coils. The duplicated converter may introduce high power loss to the WPT system. This method also requires complicated sensing and control strategy.

2.3 Free-loading Design

WPT systems are desired to offer “free-loading”, where the operation of one receiver has no cross-channel disturbance to other devices. Previous literature have made efforts to solve

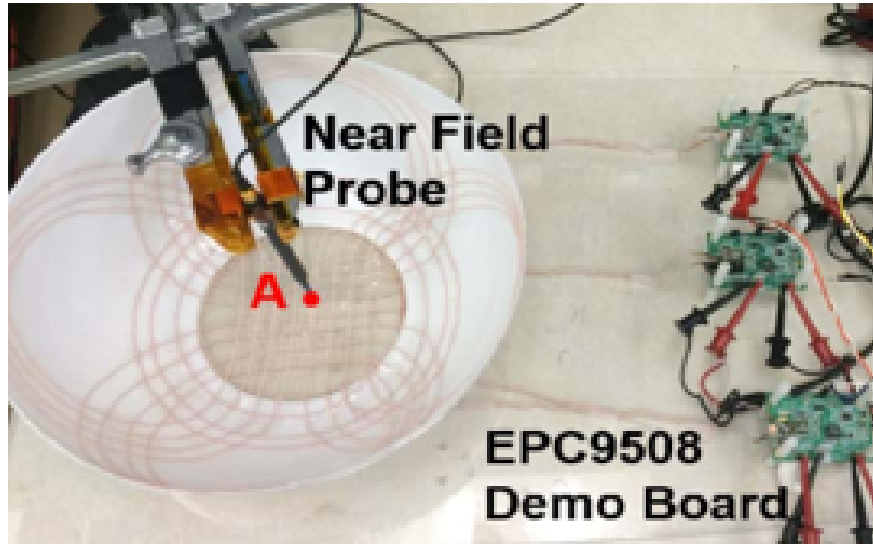


(a)

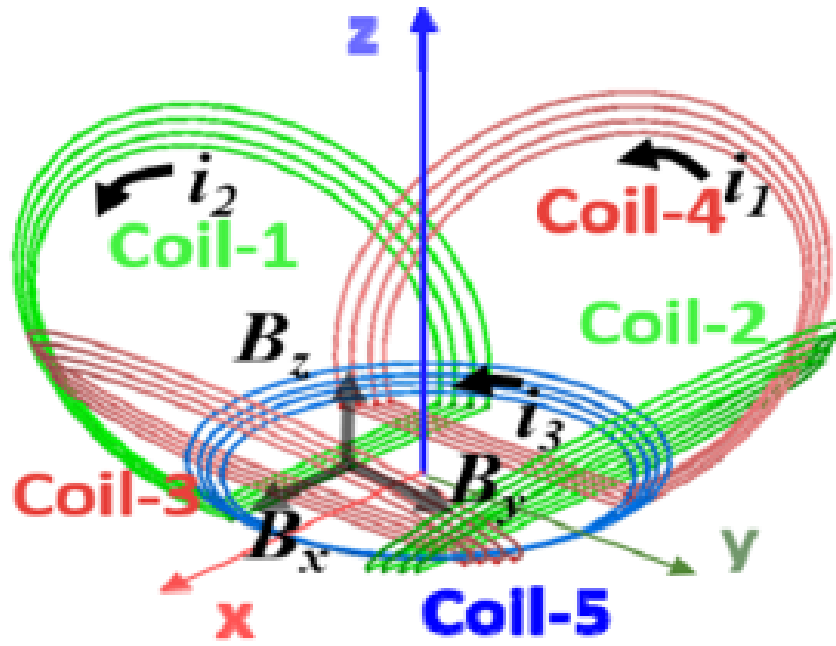


(b)

Figure 2.4: Coil array schematic (a) [50] and the field distribution (b) [78].



(a)



(b)

Figure 2.5: Bow coil schematic (a) [30] and the field distribution (b) [30].

this problem on either 1) transmitter side, or 2) receiver side. On the transmitter side, if the transmitter coil current is constant regardless of the load change, the generated magnetic field and induced voltage in other receivers are constant. On the receiver side, if the power converter is controlled to generate a constant output DC regardless of the magnetic field change, the receiver gains resistance to cross-channel disturbance.

2.3.1 Constant Transmitter Coil Current Regardless of Loading

To achieve a constant transmitter coil current, references [42, 53, 52] use a closed-loop control method. This method senses the transmitter coil current and then controls the transmitter inverter or the front-end dc-dc converter for a constant coil current. Pulse-width-modulation (PWM) [53] is widely used in controlling dc-dc stages. For high-frequency inverters, pulse-density-modulation (PDM) with constant duty-cycle [52] is more suitable because it maintains soft-switching and therefore reduces power loss [52]. Using PDM, the inverter is turned on for a designed number of cycles and then turned off for another designed number of cycles. The averaged output voltage is controlled by the ratio of on and off cycles. Though the averaged output current amplitude is constant, the amplitude increases when the inverter is turned on and decreases when the inverter is turned off. The block diagram of the PWM closed-loop control is shown in Fig. 2.6. Both PWM and PDM control require complicated sensor and controller design.

An alternative method to achieve a constant transmitter coil current is using a passive immittance matching network (IMN) [46], eliminating the requirement of sensor and closed-loop controller design. The circuit schematic of the IMN is shown in Fig. 2.7, using fundamental harmonic analysis. When L_{imn} resonates with C_{imn} , i_t is independent of Z_{n2} thus achieving constant current [46]. However, the passive components conduct high resonant current which may add high power loss to the system, if not designed properly. Unfortunately, previous literature [46] only demonstrates the constant current behavior, and presents no design method in a WPT system.

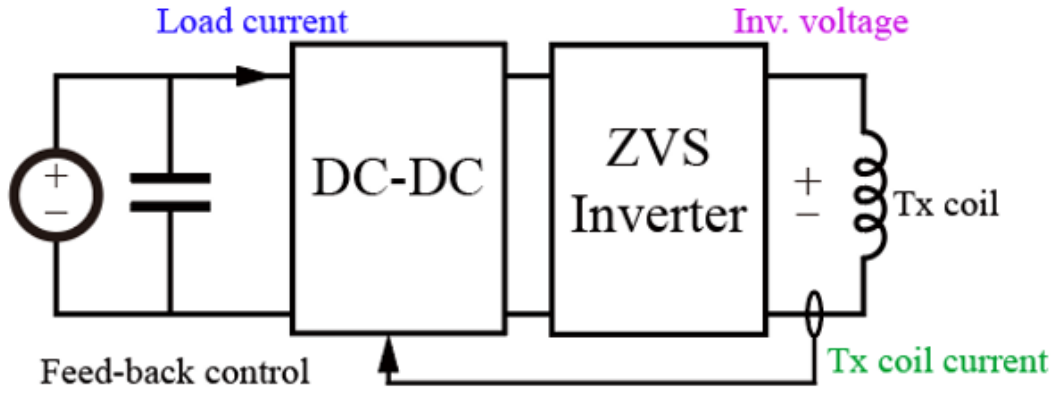


Figure 2.6: Block diagram showing the closed-loop control method to achieve constant coil current.

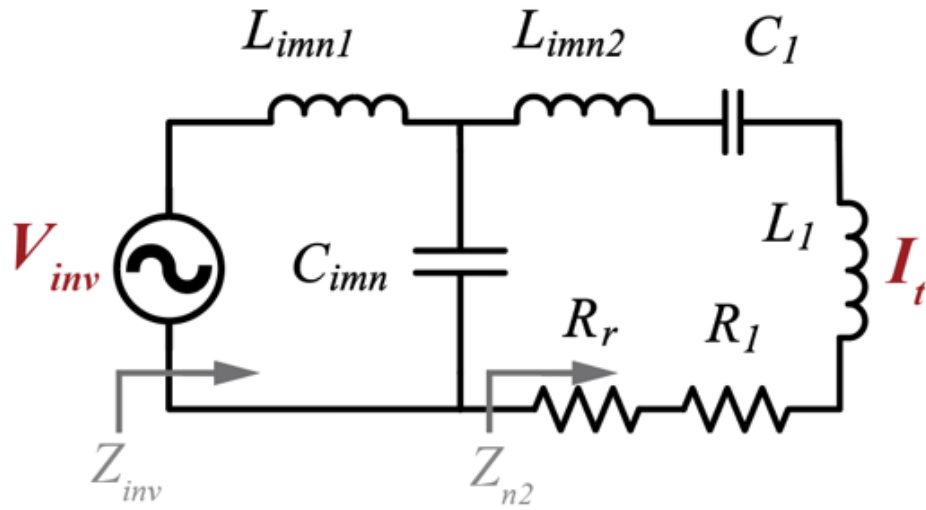


Figure 2.7: Circuit schematic of the IMN.

2.3.2 Constant Rectified Voltage using Active Converter

Free-loading can also be achieved on the receiver side regardless of the change of transmitter coil and field. Reference [66] uses an active rectifier to regulate the output voltage, thus gaining resistance towards the change of transmitter coil current and receiver induced voltage. However, 6.78 MHz synchronization needs careful sensor design and PLL control and requires an additional inductor for soft-switching operation, which adds power loss and system complexity. Reference [53] uses a diode rectifier and post-dc-dc for constant rectifier voltage. Still, the additional dc-dc adds power loss and system complexity. In consumer electronics applications, the receiver side is usually designed to have a small volume [42].

2.4 High-efficiency Design

Examining the WPT circuit in Fig. 2.2, the efficiency of the circuit is maximized [53] when $Z_s = -j\omega_s L_s$ and

$$R_L = R_s \sqrt{1 + k^2 Q_1 Q_2} \quad (2.1)$$

where ω_s is the angular switching frequency, $Q_1 = \omega_s L_1 / R_1$ and $Q_2 = \omega_s L_2 / R_2$.

At this load resistance, the maximum efficiency is achieved [53, 42, 17, 18]

$$\eta_{max} = \frac{k^2 Q_p Q_s}{1 + \sqrt{1 + k^2 Q_p Q_s}} \quad (2.2)$$

From equation (2.2), effort must be made to increase the quality factor of transmitter and receiver coils for high system efficiency. The benefit from high-Q coil design is shown in Fig. 2.8. Thus, many systems are designed following the sequential design methodology that first design high-Q coils, then design the converter operation to achieve the optimal load (2.1) for maximized efficiency [53, 42, 17, 18].

The sequential design method has been verified in kHz WPT systems [53, 42, 17, 18]. However, for MHz system, the inverter design must consider the soft-switching condition to avoid excessive switching loss and system failure [46], instead of solely considering the output

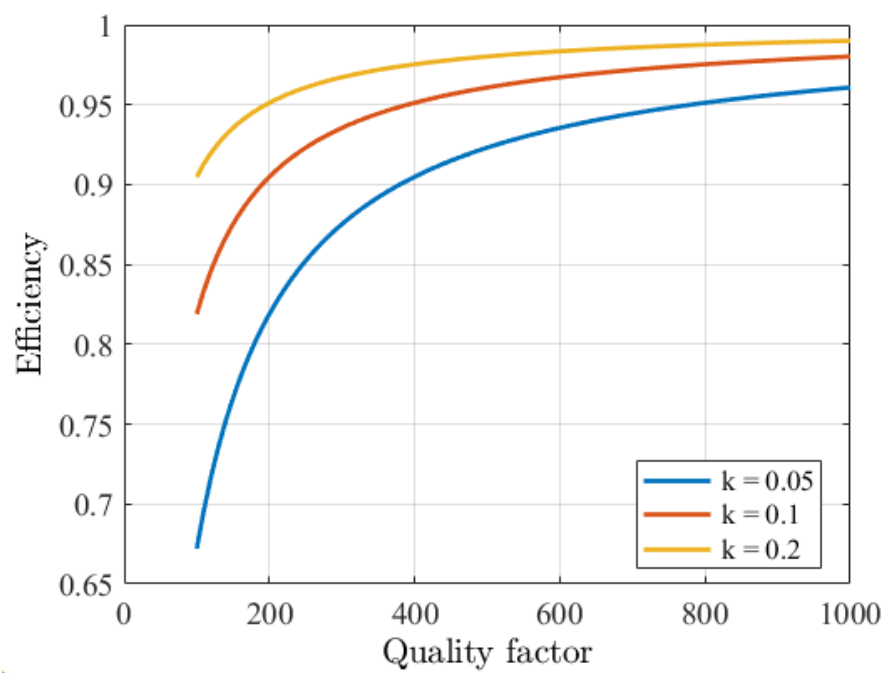


Figure 2.8: Maximum achievable efficiency vs. various coil design.

voltage and current. MHz diode rectifier also exhibits different characteristics compared to kHz operation, which needs considering when designing a high-efficiency system.

2.4.1 High-Q Coil Design

At high frequency, in addition to DC copper resistivity, an increased copper loss incurs in the coil windings due to the eddy current effect. The skin effect is current increasingly crowding to the surface of the conductor due to the presence of eddy current inside the copper, which is caused by the magnetic field inside the copper. The effective conduction depth (termed as “skin depth”) is reduced to

$$\delta = \sqrt{\frac{\rho_{copper}}{\mu\pi f_s}} \quad (2.3)$$

where ρ_{copper} is DC resistivity, μ is the absolute magnetic permeability of the conductor, and f_s is switching frequency. Fig. 2.9 shows the skin depth vs. frequency.

The current distribution becomes more crowded considering the additional eddy current incurred in the adjacent surfaces of neighboring turns, which is produced by the magnetic field of the neighboring turns. This effect is termed the “proximity effect”. The skin and proximity effect in total is called the “eddy current effect”. The eddy current reduces effective conduction area and leads to an increased ESR.

Wire Selection for Increased Effective Conduction Area

One way of reducing the skin effect loss and increasing the effective conduction area is using Litz wire [17, 79, 60, 51]. Litz wire maximizes effective conduction area by weaving together smaller, individually-insulated strands to form larger bundles and wires. By making the individual strand thinner than a skin-depth, it does not suffer an appreciable skin effect. Furthermore, with the weaving of the wires, the length and impedance of each wire are equalized for identical current sharing when conducting ac, reducing the skin effect.

Litz wire is widely used in applications with a frequency range from tens of kHz to hundreds of kHz, and exhibits reduced loss compared to solid wire [17, 53]. When the operating frequency increases to the MHz range, the requirement for the diameter of each strand becomes increasingly difficult to meet, as shown in Fig. 2.9, and the wire becomes

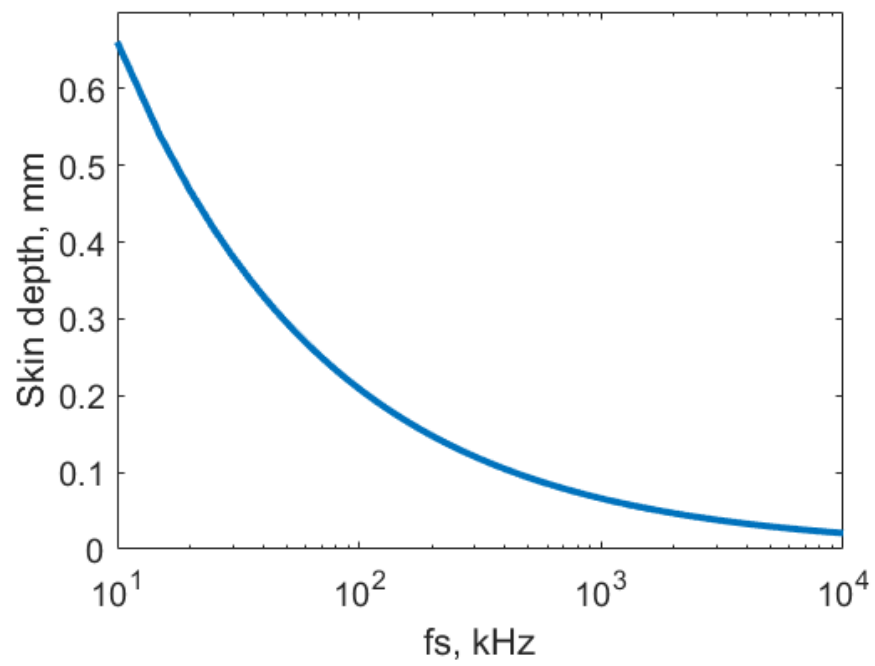


Figure 2.9: Skin depth vs. frequency.

prohibitively expensive [72]. Litz wire also has skin-effect in the bundle level, and proximity effect caused by the current in adjacent bundles [71]. The eddy current effect in the bundle level is another limiting factor for increasing the application frequency [71].

In MHz applications, thin copper foil with a comparable thickness to the skin depth is not expensive and suffers less current crowding [72]. Thus, copper foil is widely used in the MHz frequency range for high-Q coil design [50, 68, 33, 32]. Further, researchers proposed a planar Litz structure that twists copper foils in a similar way as conventional Litz wire for balanced impedance in each strand, achieving equal current sharing with reduced skin effect loss [79, 60]. The structure of an example planar Litz is shown in Fig. 2.10a [79]. Extending from this idea, the foil may be twisted on a tube and form a three-dimensional (3D) Litz to reduce the skin effect. Fig. 2.10b shows how the foil is weaved along a substrate tube forming a three-dimensional (3D) Litz called “surface spiral coil” (SSC) [51].

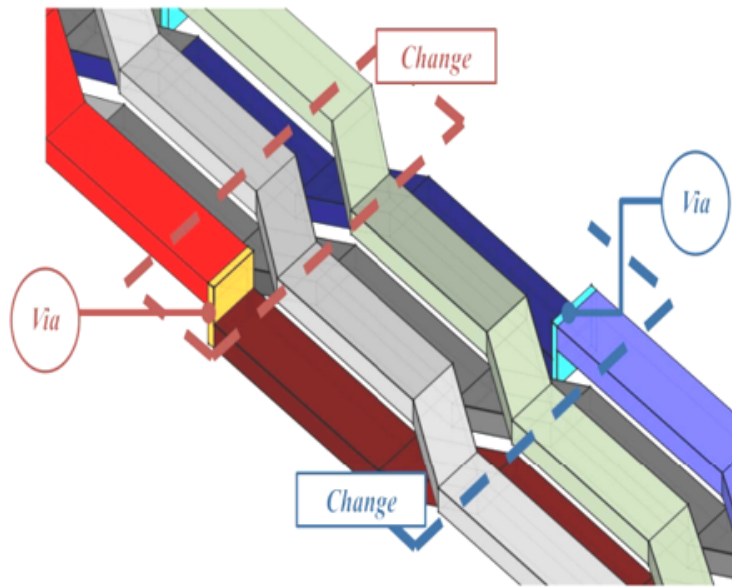
The skin effect in various wires is summarized in Fig. 2.11, where the red region shows the current crowding.

Model-based Geometric Optimization for High-Q

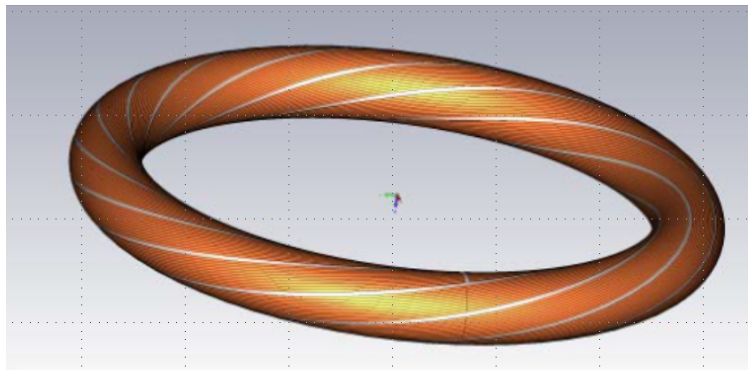
Aside from using different wires to increase effective conduction area and Q, model-based geometric optimization is another widely used method [64, 39, 48, 65]. Coil geometric optimization includes coil modeling that bridges Q with geometric parameters (i.e. the number of turns, width, and pitch), and an optimization process that specially designs the coil geometry to increase Q under certain constants [64]. Coil modeling methods have been widely studied for solid copper wire [39], Litz wire [48], copper tube [65], and planar PCB coils [20]. The modeling method potentially can be used to develop a geometric optimization design method for minimum ESR as in reference [64].

Self-resonant Coil for Reduced Compensation Capacitor Loss

In many reported IPT systems, discrete capacitors are widely used to compensate for coil reactance. Film and ceramic capacitors are commonly used [17, 52, 47, 23] because of their good thermal stability and the excellent loss characteristics [17]. However, the compensation capacitor loss may dominate over the coil loss [17].



(a)



(b)

Figure 2.10: Picture showing the schematic of the planar Litz (a) [79], and 3D Litz (b) [51].

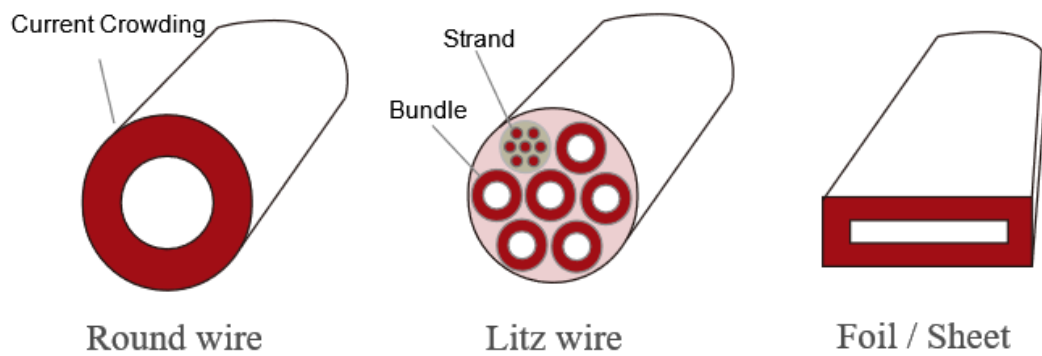


Figure 2.11: Skin effect in various wires.

To reduce the capacitor loss, coil parasitic capacitance is used to compensate coil inductance, eliminating the external lumped capacitor and forming a self-resonant (SR) coil [21, 72, 26, 48, 70, 65, 20, 39]. Coil parasitic capacitance exists in neighboring structures such as the adjacent turns of a tightly wound coil [39], the adjacent strands in a Litz wire [48], and the neighboring conductor layers in a parallel-plate-shaped coil [21, 72, 70]. More specifically, reference [21] uses two copper layers of a PCB coil as the parallel-plate structure providing the parasitic capacitance. Reference [72] is a multiple layer ring-shaped planar inductor with the adjacent layers forming the parallel-plate structure and providing the parasitic capacitance. Reference [70] adds an auxiliary conductor layer on top of the coaxial cable forming the parallel-plate structure and capacitance. The coil structures and connection methods are shown in Fig. 2.12 [21], Fig. 2.13 [72], and Fig. 2.14 [70].

Due to the exposed electric field in the air, coils using the parasitic capacitance between adjacent turns [39, 48] have the potential to interfere with, or damage, nearby electronic components. In contrast, the parallel-plate-shaped coils use a confined electric field between conduction layers inside a dielectric [21, 72, 50]. The confined field is non-sensitive to neighboring objects, making the parallel-plate-shaped resonant coils safer and have a stable capacitance.

In the above-reviewed coils, the parasitic capacitance is in parallel with the coil inductance, forming a parallel LC resonance. The coil exhibits zero impedance to AC voltage sources due to the parallel capacitance, thus requiring additional impedance matching elements when connecting to a voltage source inverter (VSI). The impedance matching elements increase loss, space, and cost. The majority of reported self-resonant WPT coils to date exhibit a parallel LC resonance and require impedance matching when connecting to a VSI [21, 72, 48, 70, 65, 20, 39].

In [26], a self-resonant (SR) coil is reported with series LC impedance and is suitable to directly connect to a VSI without impedance matching. Coil prototype and structure are shown in Fig. 2.15a and Fig. 2.15b. This coil uses multi-layer copper forming the parallel-plate capacitance. The copper layers are blended to force current flow in a circle, forming the inductance. Due to the single turn inductor structure, this coil produces a limited amount

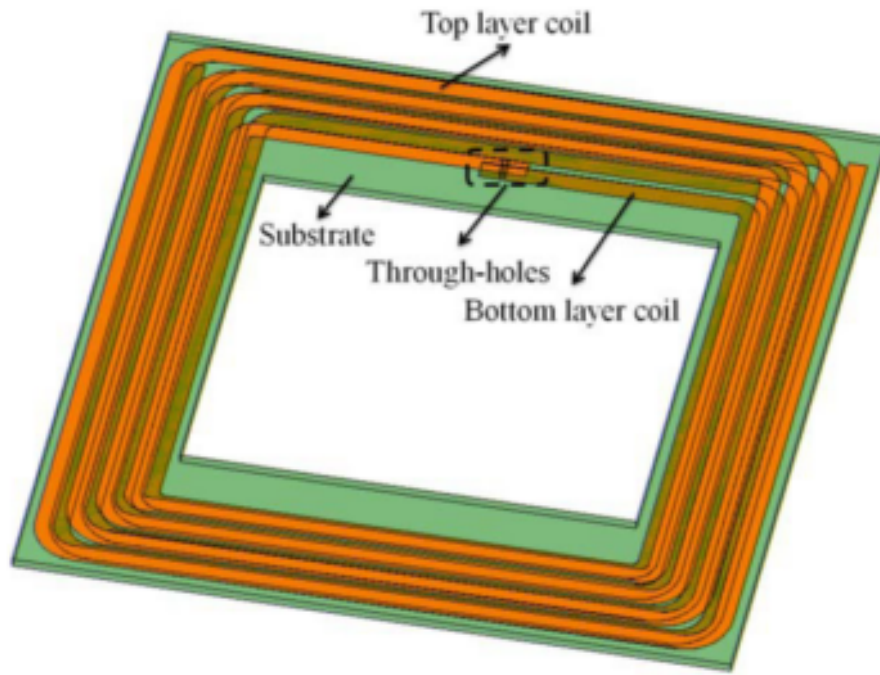


Figure 2.12: Coil structure and connection method of the double-layer PCB coil [21].

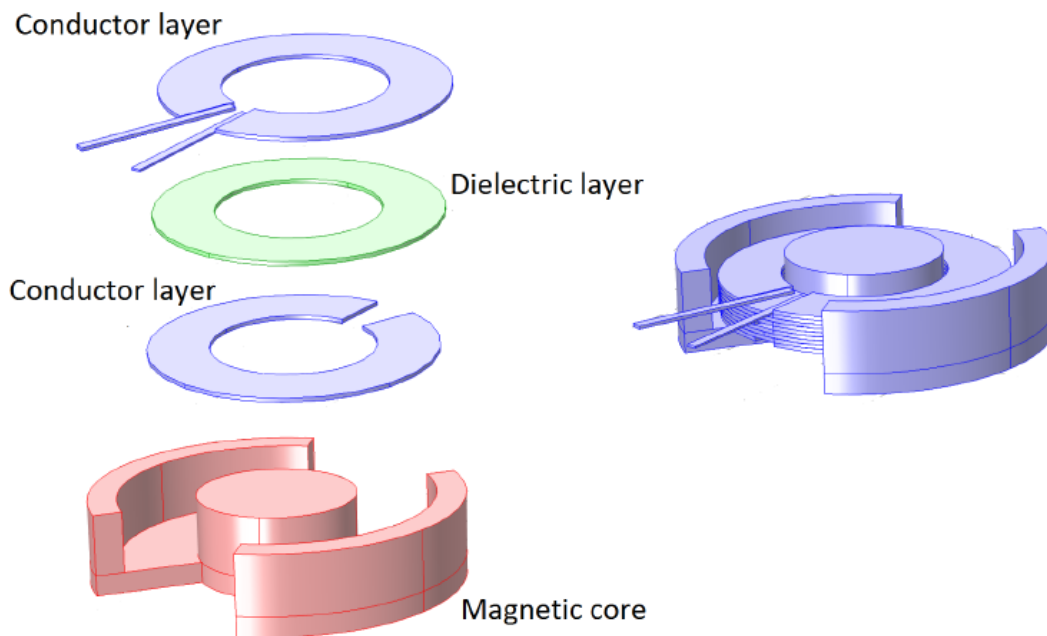


Figure 2.13: Coil structure and connection method of the ring-shaped multiple layer capacitor [72].

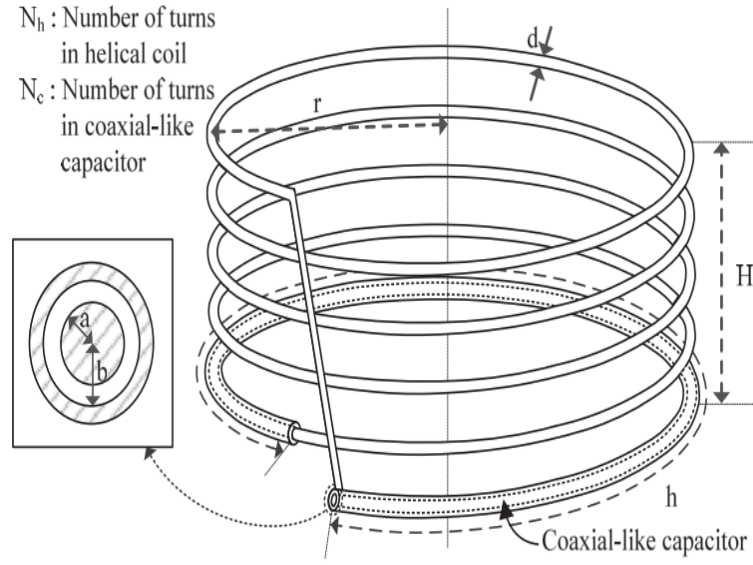
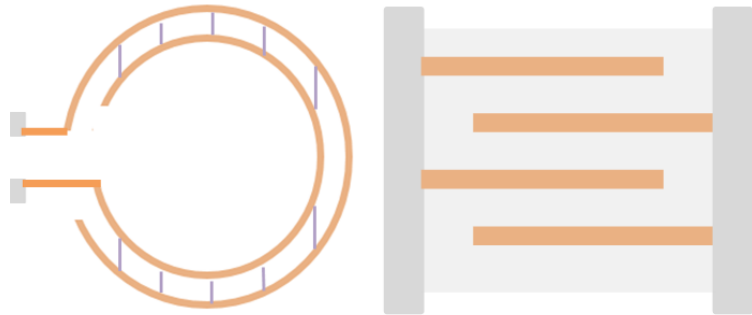


Figure 2.14: Coil structure and connection method of the coaxial cable coil with auxiliary conductor layer [70].



(a)



(b)

Figure 2.15: Coil prototype of the series LC impedance coil (a) [79], and coil structure similar to a multiple layer capacitor but with blended conductor layer (b).

of inductance. Sufficient capacitance is required to lower the resonant frequency, leading to a problematic height for consumer electronics applications. In addition, this reference provides no ESR or inductance modeling or design method for the reader to explore the limit of coil performance.

State-of-art Coil Performance

Integrating the aforementioned coil design technology, the coil prototypes' performance is summarized in Table 2.1. The key parameters, including the working frequency, coil wire, quality factor Q , quality factor including resonant capacitor ESR Q_c , diameter D , and inductance L are listed in the table.

In the kHz range, the Litz structure is widely used in WPT prototypes and achieves high Q . When frequency is near 6.78 MHz, solid wire or copper foil is widely used for high- Q coil design. Compared to conventional coils with similar diameters, SR coil using copper foil and low-loss dielectric typically have higher Q_c . However, reference [50] uses a multi-layer capacitor-like structure for high quality, at the cost of low inductance. The low inductance needs a magnified amount of current to generate similar induced voltage compared to other transmitter coils, which increases the current stress of the front-end power stage. Compared to discrete lumped capacitors, the SR coil still requires a low-loss dielectric to reduce power loss.

2.4.2 ZVS Inverter Design

Power loss in the inverters consists of device conduction loss, device switching loss, and passive component loss. In transitioning from kHz Qi to 6.78 MHz Airfuel, device switching loss increases by over one order of magnitude because switching loss scales with operating frequency. In consumer electronics applications where turn-off loss is small due to the low current low voltage application, turn-on loss dominates the switching loss and zero-voltage-switching (ZVS) operation of all devices is a primary concern [54].

ZVS techniques are widely applied in reported WPT systems [52, 53, 47, 54, 58, 34]. The fundamental principle is designing a current that lags the turn-on of the device. The lagging

Table 2.1: Coil quality factor review.

Reference	Frequency [MHz]	Coil Wire	Q	Q_c	D [cm]	L [μH]
[27]	1.0	Litz, tube	380	254	41	27.2
[54]	0.917	Litz, helical	620	382	30	75.3
[68]	7.64	Solid wire	303	232	59	39
[33]	6.00	Solid wire	90	83	6.8	8.18
[32]	6.78	Foil	201	168	23.5	1.42
[59]	6.78	Foil	178	151	11.3	3.34
[50]	7.73	Foil, SR	-	838	2.1	0.13
[56]	6.78	Foil, SR	-	188	10	3.13

current flows through and fully discharges the output capacitance C_{oss} before a gate-source voltage is applied to turn on the device, eliminating turn-on loss. Reference [52, 53] use load current to discharge the C_{oss} energy, however, in WPT applications the load may change due to position or power variations and may cause ZVS failure due to the change of current phase or amplitude. To ensure ZVS over wide load range, reference [47, 54] adds an auxiliary inductive tank across the switch node and uses the inductive current in the tank to achieve ZVS. The topology, re-produced from reference [54], is shown in Fig. 2.16.

The traditional modeling of Fig. 2.16 covers a simple situation that only the ZVS tank current transfers C_{oss} energy [47, 54]. When the load current is much smaller than the ZVS current [47], or the load is resistive [54], the load current has little contribution compared with ZVS current and thus is ignored during the ZVS transient.

In WPT applications, however, the load of the inverter may be an IMN, which is capable of offering capacitive or inductive load phase with comparable amplitude to the tank current, based on design. A ZVS inverter modeling and design method considering the ZVS tank and IMN together is missing in the literature.

2.4.3 MHz Diode Rectifier Reactive Power Issue

In relatively low frequency and high power applications, diode reactive power is generally ignored due to the negligibly small portion of the transient time. The diode rectifier is modeled as a resistive load and matches well with experiment results [54, 52, 81, 16, 53].

In transition from previous kHz WPT to the target 6.78 MHz WPT, the magnified switching frequency causes resonant transition time to constitute a significant portion of the switching period. Because the resonant transition intervals are expected to be significant in the proposed application, the diode rectifier exhibits a strong capacitive impedance, as shown in Fig. 2.17a from an example WPT system [37]. The reactive power causes a significant efficiency drop when not compensated properly, as shown in Fig. 2.17b [37].

Wide research into diode reactance modeling is reported in references [37, 15, 81, 23]. The reactance is modeled in [37, 15] using a simulation-based method, where a couple of different output voltages and powers are simulated and the reactance modeling is derived through curve fitting of the input characteristics. However, the accuracy of the simulation-assisted

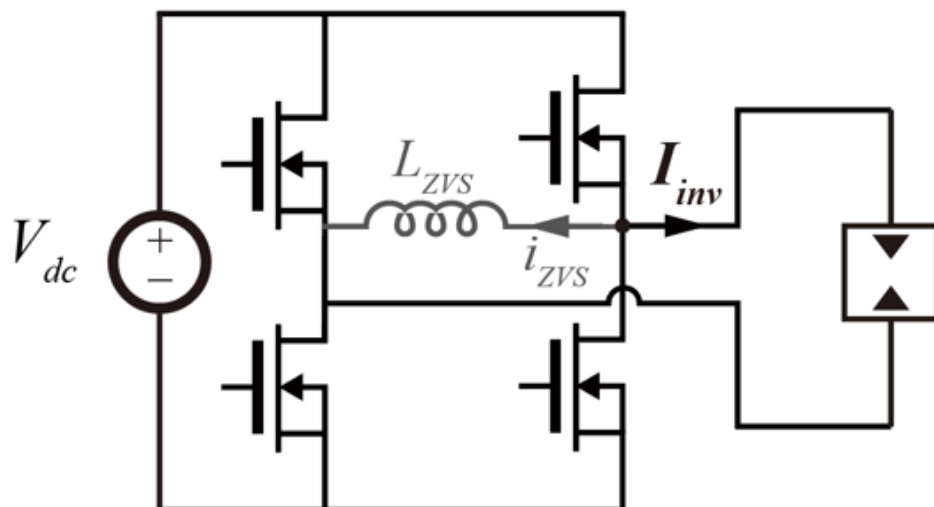


Figure 2.16: ZVS inverter with auxiliary tank [54].

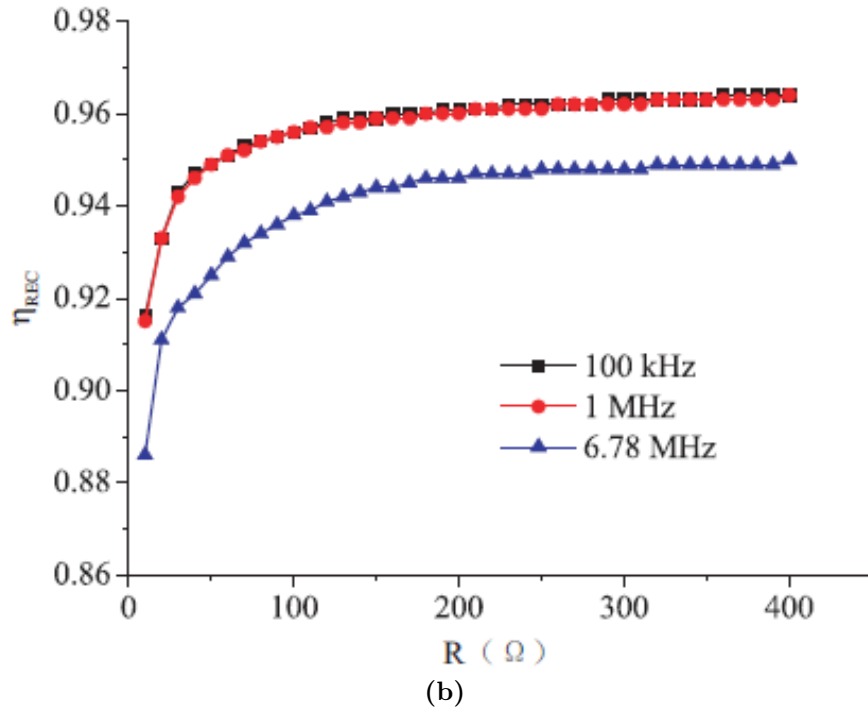
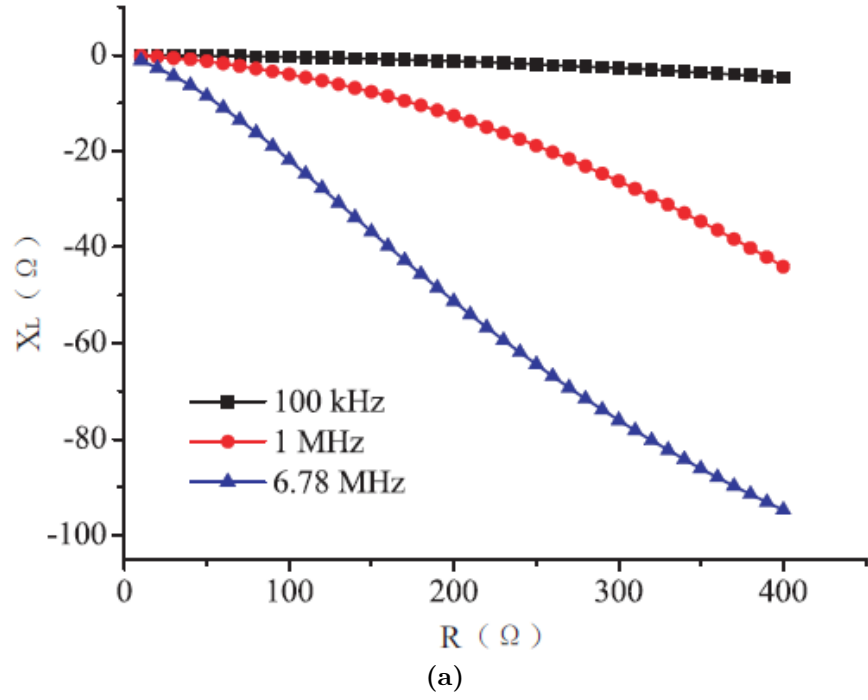


Figure 2.17: Diode reactance of a example WPT system (a) and system efficiency (b) at different frequencies [37] .

curve fitting model depends on the number of simulations. And it cannot cover arbitrary output. A diode model that covers a wide load range accurately is missing in the literature.

2.4.4 State-of-art WPT System Efficiency

The end-to-end efficiency of the state-of-art reported WPT systems are summarized in Table. 2.2. Their power levels are in the range of tens of watts and the operating frequencies are in the range of MHz. The key parameters, including operating frequency f_s , coil diameter D_c , power transfer distance d , coupling coefficient k , output power P_o , and overall efficiency η are listed in the table.

The normalized distance is defined as d/D_c . The normalized distance is considered because it is related to the coupling which is a major limiting factor in the efficiency.

Generally, 6.78 MHz systems working with tens of watts power level have an efficiency lower than 88 %. Research into WPT system design for efficiency that can approach or exceed 90 % is still needed.

2.5 Summary

The call for high efficiency and long transfer distance drives the WPT system into 6.78 MHz operating frequency presenting great challenges to WPT system designs. Motivated by the need for a WPT system that exhibits free-positioning, free-loading, and high efficiency, this chapter reviews the state-of-art literature for the target characteristics.

Previous efforts of achieving free-positioning are classified into three categories: 1) geometric optimization design for conventional rectangular or circular coils, which may have standing or radiation issue for MHz WPT covering a large charging area; 2) coil array, which exhibits low-Q due to limited coil size; and 3) current shaping using multiple controlled power converters, which requires duplicated converter hardware, and complicated sensor and controller design. For a WPT table covering a large area, previous literature lacks the coil design to offer a uniform field and no standing wave issue. The new coil structure and design method are waiting.

Table 2.2: WPT system end-to-end efficiency review.

#	Reference	f_s [MHz]	D_c [cm]	d	d/D_c	k	P_o [W]	η [%]
1	[68]	7.65	59 (28)	70	1.61	—	12	50
2	[32]	6.78	23.5	5	0.21	—	24	73
3	[59]	6.78	11.3	4	0.38	—	10	72
4	[35]	6.78	11.3	5	0.44	0.18	10	80
5	[77]	6.78	10	3	0.3	-	50	82
6	[13]	6.78	17.5	11	0.62	0.11	50	88
7	[36]	13.56	20	10	0.50	0.075	40	76
8	[74]	13.56	7.6	1.5	0.20	-	300	88
9	[54]	0.91	30	20	0.68	0.08	50	82

Previous efforts of achieving free-loading are classified into two categories: 1) transmitter side converter or IMN design to achieve constant coil current, and 2) receiver side converter design to maintain constant rectified voltage. The converter design methods in 1) and 2) require complicated sensors and control strategies, and the additional DC/DC converter will add system loss. The IMN in 1) may exhibit high power loss due to improper design in a multi-stage WPT system. Thus, the IMN design method at the system level will be explored to offer constant output current while exhibiting small power loss, which is missing in the literature.

Previous efforts of achieving high efficiency are guided by the fundamental circuit in Fig. 2.1. Based on the circuit, a sequential design methodology is widely adopted that first designs coils to achieve high-Q, then designs the converters for an operation that presents the optimal load described by (2.1). Previous efforts of achieving high-Q coils are 1) wire selection for less eddy current and more conduction area at the system working frequency; 2) model-based coil geometric optimization for maximum Q; and 3) SR coil to reduce dielectric loss. In this dissertation, both 1) and 2) will be used in the coil design, and NP0 [4] capacitors will be used to reduce dielectric loss. In a 6.78 MHz system, the converter cannot be designed solely to present the optimal load as in (2.1). ZVS must be considered to avoid excessive switching loss and system failure. In previous systems, ZVS is achieved via the inductive current from the load, or via the auxiliary inductor which offers constant inductive current regardless of the load current. However, the ZVS design together with IMN is missing. The IMN also can provide an inductive current which may change with the auxiliary inductor design. For diode rectifiers, the analytical reactance and loss modeling at 6.78 MHz is missing in the literature.

Chapter 3

Proposed Multi-load WPT System and Design Methodology

The proposed application was shown in Fig. 1.3b and is re-plotted here in Fig. 3.1. In the application, multiple receivers with coil sizes substantially smaller than the transmitter may be placed arbitrarily within a charging surface parallel to the transmitter surface but a fixed distance away. This arrangement models applications such as a wireless charging desk or table.

A schematic of the proposed application is shown in Fig. 3.1. The transmitter side consists of a ZVS inverter, a uniform field transmitter coil, and an IMN. The receiver side consists of a receiver coil, a diode rectifier, and a capacitor that compensates for the receiver reactance. The receiver side has two identical receivers in the analysis. The two sides are coupled by the magnetic field B_0 , which is near-constant due to the transmitter coil design. The proposed WPT system is desired to offer free-positioning, free-loading, and high efficiency. The free-positioning is realized by the new interleaved transmitter coil that has small field variation and no standing wave issue. Detailed modeling and design method will be given in Chapter 4. The free-loading is realized by the IMN that maintains a constant coil current. The IMN design in a WPT system is given in Chapter 5. The inverter achieves ZVS using an auxiliary inductor and IMN simultaneously. Detailed modeling and design will be given in Chapter 5. The analytical modeling and design of the receiver side will be given in Chapter 6.

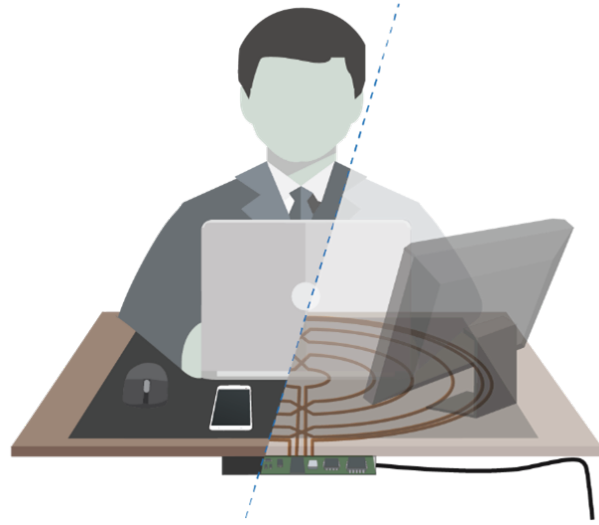


Figure 3.1: Schematic of the target application, where the transmitter coil is placed on the bottom side of a large table.

At 6.78 MHz, the inverter ZVS operation and diode reactance are coupled with their output voltages and currents. Thus, the sequential design that maximizes coil Q and then designs the converter operation for the optimal load cannot be used. Due to the system complexity, direct design optimization of the entire system may be infeasible. Before proceeding to the system-level design, each of the stages, corresponding to the dashed lines in Fig. 3.2, are examined individually, as shown in Fig. 3.3. The stages are decoupled from one another using fundamental harmonic analysis. Though each stage is now able to be analyzed individually, the operation of each is coupled through the highlighted input/output variables in Fig. 3.3. Thus, e.g. the transmitter coil can be optimized for a given combination of coil current and average flux density, $[I_t, B_0]$, the final selection of these two parameters requires system-level consideration as they affect the performance of adjacent stages.

In this dissertation, each stage is designed first in isolation, resulting in an optimized design that is parameterized by the highlighted coupling variables of Fig. 3.3. Then, each of these parameterized designs is combined in a system-level optimization to select the optimal value of the coupling variables for the system-level performance, as detailed in Chapter 7.

Specifications of Target Application

The specifications of the multi-load charging station include the total output power $P_{o,all}$ needed to charge the devices, the table thickness h across which the power is transmitted, and the geometric size for the transmitter and receiver coils. In the application, the table thickness is measured to be 1.7 cm. The size of the receivers is designed to fit the base of a computer monitor with a size of 8.5x5 inch². To offer positional charging flexibility, the B -field must have minimal variation over the charging area for a near-constant induced voltage independent of the receiver position. The system operates at 6.78 MHz frequency. The maximum B -field above the table is limited to 27 μ T according to ICNIRP 2010 [44]. In this paper, the input DC voltage is assumed to be 200 V to work with a front-end PFC [73]. The Rx side maximum output ac voltage is 40 V to work with the post battery charger ICs. This corresponds to a dc voltage limit of $V_o < (\pi/4)40 = 31.5$.

By adding a post dc-dc conversion stage, the range of the output voltage can be expanded at the cost of complexity and efficiency. The post dc-dc stage also has an reduced efficiency

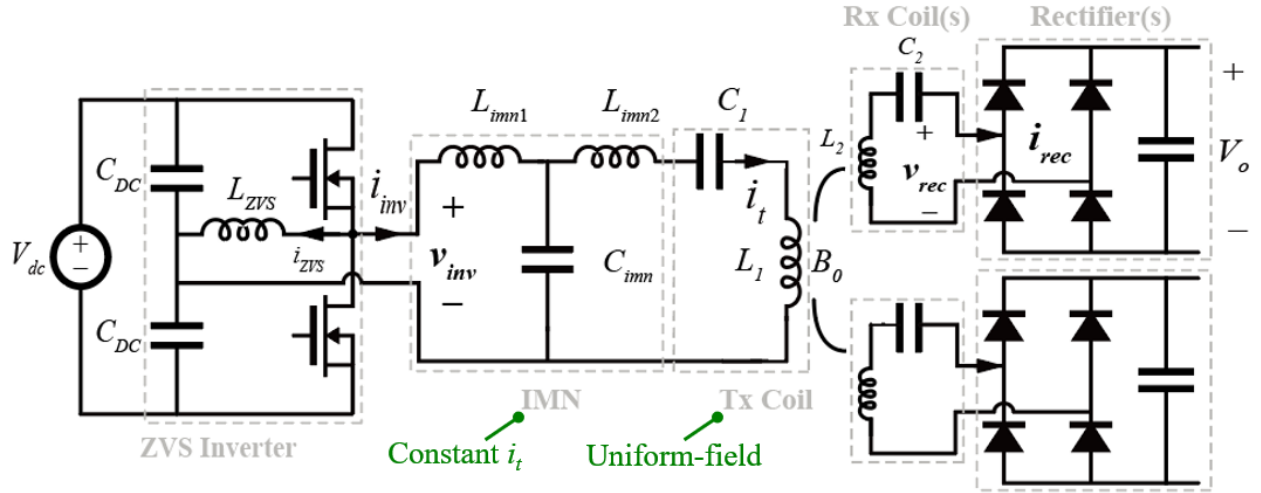


Figure 3.2: Circuit schematic of the multi-load WPT system, including inverter, IMN, transmitter and receiver coils, rectifiers, and DC output.

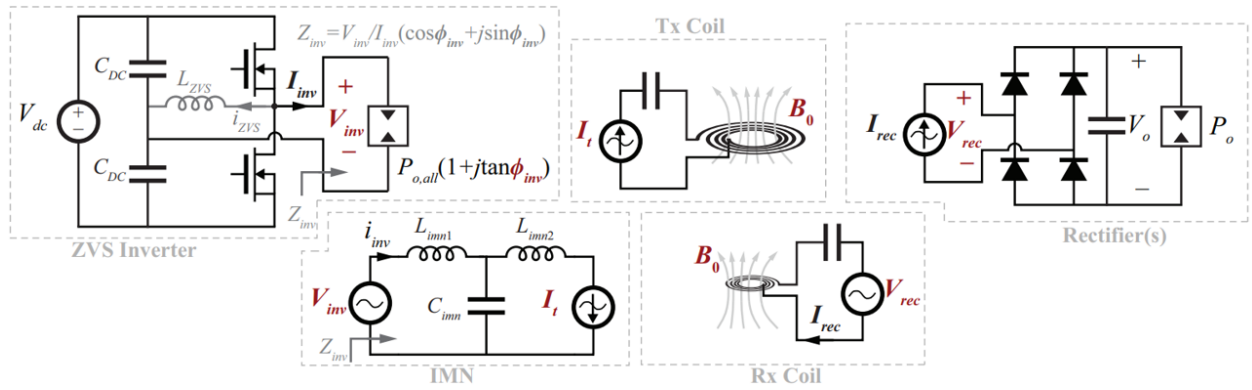


Figure 3.3: Decoupled circuit model showing the parameters that couple adjacent stages and the parameters inside an individual stage.

with higher input voltage. Thus, this dissertation uses 31.5 V as the output voltage limit for simplicity. The specifications of the desired WPT station is summarized in Table [3.1](#).

Table 3.1: Specifications of the desired WPT station.

Parameters	Variable	Value
Total output power	$P_{o,all}$	100 W
Table thickness	h	1.7 cm
Size of Tx coil	-	50x50 cm ²
Size of Rx coils	-	10.8x6.4 cm ²
Magnetic field variation	κ	30 %
System frequency	f_s	6.78 MHz
Input voltage	V_{dc}	200 V
Output voltage limit	V_o	31.5 V

Chapter 4

Analysis and Design of Transmitter Coil

4.1 Structure of the Interleaved Tx Coil

The main principle of the interleaved transmitter coil is to use paralleled windings to increase the coverage of the winding area without requiring a higher number of turns, as illustrated in Fig. 4.1. Fig. 4.1a shows a conventional 3-turn, symmetric coil with the innermost turn highlighted. The current in each of the three turns is identical and equal to the input current amplitude I_t , $i_t = [1, 1, 1]I_t$. In Fig. 4.1b, this innermost turn has been split into two parallel windings. To facilitate equal current sharing in each parallel turn, each parallel turn is swapped at the halfway point to form a symmetric structure and equalized length and impedance, shown at the top of the figure. If the impedances are perfectly balanced, the currents in each turn are now $i_t = [0.5, 0.5, 1, 1]I_t$, from innermost to outermost. In Fig. 4.1c, the innermost paralleled turn from Fig. 4.1b is further split into two parallel and impedance-balanced turns. If the impact of the crossover points is negligible, the resulting coil magnetic field is equivalent to a 5-turn coil, with currents $i_t = [0.25, 0.25, 0.5, 1, 1]I_t$ in each turn, but each complete current loop through the coil encompasses only 3 turns. This coil is described as having $n = 5$ physical turns and $N_{eq} = 3$ effective turns.

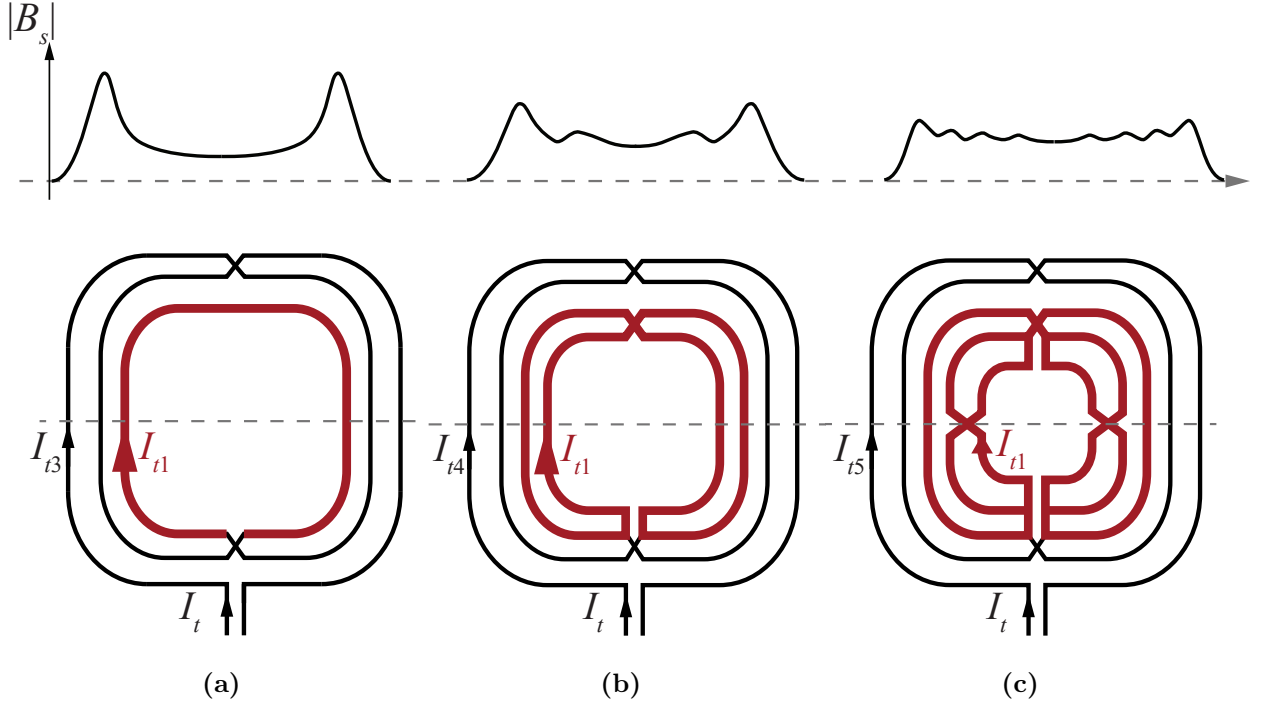


Figure 4.1: Conventional 3-turn coil (a) with innermost turn highlighted; (b) modified 3-turn coil with two parallel windings for the innermost turn; and (c) modified 3-turn coil where the innermost parallel winding has been further split into two parallel windings. Above each coil is a qualitative depiction of the B -field produced by the coil along the dashed line, some distance above the coil

The qualitative B -field of each coil at a receiver surface some distance away is shown at the top of Fig. 4.1. As the winding effective turns are split into a higher number of physical turns covering the coil area, the field can be designed with greater uniformity.

To model the interleaved coil, both the geometric parameters and current distribution are necessary. The half-length, i.e. the radius of the maximal circle that can be circumscribed within the square turn, of each turn is $a_{tx} = [a_1 \dots a_n]$, r_0 is the wire radius, and the current ratio $\lambda = [I_{t1} \dots I_{tn}]/I_t$ describes the interleaving method.

4.2 Magnetic Field Analysis

The target charging surface has a distance h above from the transmitter and covers 90% of the extents of the transmitter coil. A mutual inductance-based method is used to calculate the B -field on the charging surface and to assess the field uniformity of each candidate coil design. As shown in Fig. 4.2, n_2 centrally-located square loop with half-length of $b_1 \dots b_{n_2}$ are evenly distributed over the charging surface. The magnetic flux in the i^{th} loop is

$$\Phi_i = \sum_{j=1}^n \lambda_j M_{j,i} I_t \quad (4.1)$$

where $M_{j,i}$ is the mutual inductance between the i^{th} loop and the j^{th} turn of the transmitter coil [22],

$$\begin{aligned} M_{j,i} = \frac{\mu_0}{2\pi} & \left[2\sqrt{(a_j + b_i)^2 + (a_j + b_i)^2 + h^2} \right. \\ & - 2\sqrt{(a_j + b_i)^2 + (a_j - b_i)^2 + h^2} \\ & - (a_j + b_i) \arctan \left(\frac{a_j + b_i}{\sqrt{(a_j + b_i)^2 + (a_j + b_i)^2 + h^2}} \right) \\ & + (a_j - b_i) \arctan \left(\frac{a_j - b_i}{\sqrt{(a_j + b_i)^2 + (a_j - b_i)^2 + h^2}} \right) \\ & + (a_j + b_i) \arctan \left(\frac{a_j + b_i}{\sqrt{(a_j + b_i)^2 + (a_j - b_i)^2 + h^2}} \right) \\ & \left. - (a_j - b_i) \arctan \left(\frac{a_j - b_i}{\sqrt{(a_j - b_i)^2 + (a_j - b_i)^2 + h^2}} \right) \right] \end{aligned} \quad (4.2)$$

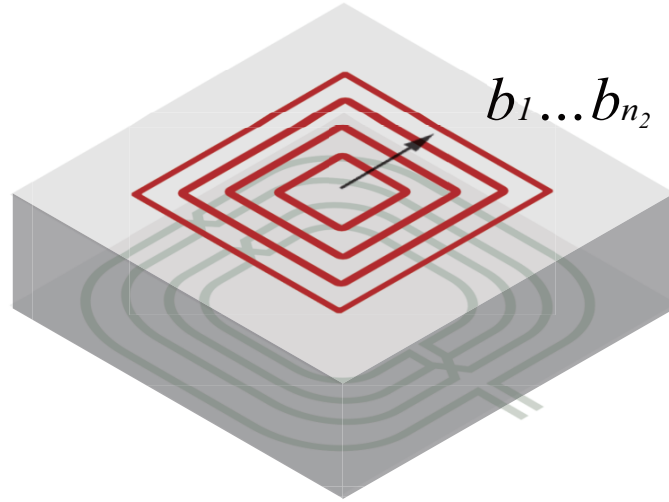


Figure 4.2: Schematic showing the transmitter coil and the centrally-located square on the charging surface for evaluating mutual inductance.

Thus, the magnetic flux in the strip region between i^{th} and $(i + 1)^{\text{th}}$ loop is $\Phi_{strip,i} = \Phi_{i+1} - \Phi_i$. The averaged B -field in the strip region is

$$B_{strip,i} = \frac{\Phi_{i+1} - \Phi_i}{4(b_{i+1}^2 - b_i^2)} \quad (4.3)$$

When $(b_{i+1} - b_i)$ is small, $B_{strip,i}$ is approximately equal to the point-by-point B -field along the arrow shown in Fig. 4.2. The qualitative B -field variation is

$$\kappa = \frac{\max_i(B_{strip,i}) - \min_i(B_{strip,i})}{\frac{1}{n_2} \sum_i B_{strip,i}} \quad (4.4)$$

The averaged B -field on the charging region is

$$B_0 = \sum_{j=1}^n \lambda_j M_{j,n_2} I_t \quad (4.5)$$

The averaged B -field per unit input current is

$$\psi_0 = \sum_{j=1}^n \lambda_j M_{j,n_2} \quad (4.6)$$

Thus, the required coil current is $I_t = B_0/\psi_0$.

With the model in this section, for each coil design with specified $[a_{tx}, i_t, N_{eq}]$, its field variation κ can be calculated using (4.4), and the required current I_t to generate target B_0 can be calculated using (4.6). I_t will be used to calculate coil power loss in section 4.5.

4.3 Circuit Parameters LR Modeling

The inductance of the Tx coil is different from conventional coils due to the possible interleaved connection and various current distribution. The self-inductance of m^{th} physical turn, when it conducts the same current as input current I_t , is [63]

$$L_{mm} = 0.002 l_m \left(\ln \left(\frac{l_m}{r_0} \right) - 1 + 0.25 \mu_0 + \frac{r_0}{l_m} \right) \quad (4.7)$$

where l_m is the perimeter of the m^{th} physical turn.

The mutual inductance between m^{th} and j^{th} physical turn M_{jm} is calculated from (4.2), replacing b_i with a_m .

For i^{th} effective turn, when it has interleaved sub-turns, including the $s_1^{\text{th}} \dots s_n^{\text{th}}$ physical turns, its inductance including the mutual inductance with other turns is

$$L_{s,i} = \sum_{m=s_1}^{s_n} \left(\lambda_m^2 L_{mm} + \sum_{j=1, j \neq m}^n \lambda_j \lambda_m M_{jm} \right) \quad (4.8)$$

where M_{jm} is the mutual inductance between the m^{th} and j^{th} physical turn. M_{jm} is calculated from (4.2), replacing b_i with a_m . Note that the current distribution plays an important role in calculating the inductance values.

When i^{th} effective turn has no interleaved sub-turns, and only corresponds to the m^{th} physical turn, (4.11) is reduced to

$$L_{s,i} = \left(\lambda_m^2 L_{mm} + \sum_{j=1, j \neq m}^n \lambda_j \lambda_m M_{jm} \right) \quad (4.9)$$

The total inductance is

$$L_1 = \sum_{i=1}^{N_{eq}} (L_{s,i}) \quad (4.10)$$

4.4 Electric Field Compensation

In addition to a uniform magnetic field at the receiver surface, it is desirable to have a minimal electric field generated by the coil. The electric field has the potential to generate an additional dielectric loss in the material that the coil is mounted within (e.g. a desk or table surface), and may cause EMI or safety concerns for objects placed on the charging surface. The electric field is generated due to the longitudinal voltage drop across the coil and may be reduced through distributing the series compensation capacitor into multiple discrete parts over the length of the coil. Though the total longitudinal voltage drop is unchanged, the electric potential between any point on the coil and an external reference (for example a dc rail of the transmitter voltage supply) is reduced over the length of the coil. The strength

of the field is related to the coil inductance and current. And the induced dielectric loss is related to the volume and loss factor of the dielectric base. In this section, it is assumed that adding two discrete capacitors for each equivalent turn is sufficient to compensate for the electric field. One capacitor is placed at the middle of the equivalent turn to compensate for the first half of the longitudinal voltage drop and the other is placed at the end to compensate for the other half. An impedance-based method is used to calculate the voltage drop of each effective turn and the discrete capacitance value which fully compensates for the voltage drop.

For the i^{th} effective turn, when it has interleaved sub-turns, including the $s_1^{\text{th}} \dots s_n^{\text{th}}$ physical turns, its inductance including the mutual inductance with other turns is

$$L_{s,i} = \sum_{m=s_1}^{s_n} \left(\lambda_m^2 L_{mm} + \sum_{j=1, j \neq m}^n \lambda_j \lambda_m M_{jm} \right) \quad (4.11)$$

where M_{jm} is the mutual inductance between the m^{th} and j^{th} physical turn. M_{jm} is calculated from (4.2), replacing b_i with a_m . Note that the current distribution plays an important role in calculating the inductance values. L_{mm} is the self-inductance of m^{th} physical turn, when it conducts the same current as input current I_t , is [63]

$$L_{mm} = 0.002 l_m \left(\ln \left(\frac{l_m}{r_0} \right) - 1 + 0.25 \mu_0 + \frac{r_0}{l_m} \right) \quad (4.12)$$

where l_m is the perimeter of the m^{th} physical turn.

When i^{th} effective turn has no interleaved sub-turns, and only corresponds to the m^{th} physical turn, (4.11) is reduced to

$$L_{s,i} = \left(\lambda_m^2 L_{mm} + \sum_{j=1, j \neq m}^n \lambda_j \lambda_m M_{jm} \right) \quad (4.13)$$

The voltage of i^{th} effective turn is

$$V_{tx,i} = \omega_s I_t L_{s,i} \quad (4.14)$$

where ω is the angular frequency.

The total inductance is

$$L_1 = \sum_{i=1}^{N_{eq}} (L_{s,i}) \quad (4.15)$$

To compensate for the longitudinal voltage change of i^{th} equivalent turn, the capacitance value of the two identical discrete capacitors are

$$C_{t,i} = 2I_t / (V_{tx,i}\omega_s) \quad (4.16)$$

The equivalent capacitance of all the distributed compensation capacitors is $C_1 = 1 / (2 \sum_{i=1}^{N_{eq}} (1/C_{t,i}))$.

An example $N_{eq} = 4$, $\lambda = [0.25 \ 0.25 \ 0.5 \ 0.5 \ 0.5 \ 1 \ 1]$, $a_{tx} = [8.1 \ 12.6 \ 15.9 \ 19.3 \ 22.6 \ 23.8 \ 25]$ cm Tx coil is shown in Fig. 4.3 to explain the discrete compensation. The capacitors are placed at the end of each half turn, which are marked by $C_{a1}, C_{b1} \dots C_{b3}$. The calculated longitudinal electric potential along the length of the coil with and without distributed capacitor compensation is compared in Fig. 4.4, where m is the middle point of the symmetric coil. Clearly, the magnitude of the electric field is reduced significantly. The reduced electric field will have a weaker interference with neighboring objects, such as the table dielectric that holds the coil. The significant reduction of the electric field and the associated dielectric loss has been verified by FEA simulation, which will be presented in section 4.7.

4.5 Power Loss Modeling

With electric field compensation, the power loss in the dielectric base is assumed to be nearly zero. The remaining power loss in the transmitter coil consists of copper loss and compensation capacitor loss. The copper ESR is determined by both coil geometry and the interleaving method

$$R_{tx,copper} = \sum_{i=1}^n \frac{8\lambda_i^2 \rho a_i}{\pi(r_0^2 - (r_0 - \delta^2))} \quad (4.17)$$

where ρ is copper resistivity, and $\delta = 1/\sqrt{\mu_0 \rho f_s} \ll r_0$ is the skin depth at switching frequency f_s .

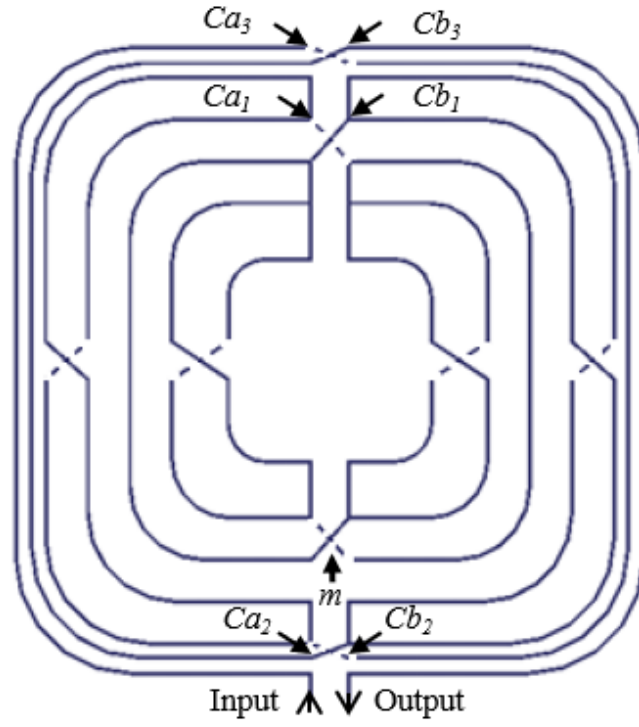


Figure 4.3: Proposed coil and the discrete compensation capacitors.

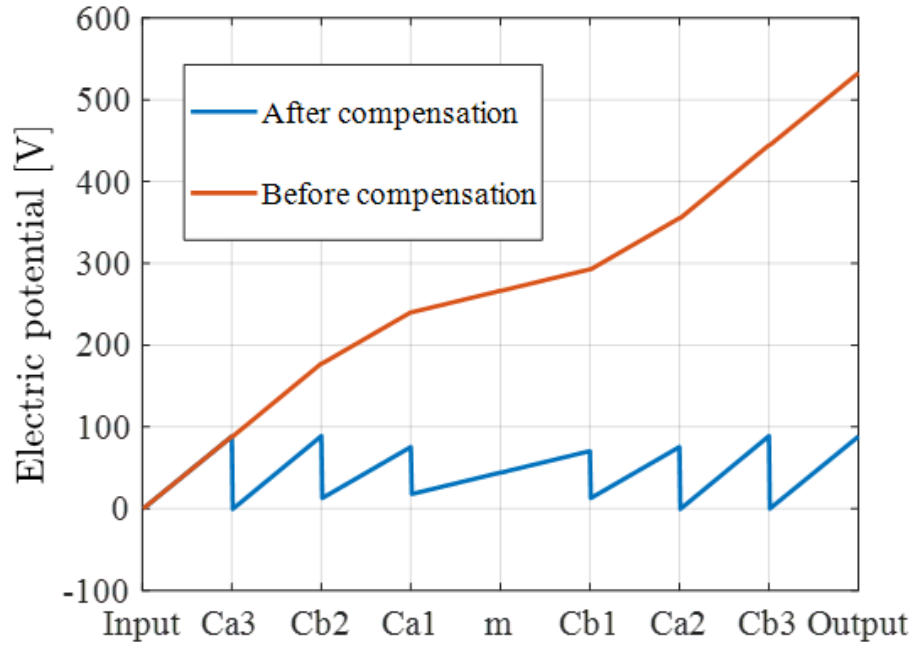


Figure 4.4: Calculated longitudinal voltage change.

The total compensation capacitors ESR is estimated by

$$R_{tx,cap} = 2 \sum_{i=1}^{N_{eq}} \frac{t_g}{\omega_s C_{t,i}} \quad (4.18)$$

where t_g is the loss tangent of the capacitor dielectric material indicated in the datasheets by the manufacturer[4]. In high-frequency applications, NP0 capacitors are used for small ESR, and low capacitance variations regardless of temperature change [4].

Combining copper and compensation capacitor ESR, the total ESR is $R_1 = R_{tx,copper} + R_{tx,cap}$. And the power loss in the transmitter coil is

$$P_{loss,tx} = 0.5 I_1^2 \left(\sum_{i=1}^n \frac{8 \lambda_i^2 \rho a_i}{\pi (r_0^2 - (r_0 - \delta^2))} + 2 \sum_{i=1}^{N_{eq}} \frac{t_g}{\omega_s C_i} \right) \quad (4.19)$$

4.6 Transmitter Coil Design

As stated at the beginning, the transmitter coil will be parameterized using the load of adjacent stages $[I_1, B_0]$ for system-level design. B_0 influences the Rx side through involvement in the induced voltage design, which further influences the reactance compensation and the rectifier design. I_1 on the other hand is coupled with the IMN output current design. The goal of the Tx coil design in isolation is to meet the field variation requirement and achieve minimum power loss for a range of $[I_1, B_0]$, through designing coil physical parameters, e.g. the number of turns N_{eq} , turn radii a_{tx} and paralleling scheme λ .

In the design process, N_{eq} , a_{tx} and λ are first iterated to calculate $[\kappa, \psi_0, R_1]$. Designs with κ larger than application requirements are discarded. To generate a target B_0 , the required I_1 is calculated using (4.5) and (4.6), and $P_{loss,tx}$ is calculated afterwards using (4.19). For a specified B_0 , multiple designs may exist that need identical I_1 . Only the design with minimum power loss is saved and all other designs are eliminated. Consequently, the coil design is converted from a geometrical design to a section of B_0 and I_1 . For each combination, the associated minimum loss is stored in a two-dimensional coil design look-up-table. The selection of B_0 and I_1 will be determined at the system-level in Chapter 7 to capture the dependence on adjacent stages.

For the example design shown in Table 3.1, B_0 is iterated from 10 mT to 27 μ T with a step of 1 μ T. I_1 is iterated from 0.5 A to 3 A with a step of 0.1 A. Coil N_{eq} is iterated from 2 to 4. Each effective turn is allowed to be implemented as a single turn, $\lambda = [1]$, two parallel turns, $\lambda = [0.5, 0.5]$, or three turns with $\lambda = [0.25, 0.25, 0.5]$. The possible position for each turn ranges from 1 cm to 25 cm with a 0.5 cm step, except for the outermost turn with a fixed position of 25 cm. Considering the high frequency, AWG 12 solid copper wire is used for small conduction loss. The resulting candidate coils performances $[\kappa, \lambda_0, \text{ and } R_1]$ are shown in Fig. 4.5. The designs that exceed the 30 % maximum limit on κ are plotted with a lighter color and will not be used in further analysis. As N_{eq} increases and provides finer shaping of the field, the field can be designed with greater uniformity.

Within the range of B_0 and I_1 , the designs that have minimum power loss are shown in Fig. 4.6. The optimized designs form two strips because ψ_0 and R_1 are highly related to N_{eq} , as shown in Fig. 4.5.

The parameters of the final Tx coil design, which result from the overall design in Chapter 7, are summarized in Table 4.1.

4.7 FEA Simulation Verification

Using the parameters in Table 4.1, an FEA simulation model is built with Ansys HFSS. A picture of the simulation model is given in Fig. 4.7. The coil is placed under an HDPE plastic table with a thickness of 1.7 mm. HDPE has a low dielectric constant of 2.2 and a low dielectric loss tangent of 0.0002, which makes it ideal for high-frequency applications.

In the simulation, the compensation capacitors are modeled by cutting the copper wire and inserting an NP0 dielectric material [4], forming a parallel-plate capacitor to represent the lumped discrete capacitor. The capacitor dielectric thickness is adjusted to form the target capacitance. The model is shown upside-down so the windings are visible.

4.7.1 Electric Field Verification

The electric field between adjacent turns causes additional power loss in the table. Thus, the dielectric loss density is used to clearly show the electric field.

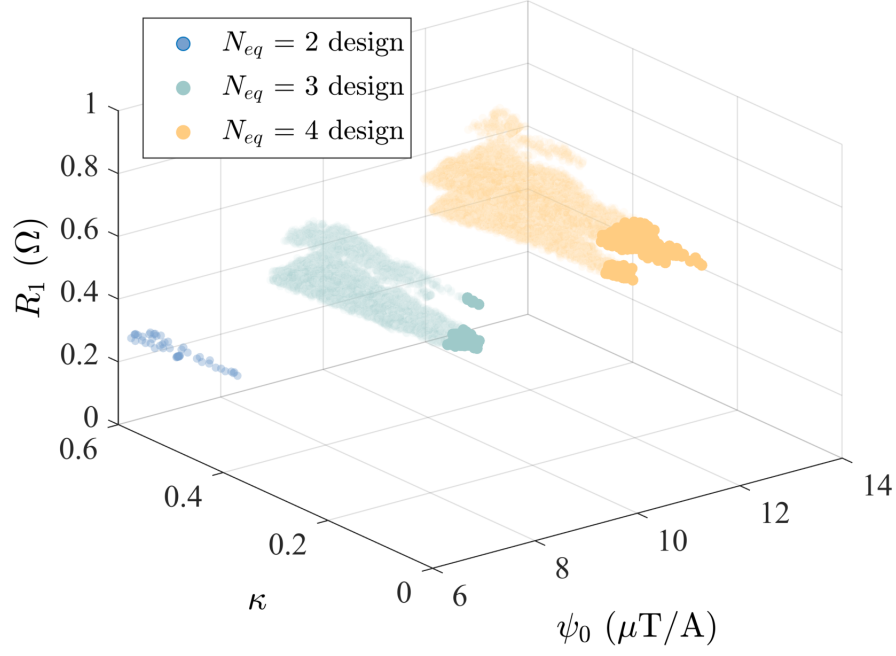


Figure 4.5: Transmitter coil characteristics after iteration. For each N_{eq} , designs shown with lighter color exceed λ limit.

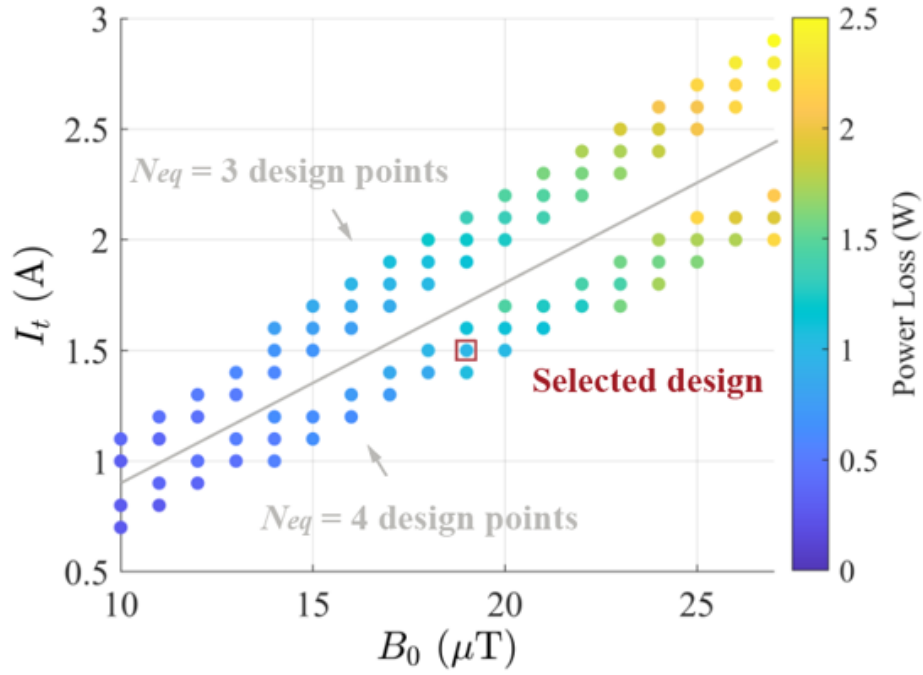


Figure 4.6: Total transmitter coil power loss, including the wire conduction loss and compensation capacitor loss.

Table 4.1: Specifications of the proposed Tx coil design.

Parameters	Variable	Value
Table thickness (cm)	h	1.7
Half length (cm)	a_{tx}	[8.1 12.6 15.9 19.3 22.6 23.8 25]
Current ratio	λ	[0.25 0.25 0.5 0.5 0.5 1 1]
Magnetic field variation	κ	15 %
Inductance (μH)	L_1	12.8
Capacitance (pF)	C_1	44.8
ESR (Ω)	R_1	1.07

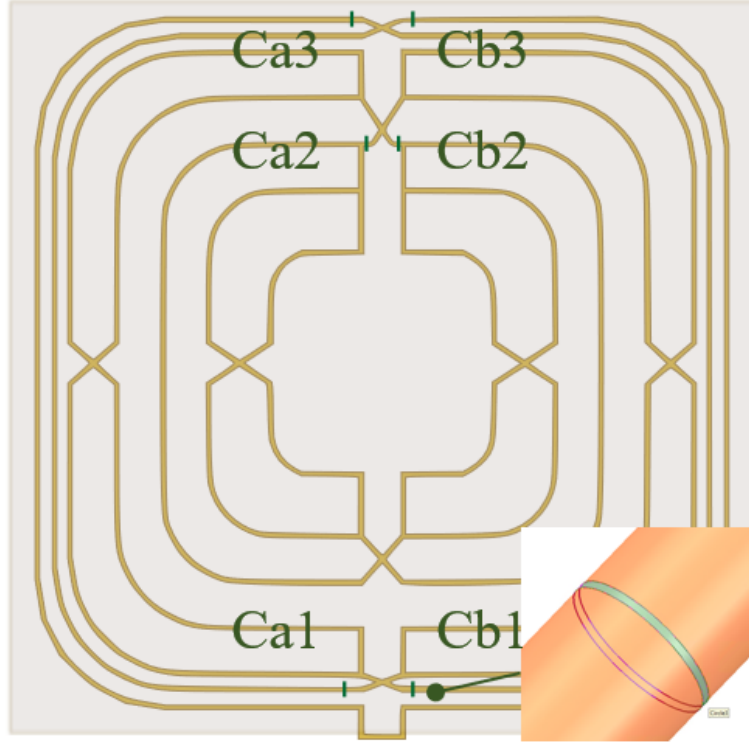


Figure 4.7: Proposed wireless charging coil, shown upside-down for visibility.

As shown in Fig. 4.8, the dielectric loss is almost eliminated after adding the discrete compensation, where both cases are simulated when the coil conducts 1.55 A, 6.78 MHz current. The significant reduction of the electric field verifies that the longitude voltage drop is canceled. The simulated ESR before adding compensation was 1.73Ω and drops to 1.11Ω after compensation, which is close to the calculated result of 1.07Ω .

4.7.2 Magnetic Field Verification

The B -field is simulated and the magnitude of the z -axis component is plotted from the center towards the middle of the edge, along the arrow shown in Fig. 4.10. The simulation step is 2.25 cm and the position of the last measurement is 22.5 cm, which covers 90 % of the width of the coil. The simulated B -field on the charging surface, together with calculations, is shown in Fig. 4.9 when the coil conducts 1.55 A, 6.78 MHz current. The simulation matches the calculation closely, proving the accuracy of the magnetic field modeling.

4.8 Experimental Coil Evaluation

To verify the Tx coil model experimentally, a $0.5 \times 0.5 \text{ m}^2$ coil with the target geometry is fabricated. A picture of the prototype is given in Fig. 4.10, together with other stages of the WPT system which will be introduced in later chapters. The transmitter coil is shown upside-down so the windings are visible.

To precisely control the shape of the coil, the geometries are engraved into sheets of 1.7 cm HDPE dielectric using a CNC mill.

An Agilent 4294A impedance analyzer is used to measure resistance and inductance. The measurement results are summarized in Table 4.2 and compared with the designed parameters, proving the accuracy of the modeling.

To test the uniformity of the transmitter coil magnetic field, a field probe [18] is used to test the magnetic field when the system works with the nominal input voltage and no load is present. The field is measured from the center towards the middle of the edge, along the arrow shown in Fig. 4.10. The measurement step is 2.25 cm and the position of the last measurement is 22.5 cm, which covers 90 % of the width of the coil. The measurements are

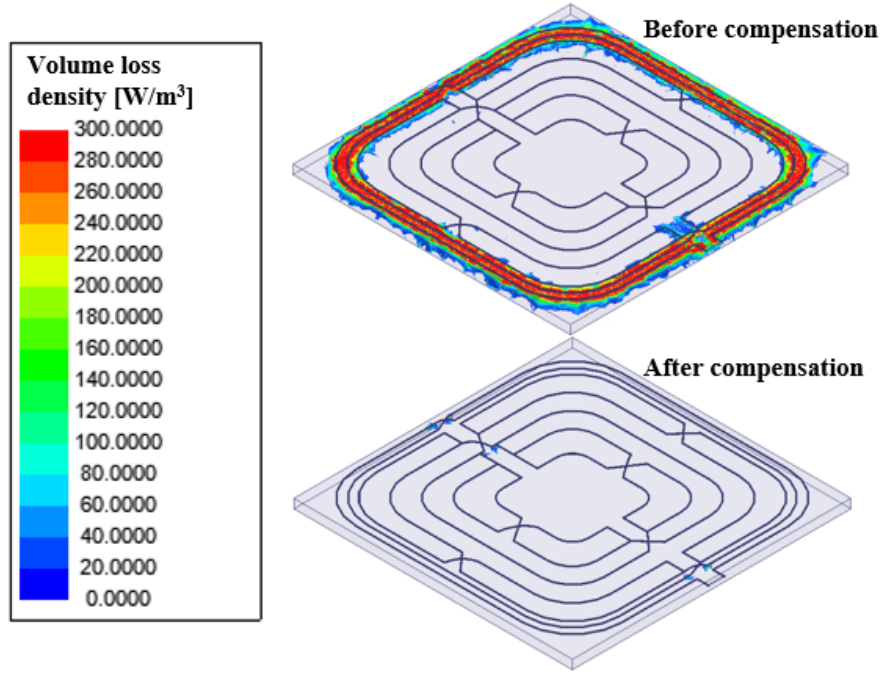


Figure 4.8: FEA simulation showing the dielectric loss density.

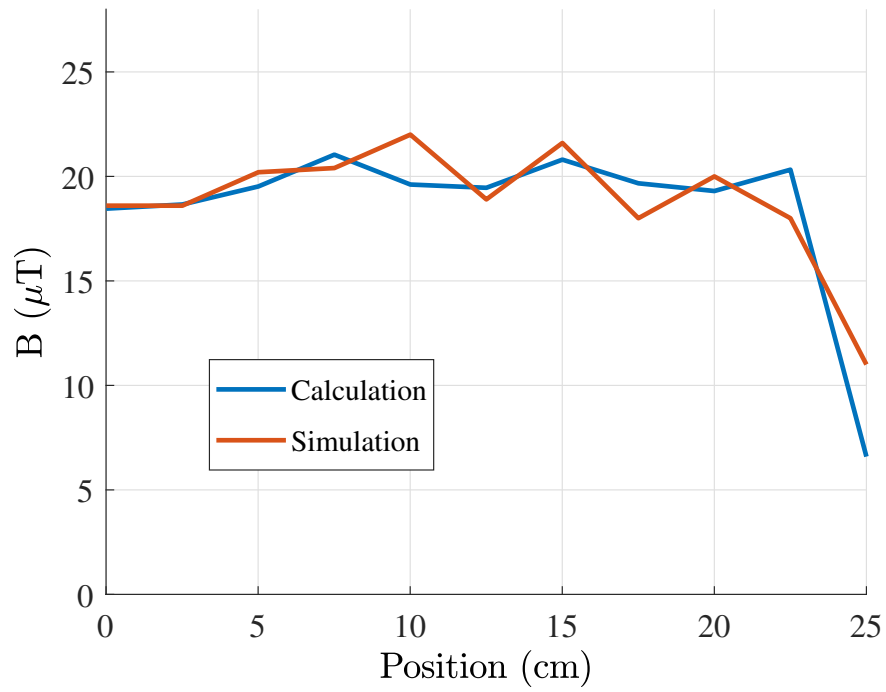


Figure 4.9: Simulated B -field, compared with the calculation.

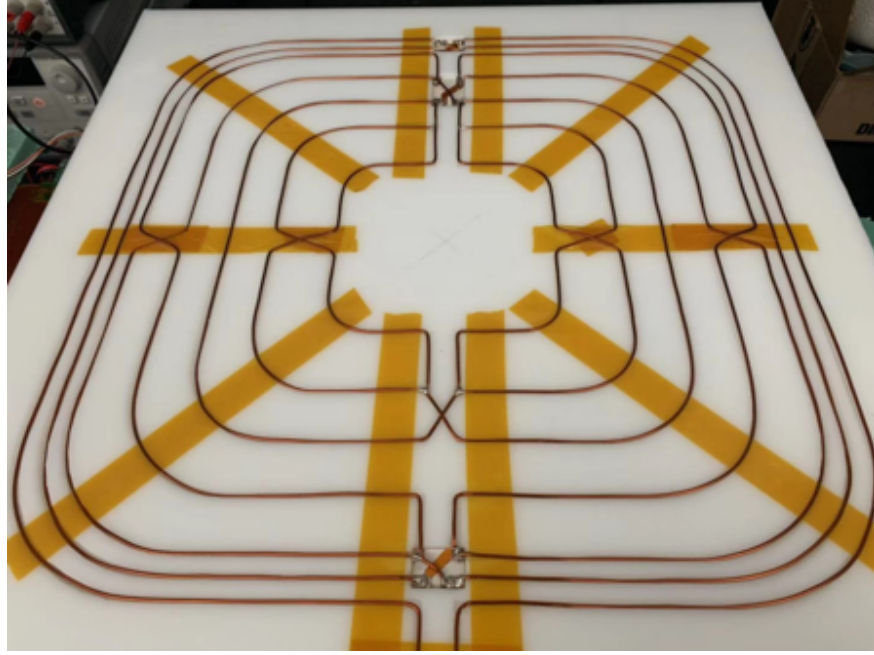


Figure 4.10: Picture of the prototype.

Table 4.2: Comparison of the measured and calculated circuit parameters.

Parameters	Variable	Design	Measurement
Tx coil inductance (μH)	L_1	12.3	12.6 (+2.4%)
Tx coil compensation (pF)	C_1	44.8	44 (-1.8%)
Tx coil ESR (Ω)	R_1	1.07	1 (-6.5%)

shown in Fig. 4.11 together with the model prediction showing that a near-uniform magnetic field is achieved. The measured $B_0 = 20.5 \text{ } \mu\text{T}$ and field variation $rB = 15.9 \%$, which is close to the calculation value of 15% . A picture of the field probe[18] is also included in Fig. 4.11.

4.9 Summary

This chapter presents the design of a transmitter coil with an interleaved structure for multi-load wireless power transfer applications covering a large charging area. The proposed coil achieves a uniform magnetic field, low electric field generation, and high efficiency. Analytical modeling of the coil magnetic field distribution, electric field generation, and power loss is presented and utilized in the coil design process. FEA simulation verifies the uniform magnetic field and the significantly reduced electric field. A prototype $0.5 \text{ m} \times 0.5 \text{ m}$ transmitter coil is built and tested in a 6.78 MHz WPT system. The magnetic field is uniform across the entire charging area with less than 16% variation in amplitude. Following the overall design method, this chapter optimizes the transmitter coil internally, leading to reduced design spaces parameterized by the $[I_1, B_0]$ which emphasizes the interdependence of adjacent stages.

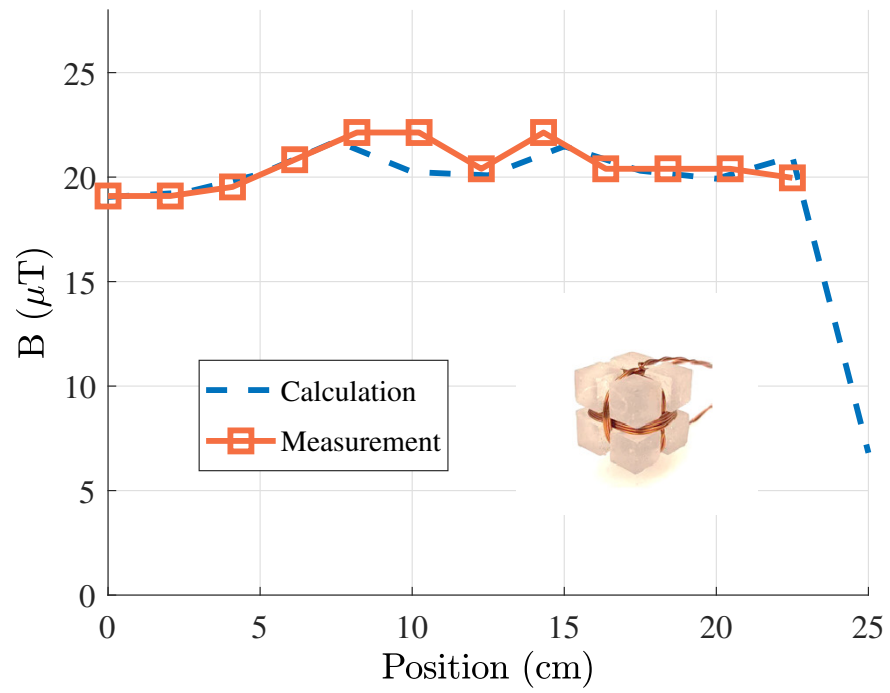


Figure 4.11: Comparison of the designed and measured magnetic field.

Chapter 5

Analysis and Design of Transmitter Power Stage

Chapter 4.1 has analyzed the transmitter coil with a uniform magnetic field. This chapter details the transmitter power stage modeling and design, including the ZVS inverter and the following IMN which converts the output voltage of the inverter to a constant current source for the transmitter coil.

5.1 Analysis and Design of ZVS Inverter

At 6.78 MHz, it is critical to ensure ZVS of inverter devices for safe operation. The circuit schematic of the ZVS inverter has been shown in Fig. 3.3. An auxiliary inductor may be added to the output for ZVS operation when the load current is not sufficient. With this setup, the inverter may achieve ZVS using i_{inv} and i_{zvs} simultaneously.

5.1.1 ZVS Operation Modeling

The inverter schematic is shown in Fig. 3.3. At 6.78 MHz, it is critical to ensure the ZVS of inverter devices for safe operation. An auxiliary inductor L_{zvs} may be added to the output for designs where the load current is not sufficient to obtain ZVS. The inductance is designed so the inverter achieves ZVS via the combination of i_{inv} and i_{zvs} .

Typical ZVS inverter waveforms are shown in Fig. 5.1, where dt_{inv} is the deadtime, ϕ_{inv} is the phase angle between the fundamental components of v_{inv} and i_{inv} , and $t_{\phi_{inv}} = \phi_{inv}(T_s/2\pi)$.

v_{inv} is approximately trapezoidal,

$$v_{inv}(t) = \begin{cases} V_{dc}(\mp 1 \pm \frac{2t}{dt_{inv}}) & , (0 \leq t \leq dt_{inv}) \\ \pm V_{dc} & , (dt_{inv} \leq t \leq T_s/2) \end{cases} \quad (5.1)$$

The fundamental component V_{inv} is extracted from v_{inv} using Fourier analysis

$$V_{inv} = \frac{2}{T_s} \sqrt{\left(\int_0^{T_s} v_{inv} \sin(\omega_s t) dt\right)^2 + \left(\int_0^{T_s} v_{inv} \cos(\omega_s t) dt\right)^2} \quad (5.2)$$

During the ZVS transition interval, both i_{zvs} and i_{inv} flow through the output capacitors of the inverter transistors simultaneously. When the ZVS tank is designed to achieve ZVS without diode conduction, the transition is described by

$$Q_{oss,inv} = \int_0^{dt_{inv}} (i_{zvs}(t) + i_{inv}(t)) dt \quad (5.3)$$

where $Q_{oss,inv}$ is the output charge of one phase leg when blocking V_{dc} , and $C_{oss,inv} = Q_{oss,inv}/V_{dc}$ is the equivalent output capacitance calculated from the nonlinear, voltage-dependent capacitance [25] from device datasheet

$$C_{oss,inv} = \int_0^{V_{dc}} C_{oss}(V) dV \quad (5.4)$$

i_{zvs} during the ZVS transition is calculated by (5.5).

$$\begin{aligned} i_{zvs}(t) = & I_0 \cos\left(\omega_z\left(t + \frac{\beta - \alpha}{\omega_s}\right)\right) + C_{oss}V_{dc}\omega_z \sin\left(\omega_z\left(t + \frac{\beta - \alpha}{\omega_s}\right)\right) + \\ & \frac{I_1}{\left(\frac{\omega_z^2}{\omega_s^2} - 1\right)} \left[\cos\left(\omega_z\left(t + \frac{\beta - \alpha}{\omega_s}\right)\right) \sin(\alpha - \beta) + \right. \\ & \left. \frac{\omega_s}{\omega_z} \sin\left(\omega_z\left(t + \frac{\beta - \alpha}{\omega_s}\right)\right) \cos(\alpha - \beta) + \sin(\omega_s t) \right] \end{aligned} \quad (5.5)$$

where $\omega_z = 1/\sqrt{L_{zvs}C_{oss,inv}}$, $I_0 = i_{zvs}(0)$, and $\beta = dt_{inv}f_s\pi$ is half of the transient time in radians.

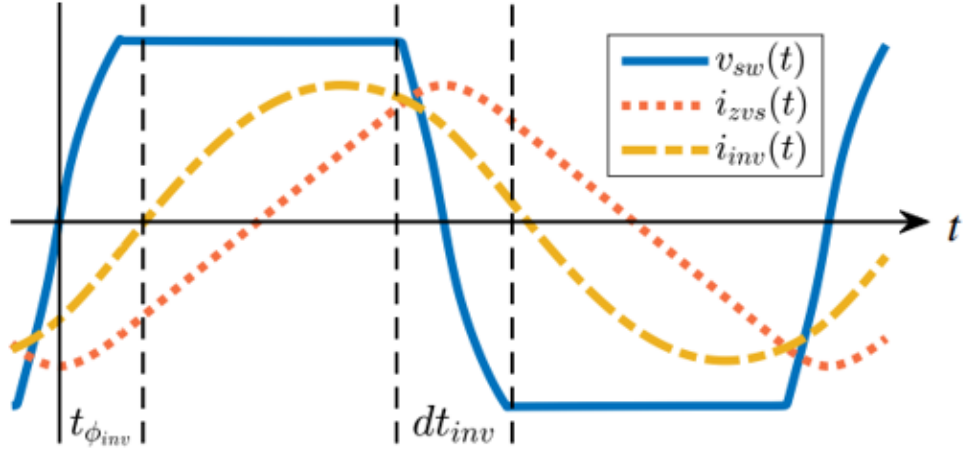


Figure 5.1: Waveform of the ZVS inverter showing the output voltage and current.

The solution for I_0 is given in (5.6)

$$I_0 = \frac{1}{1 + \cos(\omega_z dt)} \left[\frac{V_{dc}}{2L_{zvs}} \left(\frac{T}{2} - dt \right) - \frac{V_{dc}}{2} \sqrt{\frac{C_{oss}}{L_{zvs}}} \sin(\omega_z dt) + \right. \\ \left. \frac{I_1}{L_{zvs} C_{oss} \omega^2 - 1} \left(\sin(\omega_s dt + \alpha - \beta) - \cos(\omega_z dt) \sin(\alpha - \beta) - \right. \right. \\ \left. \left. \frac{\omega_s}{\omega_z} \sin(\omega_z dt) \cos(\alpha - \beta) \right) \right] \quad (5.6)$$

For a specified load V_{inv} and ϕ_{inv} , L_{zvs} is solved from combining (5.6) and (5.5).

5.1.2 Power Loss Modeling

Note that the inverter is designed for full ZVS and no reverse conduction. Thus, the hard switching loss and body diode reverse conduction loss are eliminated. The total power loss consists of three parts: the conduction loss in the drain-to-source on-resistance P_{ds} , the output capacitance hysteresis loss under ZVS condition $P_{ZVS,Coss}$ [82], and the power loss in the auxiliary inductor P_{tank}

$$P_{loss,inv} = P_{ds} + P_{ZVS,Coss} + P_{tank} \quad (5.7)$$

Switch Conduction Loss

During power delivery interval, $i_{tot} = (i_{zvs} + i_{inv})$ flows through the device channel and causes power loss. The conduction loss $P_{ds} \approx I_{tot,rms}^2 \cdot R_{ds}$. where R_{ds} is the on-resistance of the transistor, and $I_{tot,rms}$ is the RMS value of $(i_{inv} + i_{zvs})$

$$I_{tot,rms} = \sqrt{\frac{1}{0.5T - dt_{inv}} \int_{dt_{inv}}^{0.5T} i_{tot}(t)^2 dt} \approx \sqrt{\frac{2}{T} \int_0^{0.5T} i_{tot}(t)^2 dt} \quad (5.8)$$

where the approximation is valid when $dt_{inv} \ll 0.5T_s$.

The load current i_{inv} is calculated from the inverter loading conditions $P_{o,inv} = V_{inv} I_{inv} \cos(\phi_{inv})$

$$i_{inv}(t) = I_{inv} \sin(\omega_s(t - t_{\phi_{inv}})) \quad (5.9)$$

The inductor current i_{zvs} is approximated as a triangular wave when calculating the conduction loss, which is valid for $dt_{inv} \ll 0.5T_s$ [54]. The peak value of i_{zvs} is

$$I_{zvs,pk} = \frac{V_{dc}}{8L_{zvs}f_s} \quad (5.10)$$

i_{zvs} is further expressed as

$$i_{zvs}(t) = \begin{cases} I_{zvs,pk} \left(\frac{4t}{T_s} - 1 \right) & , (0 \leq t \leq 0.5T_s) \\ I_{zvs,pk} \left(1 - \frac{4(t - 0.5T_s)}{T_s} \right) & , (0.5T_s \leq t \leq T_s) \end{cases} \quad (5.11)$$

Coss Hysteresis Loss

For ZVS converters operating in the MHz range and using high-voltage GaN switches, a significant additional power loss is observed in the transistors with R_{ds} conduction loss. The additional loss increases with V_{ds} and f_s , and is highly related to the device in use. In a recent study [82], this part of power loss is explained by the energy loss during the repeated charging and discharging of the C_{oss} capacitor. A plot of the power loss is shown in Fig. 5.2 [82].

This type of energy loss is summarized as hysteresis loss and is modeled using a curve fit based on an experimental test of sample devices. In the experiments [82], various devices with differing C_{oss} are tested under a wide range of V_{dc} and f_s to find the empirical fit parameters. For the 650 V GaN transistors made by GaN System Inc., the C_{oss} loss under ZVS condition is

$$P_{ZVS,Coss} = 4.5 \times 10^{-11} \frac{C_{oss}}{C_{ref}} \left(\frac{V_{dc}}{650} \right)^{1.6} f_s^{1.7} \quad (5.12)$$

where C_{ref} is the output capacitance of the reference device GS66504B [82].

It should be noted that the parameters are only for the GaN system 650 V family and may be accurate only when device C_{oss} is around the reference device C_{ref} . The parameters have an obvious shift when using other device families [82].

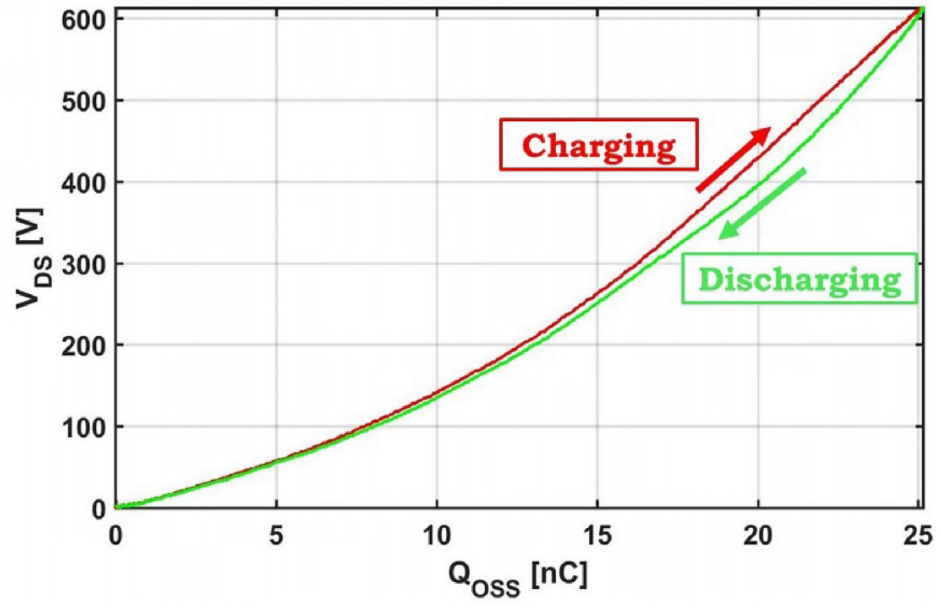


Figure 5.2: Charge and discharge periods for a typical GaN transistor showing the energy lost [82].

Auxiliary Inductor Design and Power Loss

The power loss in the auxiliary inductor consists of copper loss P_{wire} and core loss P_{core} , which are calculated under a specified L_{zvs} . In this work, T-68 iron powder toroid core [11] is used to fit in the target application. Iron powder is commonly used in RF applications and has low core loss. The selection of wire depends on the available space and power loss.

To achieve a specified L_{zvs} , the number of turns N_z is [11]

$$N_z = \sqrt{\frac{L_{zvs}}{k_z}} \quad (5.13)$$

where k_z is an empirical inductance coefficient provided by the manufacturer datasheet.

The length of wire l_{zw} is

$$l_{zw} = MLT \cdot N_z \quad (5.14)$$

where MLT is the mean length per turn.

At 6.78 MHz, the skin depth of copper δ is 25 μm and Litz wire for this frequency is not commercially available. Thus, solid copper wire is used, whose outer radius is denoted as $r_{z,0}$. The effective conduction area A_{ac} is, assuming $r_{z,0} \gg \delta$,

$$A_{ac} = (r_{z,0}^2 - (r_{z,0} - \delta)^2) \quad (5.15)$$

Combining l_{zw} and A_{ac} , the copper ESR is

$$R_{zvs,wire} = \frac{MLT \rho_{cu} \sqrt{\frac{L_{zvs}}{k_z}}}{(r_{z,0}^2 - (r_{z,0} - \delta)^2)} \quad (5.16)$$

From (5.16), the wire ESR decreases as $r_{z,0}$ and effective conductive area increases. Thus, for each L_{zvs} , the thickest wire is preferred for minimum ESR. With the isolation gap between adjacent turn l_{gap} , the maximum wire radius is

$$r_{z,max} = \frac{2\pi r_{z,in} - N_z l_{gap}}{2(N_z + \pi)} \quad (5.17)$$

where $r_{z,in}$ is the inner radius of the T-68 core provided by the manufacture datasheet.

The final selection of $r_{z,0}$ depends on $r_{z,max}$ and the closest available AWG diameter.

The inductor current i_{zvs} has been modeled in (5.10).

Combining (5.10) and (5.16), the power loss in the wire is

$$P_{wire} = \frac{1}{3} \cdot \left(\frac{V_{dc}}{8L_{zvs}f_s} \right)^2 \frac{\rho_{cu}MLT\sqrt{\frac{L_{zvs}}{k_z}}}{\pi(r_{z,0}^2 - (r_{z,0} - \delta)^2)} \quad (5.18)$$

The core loss is calculated using the empirical equation provided by the manufacturer [11]

$$P_{core} = \left(\frac{f}{\frac{4 \times 10^9}{B^3} + \frac{3 \times 10^8}{B^{2.3}} + \frac{2.7 \times 10^6}{B^{1.65}}} + 3 \times 10^{-15} f^2 B^2 \right) V_{ol} \quad (5.19)$$

where $B = I_{zvs,pk}L_{zvs} \cdot 10^8 / (A_e N_z f_s)$ is the magnetic flux density inside the core, A_e is the effective magnetic cross section, and V_{ol} is the core volume.

Using the design method, the inductor power loss vs. target inductance is shown in Fig. 5.3. The inductor loss and each of its components decrease rapidly when the requirement of ZVS current reduces. The glitches in the copper loss indicate the change of wire gauge, which is explicitly shown in Fig. 5.4, together with the peak auxiliary current. It should be noted that the peak current is an approximation assuming $dt_{inv} \approx 0$. The real peak current will be smaller.

5.1.3 Switching Device Selection

From the ZVS modeling and power loss analysis, switching device characteristics, i.e. output charge Q_{oss} and drain-to-source on-resistance R_{ds} , play an important role and must be considered when designing the inverter. Ideally, the output capacitance is desired to be near zero to reduce the requirement of additional inductive current for ZVS operation and small power loss. And the device R_{ds} is desired to be near zero for low power loss. The two proprieties, however, cannot be achieved simultaneously.

A family of 650 V GaN switches produced by GaN Systems Inc. is studied in this work. Their R_{ds} ranges from 67 mΩ to 450 mΩ. The instantaneous non-linear output capacitance vs. blocking voltage is summarized in Fig. 5.5 from the datasheet.

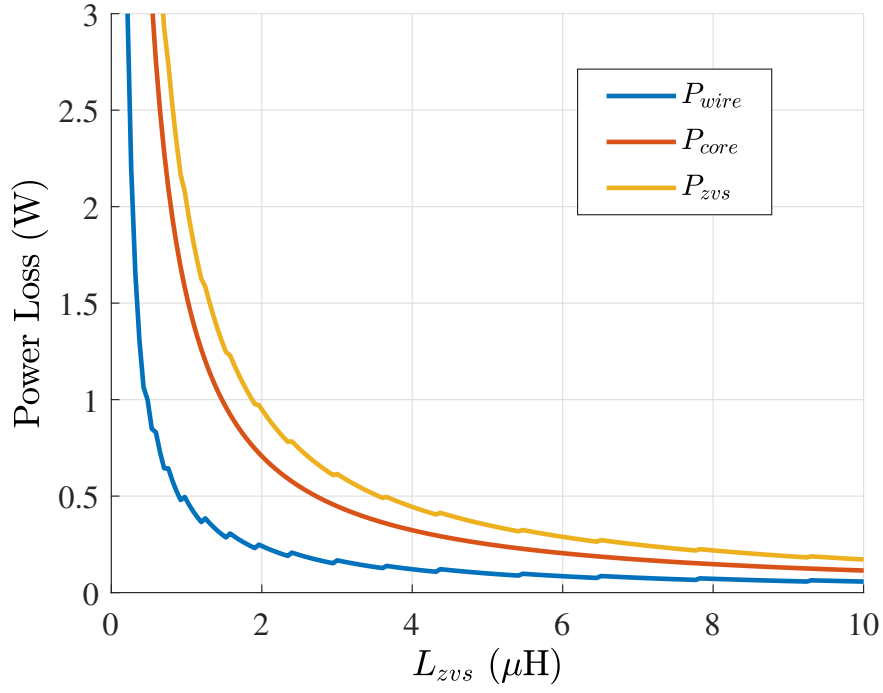


Figure 5.3: Auxiliary inductor power loss vs. a range of target inductance.

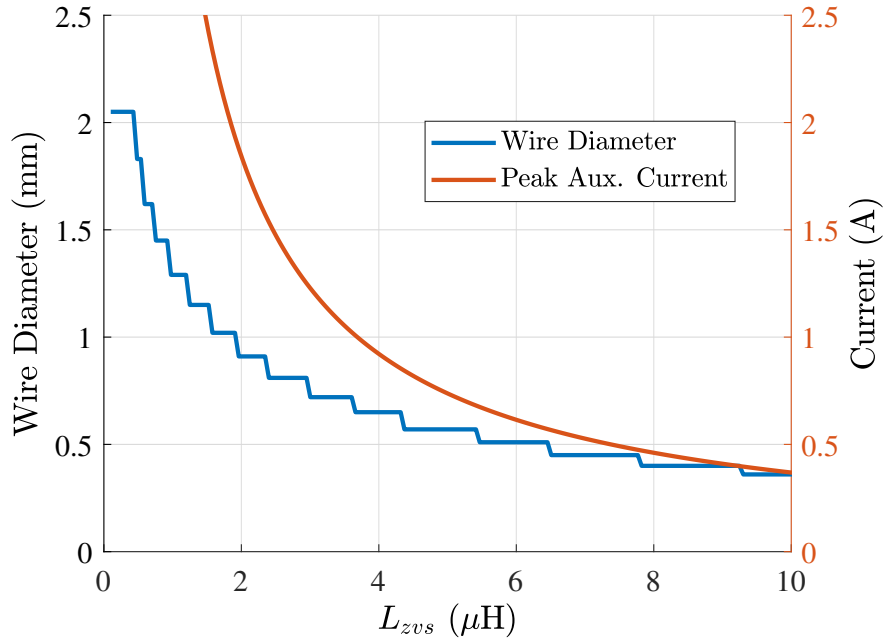


Figure 5.4: Designed wire gauge and peak auxiliary current vs. a range of target inductance.

The equivalent Q_{oss} and C_{oss} are calculated using the aforementioned (5.4), and are presented in Table 5.1 together with R_{ds} . Devices with small R_{ds} have relatively large Q_{oss} .

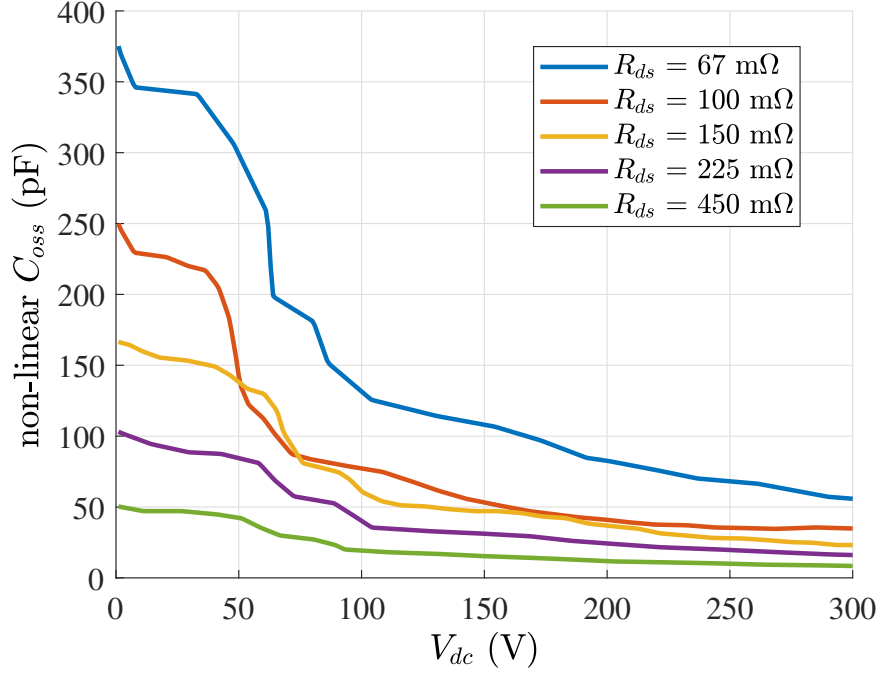


Figure 5.5: Instantaneous non-linear output capacitance vs. blocking voltage of the 650 V GaN devices produced by GaN Systems Inc..

Table 5.1: Characteristics of the GaN switches.

Part number	R_{ds} (Ω)	Q_{oss} @ V_{dc} (nC)	C_{oss} @ V_{dc} (pF)	FOM
GS66506T	0.067	37.2	186	0.40
GS66504B	0.100	21.1	106	0.47
GS065011	0.150	17.2	84	0.40
GS065008	0.225	10.7	52	0.43
GS065004	0.450	5.3	26	0.42

Reviewing the characteristics, small R_{ds} and Q_{oss} cannot be achieved simultaneously. The product of R_{ds} and Q_{oss} is near-constant within the 650 V GaN System device family. The device figure-of-merit is

$$FOM = \frac{1}{R_{ds} \cdot Q_{oss}} \quad (5.20)$$

where R_{ds} is expressed in Ω and Q_{oss} is expressed in nC.

Since the output capacitance of the last two devices under study are much smaller ($\leq 50\%$) than the reference device, the C_{oss} hysteresis loss modeling may not be accurate [82]. In the device optimization process, only the first three devices are studied which already have a wide range of R_{ds} and C_{oss} .

5.1.4 ZVS Inverter Design Method

As stated at the beginning, the inverter stage in a WPT system is desired to provide a voltage source and power the rest of the system with high efficiency. Using a combination of the high-efficiency approximation and fundamental harmonic analysis, the inverter output power $P_{o,inv}$ is held constant and equal to the total output power $P_{o,all}$ across varying designs. The coupling parameters of the inverter $[V_{inv}, \phi_{inv}]$ are highlighted in Fig. 3.3.

Thus, the goal of the inverter design in isolation is to minimize power loss for each combination of $[V_{inv}, \phi_{inv}]$ within a possible range, through designing inverter internal parameters, e.g. device selection, dt_{inv} , L_{zvs} and the inductor wire implementation.

For a specified $[V_{inv}, \phi_{inv}]$ and fixed device, dt_{inv} is calculated from (8.5) and (5.2), and L_{zvs} is calculated from (5.3)-(5.6). With L_{zvs} , the inductor is designed using the method in 5.1.2 and P_{core} is calculated. With L_{zvs} , i_{zvs} and i_{tot} are derived and used to calculate P_{ds} . The inverter design is completed for a fixed device and a range of load. Iteration among different switching devices gives multiple inverter designs that achieve the same $[V_{inv}, \phi_{inv}]$ with various power loss. Only the design with minimum power loss is kept and other designs with higher power loss are eliminated. Consequently, the inverter design is converted to a selection of $[V_{inv}, \phi_{inv}]$. For each combination, the associated minimum loss is stored in a two-dimensional look-up-table. The selection of $[V_{inv}, \phi_{inv}]$ will be determined at the system-level in Chapter 7.

In the example design, V_{inv} is iterated from 116 V to 126 V which corresponds to a dead-time range of 10 ns to 35 ns when operating from a 200 V PFC supply [73]. ϕ_{inv} is iterated from resistive load to 60° inductive load. The GaN transistors in the analysis have been summarized in Table 5.1.

The inverter power loss using GS065011 is shown in Fig. 5.6. And the auxiliary inductor loss is shown in Fig. 5.7. The required auxiliary current decreases as the inductive ϕ_{inv} and

current I_{inv} increases, which leads to a reduced P_{tank} . However, because I_{inv} increases with ϕ_{inv} , P_{ds} also increases which limits ϕ_{inv} . The blank region is where the inverter cannot achieve ZVS at the corresponding dt_{inv} . The magnitude of $(i_{inv} + i_{zvs})$ at dt_{inv} is shown in Fig. 5.8 together with each of its components for $\phi_{inv} = 30^\circ$ design. The magnitude reduces to zero at the edge of the blank region. The total current would go negative and harm ZVS if further extending the dead time from the edge.

Iterating different devices, the minimum inverter loss design is shown in Fig. 5.9. The blank region of GS065011 is filled by other devices with larger capacitance. Fig. 5.9 will be used in the system-level design in Chapter 7

5.2 Analysis and Design of IMN

5.2.1 Basics for Impedance Matching

In the proposed multi-load WPT system, the IMN stage provides a constant current to the Tx coils regardless of receiver side loading or positional change. The IMN stage consists of multiple passive inductors and capacitors, connected in a special topology to perform impedance matching.

Since the receiver side is usually designed to achieve no circulating current and high-efficiency [31, 58, 53], it is assumed that the reflected power to the transmitter side approximately is resistive and equals the total output power $P_{o,all}$.

Thus, the equivalent resistance of the receivers are

$$R_r = \frac{P_{o,all}^2}{I_{t,rms}^2} \quad (5.21)$$

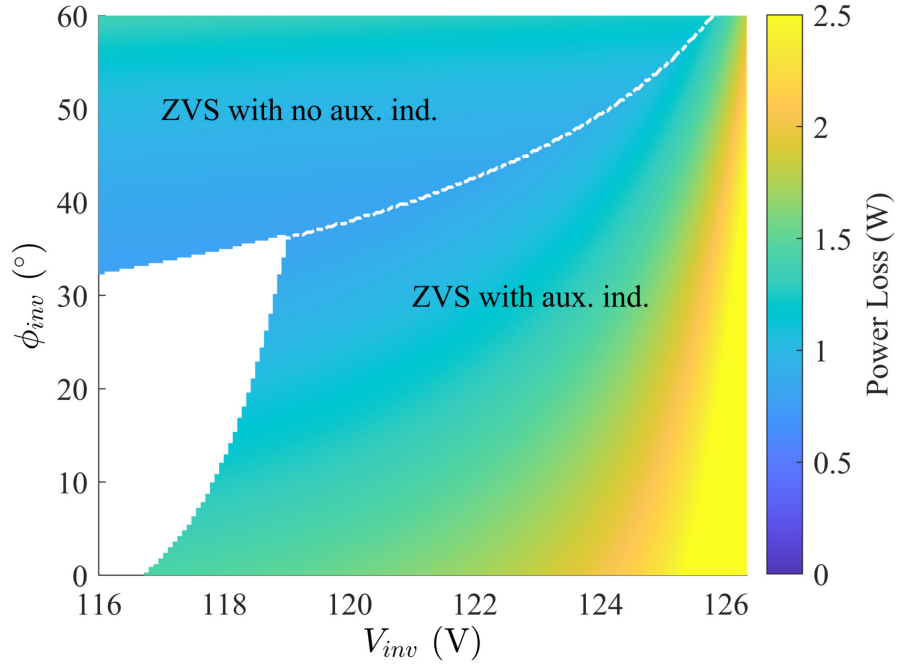


Figure 5.6: Inverter power loss design result for GS065011, including device conduction loss and auxiliary inductor loss.

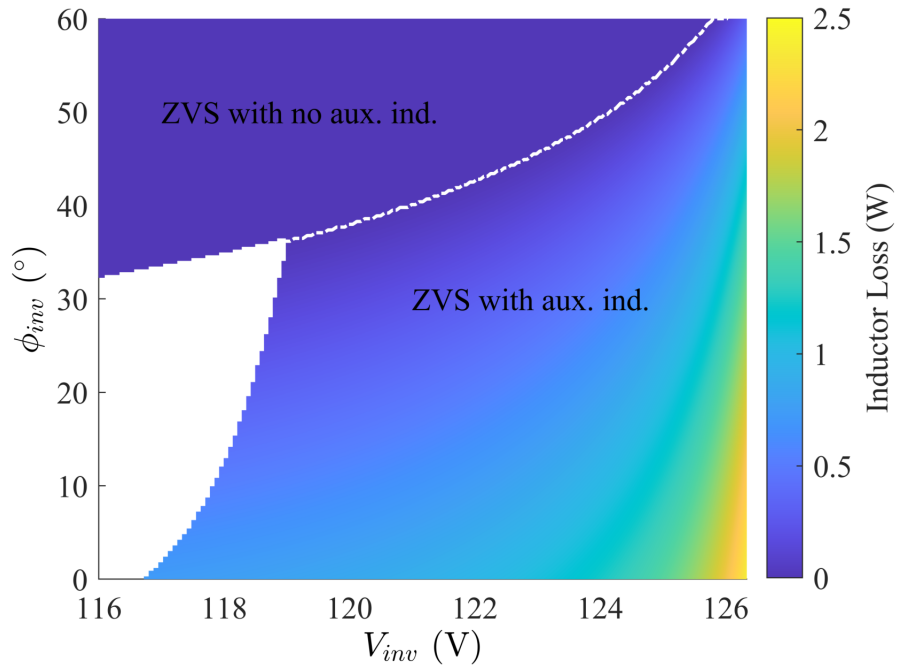


Figure 5.7: Inverter power loss design result for GS065011, only showing the power loss in the inductor.

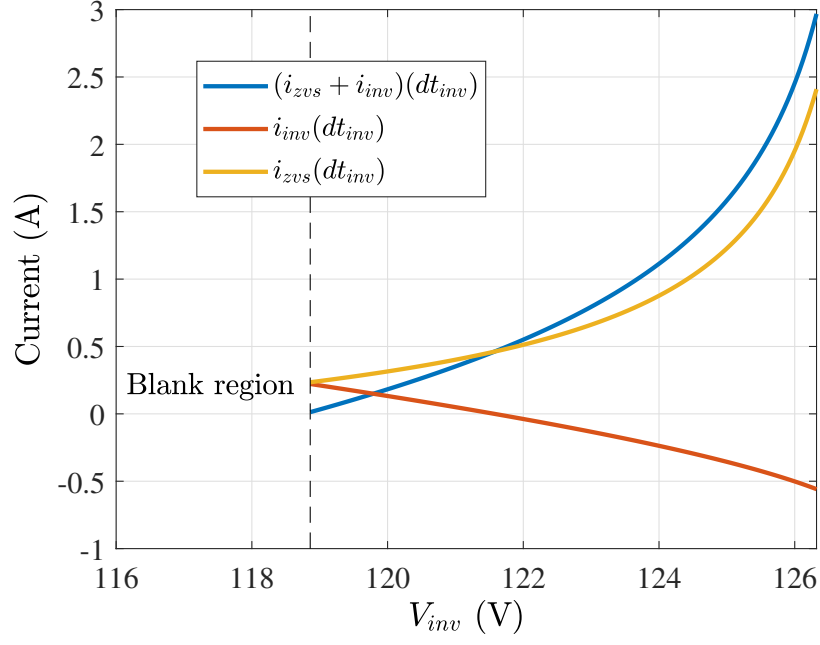


Figure 5.8: Inverter output current for load $\phi_{inv} = 30^\circ$.

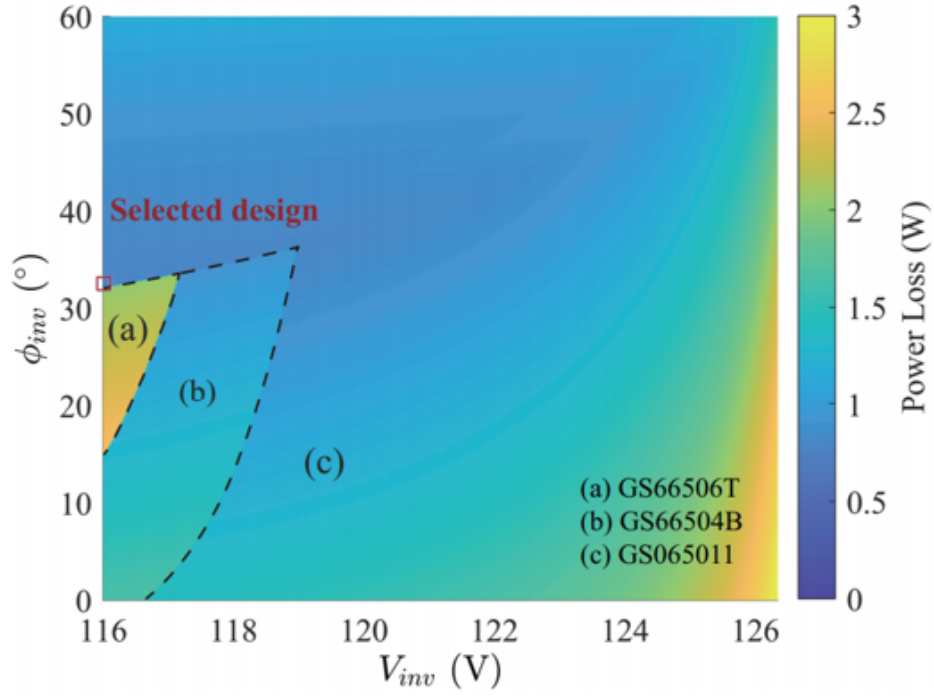


Figure 5.9: Optimized inverter power loss with all iterated devices.

The circuit schematic of the IMN stage, together with the Tx coil and the reflected load of receivers, are shown in Fig. 5.10.

In the transmitter coil design, discrete compensation is used to compensate the inductance, which gives $L_1 = -1/(\omega_s^2 C_1)$. Solving the circuit using KCL and KVL, the total impedance seen from the inverter is

$$Z_{imn} = j\omega_s L_{imn1} + \frac{\frac{1}{j\omega_s C_{imn}} Z_{n2}}{\frac{1}{j\omega_s C_{imn}} + Z_{n2}} \quad (5.22)$$

where Z_{n2} is the total impedance of the right side branch

$$Z_{n2} = j\omega_s L_{imn2} + R_r + R_1 \quad (5.23)$$

Assuming the phase of v_{sw} is zero, the inverter output current is

$$i_{inv} = \frac{V_{inv}}{Z_{inv}} \quad (5.24)$$

Further solving the circuit, the Tx coil current i_{tx} is

$$i_{tx} = i_{inv} \frac{\frac{1}{j\omega_s C_{imn} Z_{n2}}}{j\omega_s C_{imn} + Z_{n2}} \quad (5.25)$$

Simplifying i_{tx} gives

$$i_{tx} = \frac{V_{inv}}{j\omega_s L_{imn1} + Z_{n2}(1 - \omega_s^2 L_{imn1} C_{imn})} \quad (5.26)$$

If L_{imn1} and C_{imn} are designed such that their impedances are canceled by the other, $(1 - \omega_s^2 L_{imn1} C_{imn})$ will equal zero [46]. L_{imn1} is [46]

$$L_{imn1} = \frac{V_{inv}}{I_t} \quad (5.27)$$

C_{imn} is designed to compensate L_{imn1}

$$C_{imn} = \frac{1}{\omega_s^2 L_{imn1}} \quad (5.28)$$

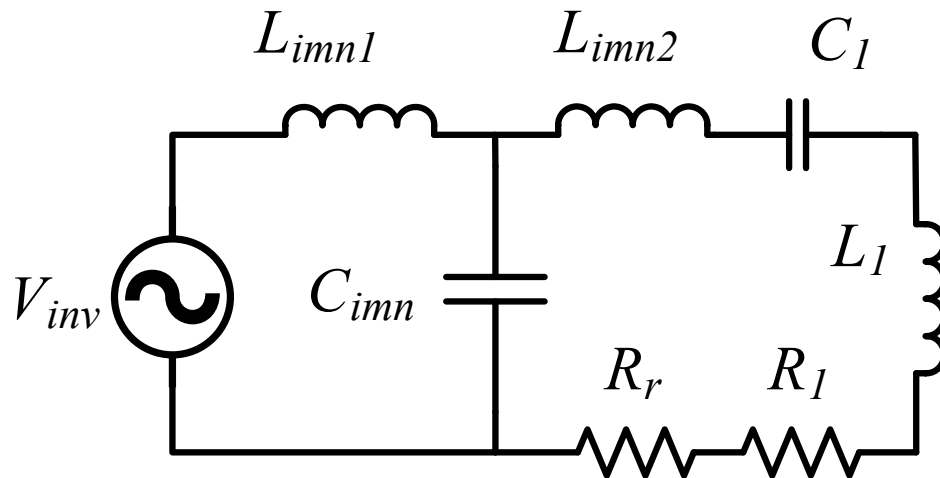


Figure 5.10: Circuit schematic of the IMN stage, with receiver side modeled as an equivalent reflected resistance.

The inverter output impedance is

$$Z_{inv} = \frac{V_{inv}}{I_{inv}} (\cos(\phi_{inv}) + j \sin(\phi_{inv})) \quad (5.29)$$

Thus, i_{tx} is independent of Z_{n2}

$$i_{tx} = \frac{V_{inv}}{j\omega_s L_{imn1}} \quad (5.30)$$

when $L_{imn1} = 1/(\omega_s^2 C_{imn})$.

The inverter output impedance is

$$Z_{inv} = \frac{V_{inv}}{I_{inv}} (\cos(\phi_{inv}) + j \sin(\phi_{inv})) \quad (5.31)$$

The reflected load from the receivers is $R_r = 2P_{o,all}I_t^{-2}$, which typically is much greater than R_1 in a high-efficiency system. Solving the equivalent circuit, the impedance of Z_{imn2} is

$$Z_{imn2} = \frac{(Z_{inv} - j\omega_s L_{imn1})}{1 - (Z_{inv} - j\omega_s L_{imn1})(j\omega_s C_{imn})} - R_r \quad (5.32)$$

The imaginary part of Z_{imn2} may range from negative to positive, and thus may be physically implemented as either an inductance or capacitance.

The current in C_{imn} , which will be used to calculate capacitor power loss, is derived as

$$i_c = -\frac{V_{inv}(R_1 + R_r + Z_{imn2})}{Z_{inv} \cdot j\omega_s L_{imn1}} \quad (5.33)$$

Note that the IMN stage is also desired to provide an inductive load current such that the inverter achieves ZVS. When the target phase angle is ϕ_{inv} , the desired inverter output impedance is

$$Z_{inv} = \frac{V_{inv}}{I_{inv}} (\cos(\phi_{inv}) + j \sin(\phi_{inv})) \quad (5.34)$$

Comparing (5.22) with (5.34), the impedance of L_{imn2} is

$$Z_{imn2} = \frac{(Z_{inv} - j\omega_s L_{imn1})}{1 - (Z_{inv} - j\omega_s L_{imn1})(j\omega_s C_{imn})} - R_r \quad (5.35)$$

The imaginary part of Z_{imn2} may range from negative to positive. When it is positive, the branch is an inductor with $L_{imn2} = \Re\{Z_{imn2}/(j\omega_s)\}$. When the imaginary part is negative, the branch is a capacitor with $C_{imn2} = \Re\{1/(j\omega_s Z_{imn2})\}$.

The IMN stage power loss $P_{loss,IMN}$ consists of inductor loss and dielectric loss in the capacitor. For any specified inductance value, the inductor design and power loss calculation use the same method as the ZVS inductor. The core of the IMN inductors is implemented using T-68, which is the same as the ZVS inductor to fit in the application and for low core loss. The IMN capacitors are also implemented with NP0 dielectric for small power loss. The capacitor power loss is calculated using (4.18).

5.2.2 IMN Design

The IMN stage is coupled with the inverter and desired to generate a constant I_t for the Tx coil and exhibit an inductive phase ϕ_{inv} for the inverter ZVS operation. The coupling parameters are highlighted in Fig. 3.3. For a possible range of $[V_{inv}, \phi_{inv}, I_t]$, the IMN parameters L_{imn1} , C_{imn} , and L_{imn2} (or C_{imn2}) are designed using (5.27), (5.28), and (5.35). For any specified L_{imn1} or L_{imn2} , the inductor is designed using the same method in Section 5.1.2 to minimize power loss. For any specified C_{imn} or C_{imn2} , the power loss is calculated using (4.18).

Consequently, the IMN design is converted to a selection of $[V_{inv}, \phi_{inv}, I_t]$, where each combination is associated with a minimized power loss. The selection of $[V_{inv}, \phi_{inv}, I_t]$ will be determined at the system-level in Section 7 to capture the dependence on adjacent stages.

In the example design, $[V_{inv}, \phi_{inv}, I_t]$ use the same range as in the previous inverter and Tx coil example design. The resulting IMN design space is shown in Fig. 5.11. In this figure, multiple surfaces for varying V_{inv} are nearly completely overlapped, which indicates that the IMN design is insensitive to the range of V_{inv} from the inverter design.

To better understand the power loss distribution, a detailed loss breakdown is given in Fig. 5.12 for an example design with $I_t = 1.5$ A, $\phi_{inv} = 32^\circ$, and $V_{inv} = 116$ V. Because L_{imn1} decreases when I_t increases, and I_{inv} remains constant for a given ϕ_{inv} , the power loss in L_{imn1} decreases with I_t . When I_t is greater than 1 A, Z_{imn2} is implemented with an inductor and has an increased power loss. The final selection of IMN design is marked in

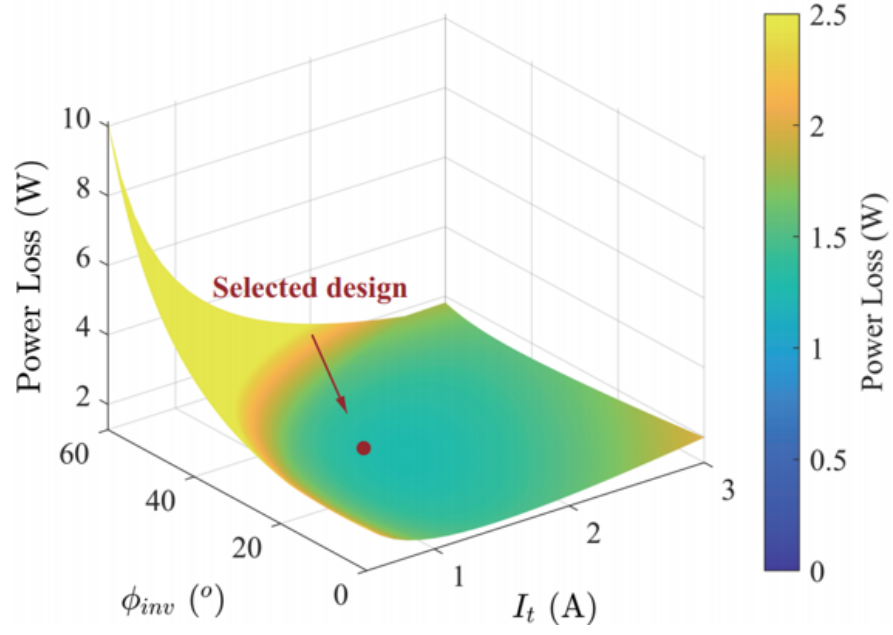


Figure 5.11: IMN power loss result. The surfaces with various V_{inv} are close.

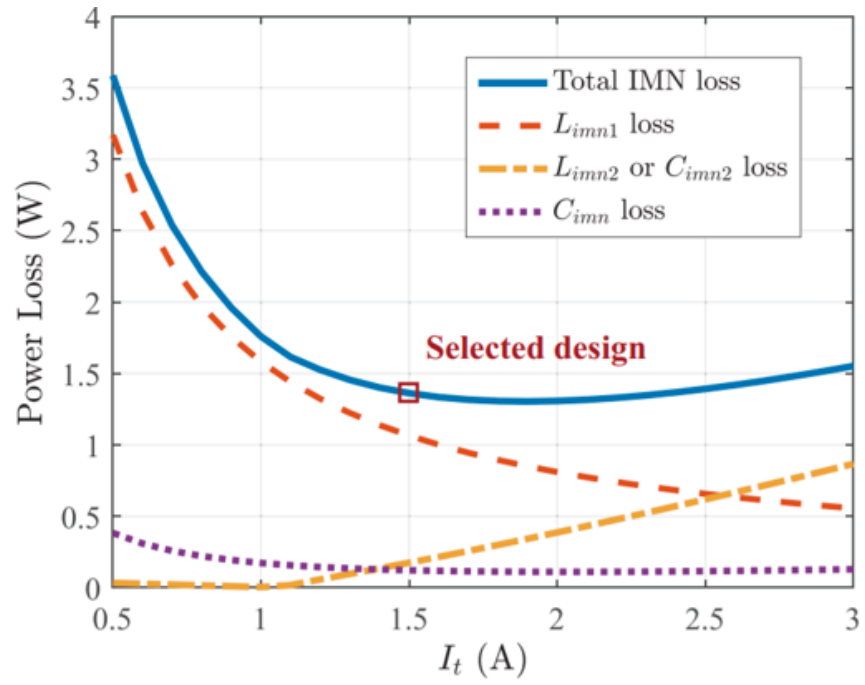


Figure 5.12: IMN power loss breakdown at the proposed working point.

both figures, resulted from the system-level design which will be detailed in Chapter 7. The selected design is not the minimum power loss IMN design but is a design that will result in minimum system-wide power loss.

5.3 Summary

This chapter details the transmitter side power stage modeling and design, including the ZVS inverter and the following IMN which converts the output voltage of the inverter to a constant current source for the transmitter coil. Analytical modeling of the ZVS and the impedance matching are provided. An LTSpice simulation of an example 6.78MHz resonant inverter verifies the ZVS design and the constant current behavior. Following the overall design method, this chapter optimizes the inverter stage and the IMN stage internally, leading to reduced design spaces parameterized by the coupling parameters of adjacent stages for system-level design.

Chapter 6

Receiver Design and Optimization

Chapter 4.1 has analyzed the transmitter coil with a uniform magnetic field. Chapter 5 has discussed the transmitter side power stage that maintains a constant amplitude current to the transmitter coil. Combining the coil and the power stage, the transmitter generates a magnetic field that is uniform regardless of receiver side positional or load changes. In this chapter, receivers that are working in the uniform magnetic field are investigated, including the diode rectifier working in the MHz range and the receiver coil.

6.1 Rectifier Operating in 6.78 MHz

6.1.1 High-frequency Rectifier Modeling

Previously, diode rectifiers operating in the kHz range are usually modeled as a resistive load, neglecting the turn-on and turn-off transient. The accuracy has been verified in many reported systems [52, 17]. However, when the operating frequency moves further to 6.78 MHz, the transition of the diode rectifier constitutes a significant portion of the overall period and causes reactive power and circulating current. Failing to consider the magnified transient interval usually leads to a significant reduction of system efficiency, or even resonant failure [37, 58, 23].

Typical waveforms of a 6.78 MHz diode rectifier working in a series-resonant wireless power transfer system are shown in Fig. 6.1. i_{rec} is approximated as sinusoidal because of the series resonance formed by L_2 and C_2 and tuned near the system switching frequency.

One symmetric half-period of operation is divided into two intervals: the switching transition interval and the power delivery interval. During the transition interval, i_{rec} flows through the parallel junction capacitances of the reverse-biased diodes

$$Q_{j,rec} = \int_0^{dt_{rec}} I_{rec} \sin(\omega_s t) dt \quad (6.1)$$

where dt_{rec} is the transient time, I_{rec} is the amplitude of i_{rec} , and $Q_{j,rec} = C_{j,rec} V_o$ is two times the charge stored by a single diode capacitance when blocking V_o . It should be noted $C_{j,rec}$ is the equivalent junction capacitance at V_o , which is calculated similarly to the inverter switching device (5.4).

During the power delivery interval, i_{rec} flows through the forward-biased diodes to the output, delivering an average power

$$P_o = \frac{2}{T_s} \int_{dt_{rec}}^{T_s/2} V_o I_{rec} \sin(\omega_s t) dt \quad (6.2)$$

with power loss

$$P_{loss,rec} = \frac{4}{T_s} \int_{dt_{rec}}^{T_s/2} V_f I_{rec} \sin(\omega_s t) dt \quad (6.3)$$

where $V_f(t)$ is the forward voltage drop on each diode, which is approximately determined by the junction barrier voltage V_0 and the equivalent diode resistance R_d , $V_f(t) = V_0 + R_d \cdot i_{rec}(t)$.

The total input power is the sum of output power and power loss $P_{rec,in} = P_{loss,rec} + P_o$.

I_{rec} is solved from combining (6.1) and (6.2)

$$I_{rec} = \frac{\pi P_o + V_o \omega_s Q_{j,rec}}{2V_o} \quad (6.4)$$

The analytical expression of v_{rec} is

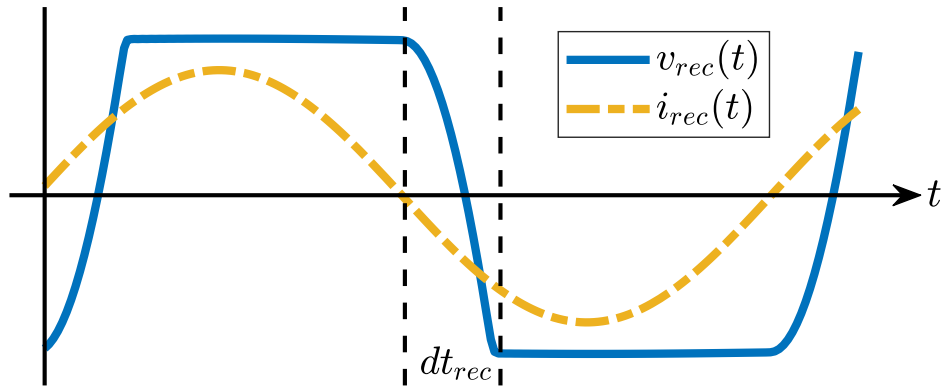


Figure 6.1: Waveform of the diode rectifier showing the input voltage and current.

$$v_{rec}(t) = \begin{cases} \mp V_o + 2 \int_0^t \frac{I_{rec} \sin(\omega_s t)}{C_{oss,rec}} dt & , 0 \leq t \leq dt_{rec} \\ \pm V_o \pm 2(V_0 + R_d \cdot I_{rec} \sin(\omega_s t)) & , dt_{rec} \leq t \leq \frac{T_s}{2} \end{cases} \quad (6.5)$$

where the terms with varying signs take on alternating polarity in subsequent half-periods. The fundamental component V_{rec} of v_{rec} is extracted using Fourier analysis

$$V_{rec} = \frac{2}{T_s} \sqrt{\left(\int_0^{T_s} v_{rec} \sin(\omega_s t) dt \right)^2 + \left(\int_0^{T_s} v_{rec} \cos(\omega_s t) dt \right)^2} \quad (6.6)$$

The input phase ϕ_{rec} is

$$\phi_{rec} = \arccos \left(\frac{P_o + P_{loss,rec}}{I_{rec} V_{rec}} \right) \quad (6.7)$$

Thus, the equivalent impedance of the rectifier Z_{db} with any specified output is modeled by

$$Z_{db} = \frac{2(P_o + P_{loss,rec})}{I_{rec}^2} \cdot (1 - j \tan(\phi_{rec})) \quad (6.8)$$

6.1.2 Schottky Diode Selection

In the high-frequency modeling and power loss analysis, the diode characteristics, i.e. $Q_{j,rec}$ and V_f , play an important role. Ideally, $Q_{j,rec}$ is desired to be near zero to reduce the circulating current I_{rec} , and V_f is desired to be near zero for low conduction loss. In fabrication, for a fixed voltage rating, low V_f and small $Q_{j,rec}$ usually contradicts each other.

Considering the receiver output voltage limit, a family of 40 V Schottky diodes produced by Vishay Intertechnology Inc. is investigated in this work. The instantaneous non-linear junction capacitance vs. blocking voltage is summarized in Fig. 6.2 from the datasheet assuming 25° temperature.

The equivalent Q_j and C_j are calculated using the same method as calculating the GaN device (5.4), and are presented in Table 6.1 together with V_0 and R_d . Devices with small V_f have relatively large Q_j .

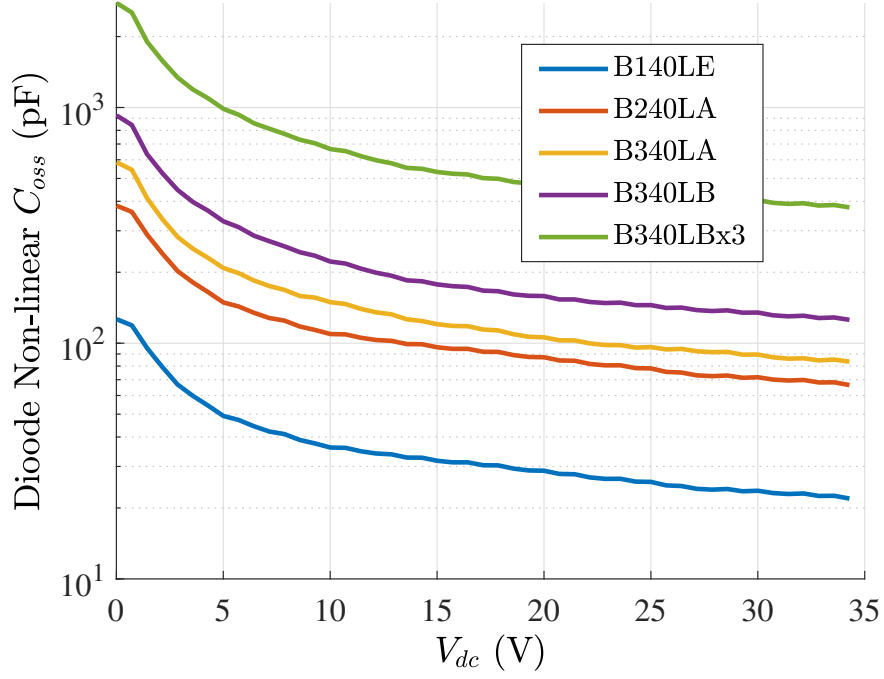


Figure 6.2: Instantaneous non-linear junction capacitance vs. blocking voltage of the 40 V Schottky diodes produced by Vishay Intertechnology Inc..

Table 6.1: Characteristics of the Schottky diodes.

Part number	V_f @ 1 A (V)	V_o (V)	R_d	C_j @ 30 V (pF)
B140E3	0.65	0.42	0.23	41
B240LA	0.42	0.32	0.1	124
B340LA	0.39	0.31	0.08	167
B340LB	0.36	0.29	0.07	255

It should be noted that B140E3, whose average current rating is 1 A, cannot power the required 50 W load and will be excluded from the rectifier design. Part of the design region using B240LA requires an input current exceeding the 2 A limit, which will also be excluded.

6.1.3 Rectifier Design

In many systems, the receiver side is desired to have high efficiency and tuned to be resistive. Using fundamental harmonic analysis, the coupling parameters of the rectifier $[V_{rec}, I_{rec}]$, have been highlighted in Fig. 3.3. Those parameters influence the Rx coil stage through involvement in the compensation design, induced voltage design, and power loss design.

From (6.4), I_{rec} is solved by P_o, V_o and specified diode device in Table 6.1. For each device, when V_o is iterated within the range constrained by the application, the coupling parameter V_{rec} is solved from combining (6.5) and (6.6), and I_{rec} is solved using (6.4). ϕ_{rec} and $P_{loss,rec}$ are solved using (6.7) and (6.3). Iterating among various devices, the resulting candidate rectifier designs are stored to a three-dimensional design space, parameterized by the coupling parameters $[V_{rec}, \phi_{rec}, I_{rec}]$, thus exhibiting differing influences on current amplitudes and power losses throughout the system.

In the example, V_o is iterated from 10 to 31.5 V according to the application requirement, with a step of 0.5 V. The resulting rectifier characteristics with the coupling parameters $[V_{rec}, \phi_{rec}, I_{rec}]$, are shown in Fig. 6.3.

For all devices, the rectifier prefers high voltage and small current operation for reducing power loss. Within the constrained V_o , devices with smaller V_f tend to achieve smaller loss but have more circulating current as indicated by the increased phase. Thus, though paralleling three B340LB yields the rectifier design with the smallest power loss, it is not clear whether such implementation is the best choice since the magnified circulating current may cause a high power loss on the receiver coil. The reactance also interacts with Rx coil compensation design. Moreover, the paralleled devices also cost more and require more space. Therefore, device selection on the rectifier needs an overall consideration with the Rx coil.

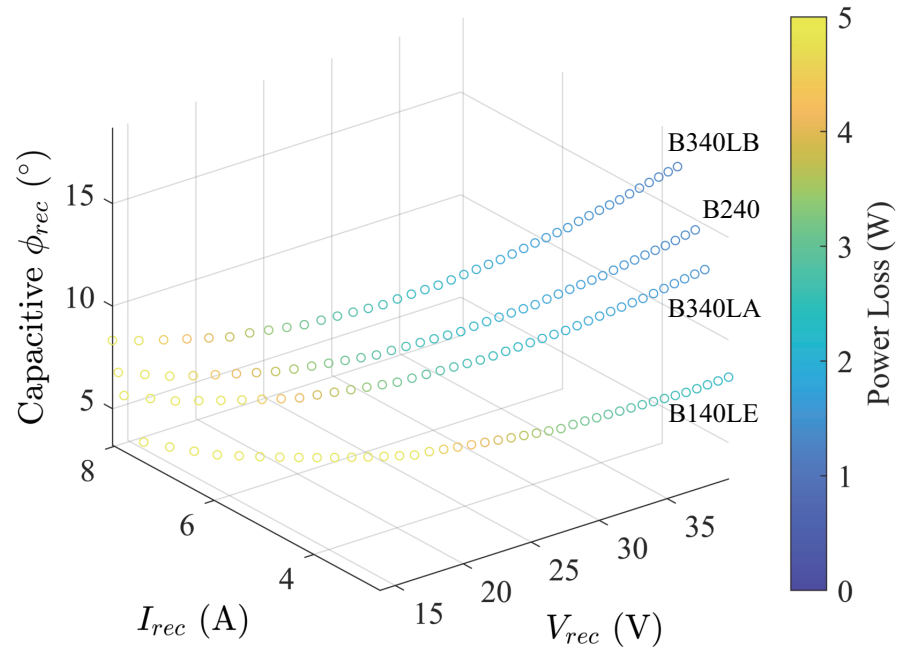


Figure 6.3: Rectifier design result including multiple paralleled Schottky diodes.

6.2 Receiver Coil Modeling

6.2.1 Coil Structure and Power Loss

In the receiver, the Rx coil generates an induced voltage V_{ind} from the B -field and provides current to the rectifier. And the compensation capacitor C_2 is added in series with the Rx coil to compensate the receiver side reactance. To fit in the specified application, conventional rectangular coil is used, which is characterized by the number of turns N_r , half-length $a_{rx} = [a_{r1} \dots a_{rN_r}]$, half-width $b_{rx} = [b_{r1} \dots b_{rN_r}]$ and wire radius. Because Litz wire is prohibitively expensive and not commercially available for 6.78 MHz application, solid copper wire is used with wire radius denoted by $r_{r,0}$.

When placed in the B -field, the effective area that has a changing magnetic flux and contributes to the induced voltage is

$$A_{ind} = 4 \sum_{i=1}^{N_r} a_{ri} \cdot b_{ri} \quad (6.9)$$

With A_{ind} , the induced voltage is

$$V_{ind} = \omega_s B_0 A_{ind} \quad (6.10)$$

The coil inductance L_2 and ESR $R_{rx,wire}$ is calculated using the same modeling in the interleaved transmitter coil design part, replacing λ with 1. The circuit schematic of the Rx coil stage, together with the compensation capacitor and the equivalent impedance of the rectifier, is shown in Fig. 6.4.

The selection of C_2 requires the information of rectifier reactance. For each specified device, Z_{db} is determined by V_{rec} (6.8). Solving the circuit in Fig. 6.4, the compensation capacitor is

$$C_2 = 1/(\omega_s^2 L_2 + \omega_s \Im(Z_{db})) \quad (6.11)$$

The capacitor ESR $R_{rx,cap}$ is calculated according to the dielectric loss definition in the transmitter coil loss calculation, assuming the capacitor uses NP0 dielectric as the Tx coil

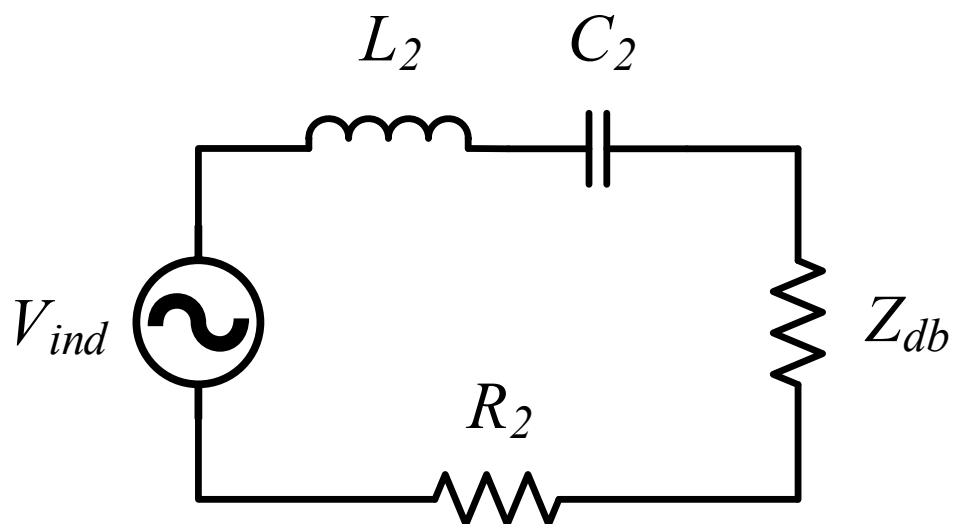


Figure 6.4: Circuit schematic of the Rx coil and compensation, with rectifier modeled as a equivalent impedance.

capacitors

$$R_{rx,cap} = t_g / (\omega_s C_2) \quad (6.12)$$

The total ESR of the Rx coil, including the compensation capacitor, is calculated by

$$R_2 = R_{rx,wire} + R_{rx,cap} \quad (6.13)$$

The power loss in the Rx coil is $P_{loss,rxcoil} = I_{rec,rms}^2 R_2$, where $I_{rec,rms} = I_{rec} / \sqrt{2}$ is the RMS value of i_{rec} .

Using the model, the C_2 and $P_{loss,rxcoil}$ can be calculated for each Rx coil design with specified diode device.

6.2.2 Receiver Coil Design

As shown in Fig. 3.3, the Rx coil is coupled with the transmitter side through the magnetic field described by B_0 , and couple with the rectifier through the load described by $[V_{rec}, I_{rec}, \phi_{rec}]$.

In the Rx coil design in isolation, coil geometries N_r , a_{rx} , and b_{rx} are iterated and the coil L_2 , $R_{rx,wire}$, and A_{ind} are calculated. Then, C_2 and $P_{loss,rxcoil}$ are calculated for each iterated coil design within a range of B_0 . Note that for each specified diode device, the relation between $[V_{rec}, I_{rec}, \phi_{rec}]$ are modeled in 6.1. Iterating among various diode devices yields different relations among $[V_{rec}, I_{rec}, \phi_{rec}]$. After this step, $P_{loss,rxcoil}$ is described using $[B_0, V_{rec}, device]$, where each design corresponds to a diode rectifier design. Multiple designs exist that exhibit various power losses. Only the design with the minimum power loss is stored in a look-up table for system-level design.

In the example, N_r is iterated from 2 to 4. Positions of a_{rx} and b_{rx} are iterated from 1 cm to maximum half length or width with a 0.2 cm step. The resulting Rx coil ESR $R_{rx,wire}$ and A_{ind} are presented in Fig. 6.5. For specified device of B340LB, the receiver loss is shown in Fig. 6.6. Iterating all devices, the receiver coil design space is shown in Fig. 6.7.

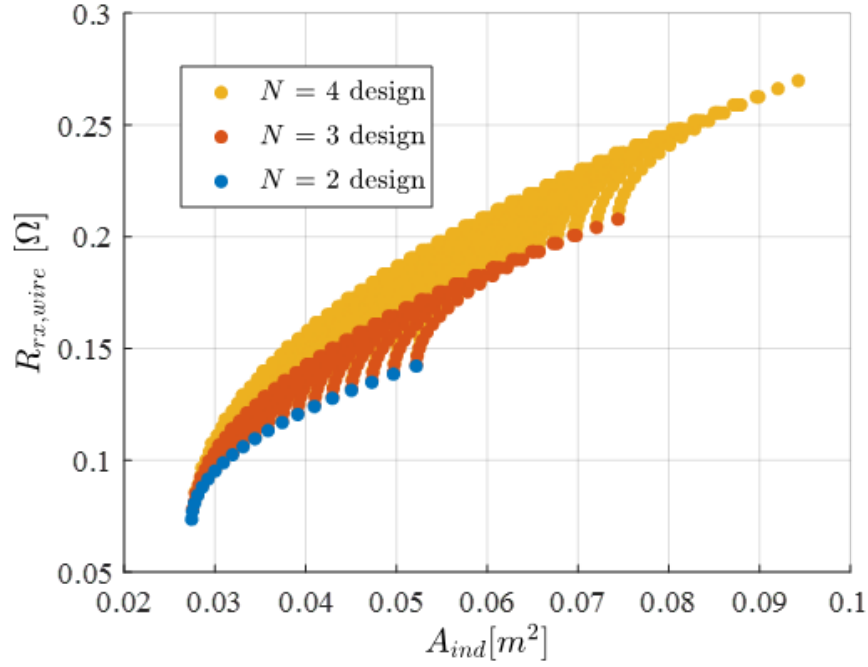


Figure 6.5: Receiver coil A_{ind} and wire ESR characteristics after geometric iteration.

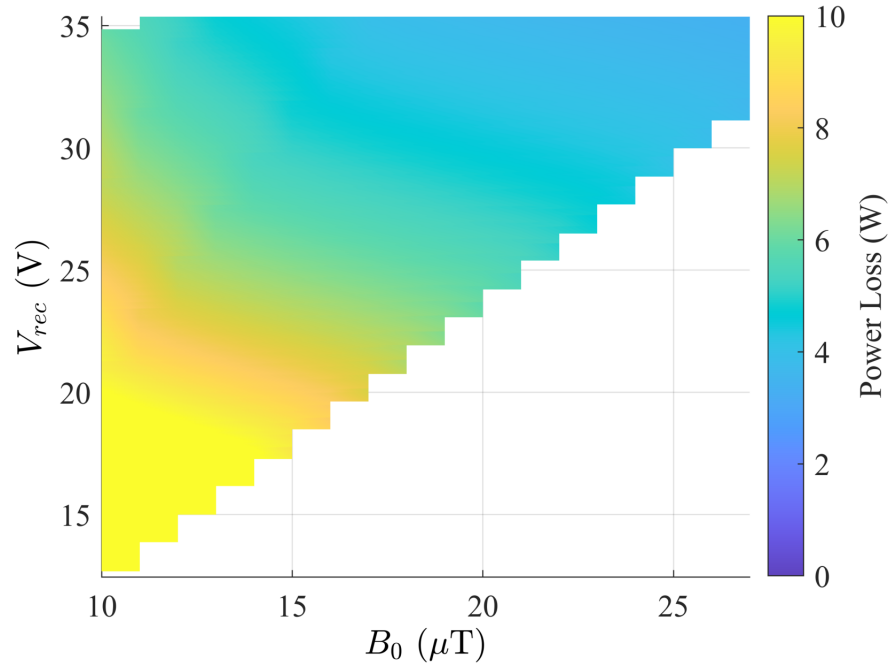


Figure 6.6: Receiver coil power loss and the required compensation for a selection of B_0 and V_{rec} . The diode in study is B340LB.

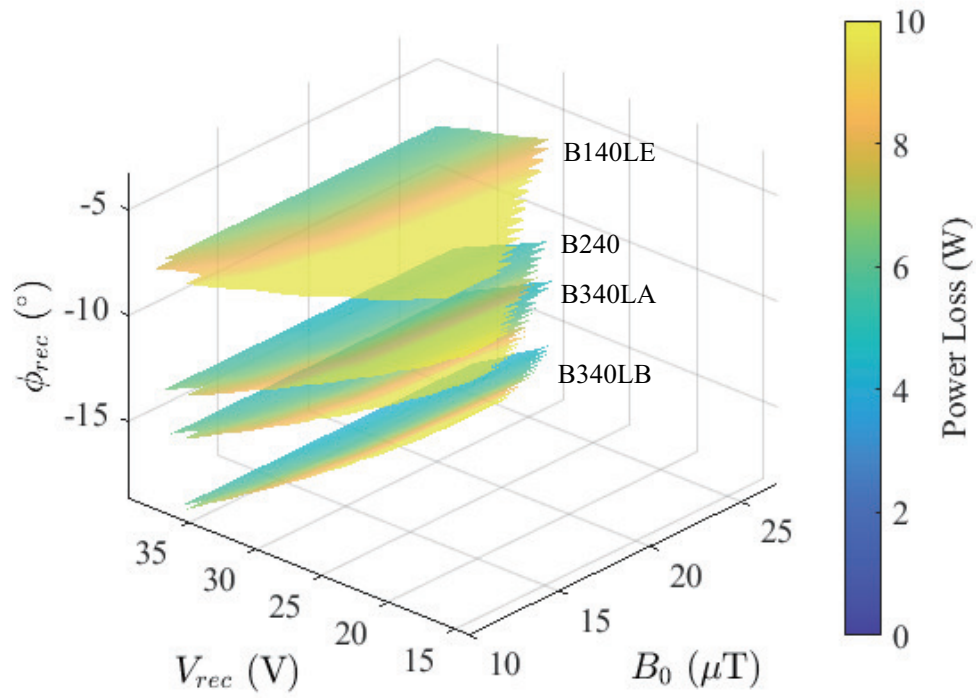


Figure 6.7: Receiver coil power loss described by B_0 and V_{rec} , ϕ_{inv} is to show the different capacitive phase when changing diodes.

6.3 Summary

This chapter details the receiver side modeling and design, including the high frequency rectifier reactance, Schottky diode device selection, and the receiver coil geometric design with compensation. Analytical models are provided. Following the overall design method, this chapter optimizes the each stage internally, leading to reduced design spaces parameterized by the coupling parameters of adjacent stages for system-level design.

Chapter 7

System-level Design and Experimental Verification

In previous chapters, the transmitter coil is designed to have a uniform magnetic field, the inverter is designed to have ZVS operation and high efficiency, the receivers have been designed to be resistive and high-efficiency. As the first step, each stage has been parameterized using the coupling parameters of adjacent stages and is optimized internally to the greatest extent possible. In this chapter, the overall design is performed to capture the dependency of each stage, based on the reduced-order models for the highest overall efficiency.

7.1 System-level Design

The goal of the WPT system-level design is optimizing all stages simultaneously to achieve high end-to-end efficiency while transferring full power. The additional design goals for multi-receiver operation are incorporated into the individual stage design methods through the design of the IMN and the constraint on κ , resulting in the designed field uniformity across spatial and power variations.

Because each stage has been converted to a reduced design space by the decoupled design, system-level optimization can be completed simply by an exhaustive search over the points of the parameterized subsystems: Inverter (Fig. 5.9), IMN (Fig. 5.11), Tx coil (Fig. 4.6),

Rx coil (Fig. 6.6) and rectifier (Fig. 6.3). The remaining design choices are all coupling parameters; thus, any choice of a parameter on one stage constrains the selections that can be made in other stages.

The coupled loss in the transmitter side is shown in Fig. 7.1. A trivial solution to the Tx coil is to design the coil near minimum $B_0 = 10 \mu\text{T}$ for minimum Tx coil loss. However, this design will result in a high power loss in the IMN stage, as shown by the red arrow that links the Tx coil with the IMN, and the red design curve in the IMN stage. In the IMN design, the red curve has higher power loss than the blue curve, which is linked with $I_t = 1.8 \text{ A}$ and leads to higher Tx coil power loss. Thus, the selection of the minimum IMN design and the minimum Tx coil design contradicts each other and needs overall consideration. Similarly, the IMN design that yields the minimum power loss in isolation (gray dot) is linked with an inverter design region that has high power loss. Each IMN design is linked with a single line of the ZVS inverter design, i.e. the IMN design marked by the blue dot constraints the inverter design to a gray line. The inverter design on the gray line has higher power loss than the minimum possible design. Each inverter design is also linked with a single curve on the IMN stage, thus influencing power loss in the rest of the transmitter side. Thus, a final, optimal transmitter design can only be selected when each stage is considered together with the models of the remainder of the system.

Considering all stages simultaneously, the system-level power loss is shown in Fig. 7.2, where B_0 is retained as a parameter for the plot. For each B_0 , multiple systems exist but only the system with minimum loss is presented. The optimal system design occurs at a magnetic field $B_0 = 20 \mu\text{T}$ and exhibits a power loss of 6.9 W. The optimized circuit parameters are summarized in Table 7.1.

7.2 Experimental Verification

To verify the WPT system modeling and design method, a 100 W, 6.78 MHz prototype WPT charging station is constructed using the optimized parameters. A picture of the prototype is given in Fig. 7.3. The transmitter coil fabrication and the uniform field test have been covered in Chapter 4. The PCB consists of all the components including the inverter,

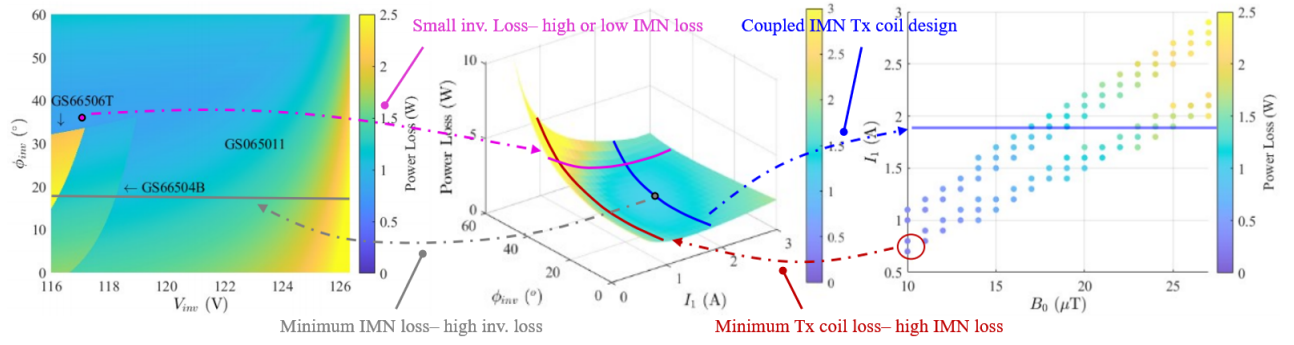


Figure 7.1: Coupled loss in the transmitter side.

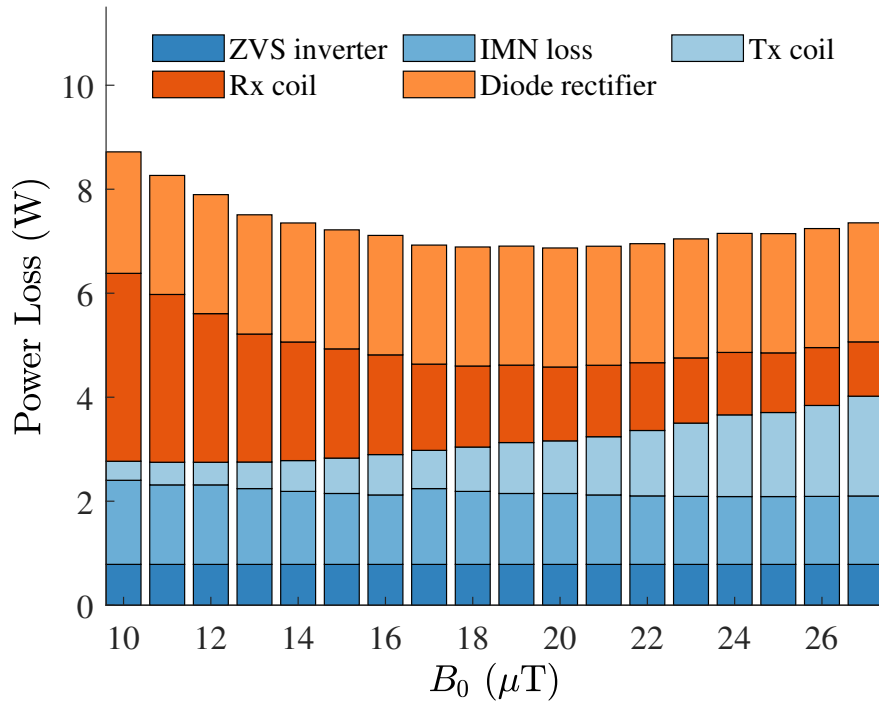


Figure 7.2: Power loss of the entire system.

Table 7.1: Comparison of measured and calculated passive components parameters.

Parameters	Variable	Design	Measurement
Tx coil inductance (μH)	L_1	12.3	12.6 (+2.4%)
Tx coil compensation (pF)	C_1	44.8	44 (-1.8%)
Tx coil ESR (Ω)	R_1	1.07	1 (-6.5%)
1 st Rx coil inductance (μH)	$L_{2,1}$	1.65	1.66 (+0.6%)
1 st Rx coil compensation (pF)	$C_{2,1}$	356	350 (-1.7%)
1 st Rx coil ESR (Ω)	$R_{2,1}$	0.17	0.18 (+6%)
2 nd Rx coil inductance (μH)	$L_{2,2}$	1.65	1.67 (+1.2%)
2 nd Rx coil compensation (pF)	$C_{2,2}$	356	350 (-1.7%)
2 nd Rx coil ESR (Ω)	$R_{2,2}$	0.17	0.18 (+6%)
IMN 1 st inductor (μH)	L_{imn1}	1.82	1.85 (+1.6%)
IMN capacitor (pF)	C_{imn}	303	300 (-1%)
IMN 2 nd inductor (μH)	L_{imn2}	0.5	0.48 (-4%)

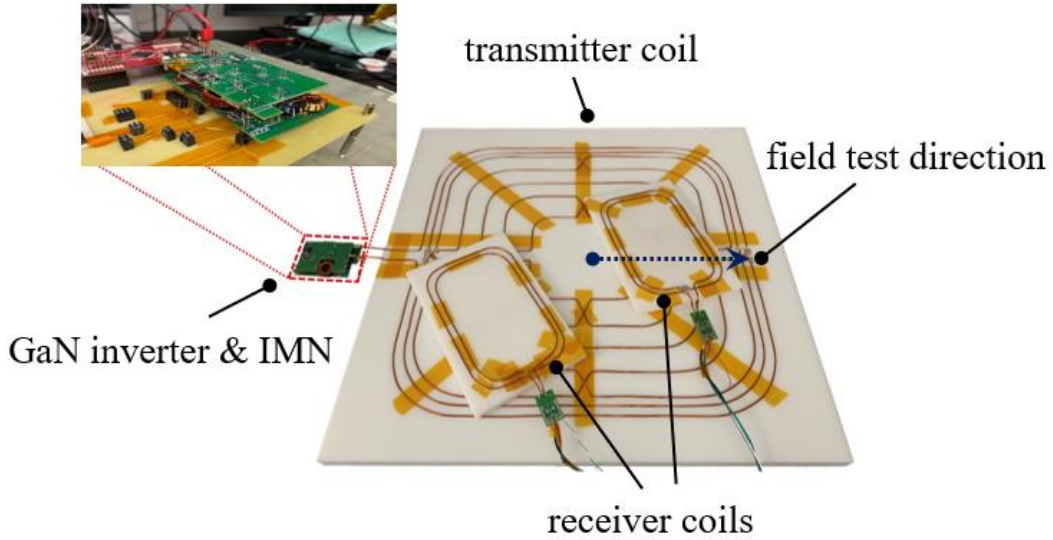


Figure 7.3: Picture of the prototype.

the IMN, the front-end PFC [73], sensors, and auxiliary power supply. All these stages are implemented into a small box with a volume of $7.36 \times 5.26 \times 1.68 \text{ cm}^3$ for 25 W/inch^3 power density. Only passive cooling is used in the experiments due to the high efficiency. A picture of the PCB design is shown in Fig. 7.4. Each receiver is connected to a BK8610 electronic load. The transmitter and receiver coils are both shown upside-down so the windings are visible.

An Agilent 4294A impedance analyzer is used to measure resistance and inductance for the coils and inductors. The measurement results are summarized in Table 7.1 and compared with the designed parameters, proving the accuracy of the modeling.

The uniform magnetic field property has been verified in Chapter 4.1. To verify the constant surface magnetic field and limited cross-channel disturbance, the prototype is measured under dynamically varying loading conditions. The load of the charging station is increased from zero to 100 W and then decreased back to zero. The load is changed by manually adding or removing the receivers on the table, which is representative of the real case of how customers use their devices. Experimental waveforms are shown in Fig. 7.5a. Measurements show that i_1 remains constant as the power level varies, and the two rectifier output voltages are free from disturbance as the alternate receiver is added to or removed from the system.

Zoomed-in waveforms of zero load, 50 W, and 100 W operation are shown in Fig. 7.5b, Fig. 7.5c, and Fig. 7.5d, respectively.

The system efficiency is defined as the total DC output power from all receivers divided by the transmitter side DC input power. Measured DC voltages and powers at the full load operating point are summarized in Table 7.2, together with a comparison to the model predictions. The measured power loss is 7.82 W, which is 6.8 % higher than the calculated value, proving the accuracy of the system modeling and design. The measured system efficiency is 92.8 %. The modeled loss breakdown at this operating point is shown in Fig. 7.6. The measured efficiency including the front-end PFC is 90.3 % [73].

Additionally, the power loss and efficiency are measured when the output power changes from 0 to 100 W with a step of 20 W. The power is changed by changing the magnetic coupling, as shown in the dynamic power test in Fig. 7.5, which represents a real situation

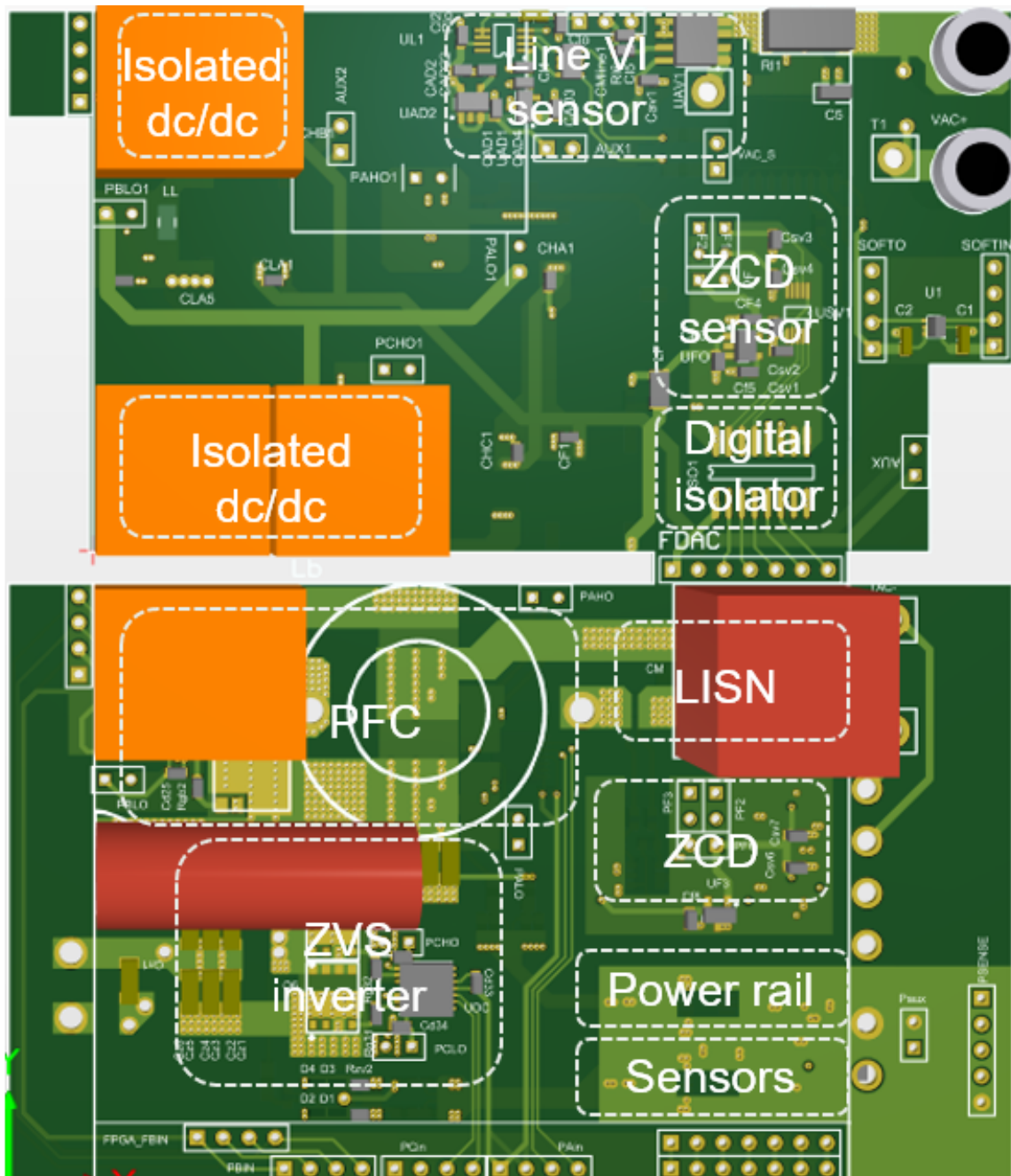
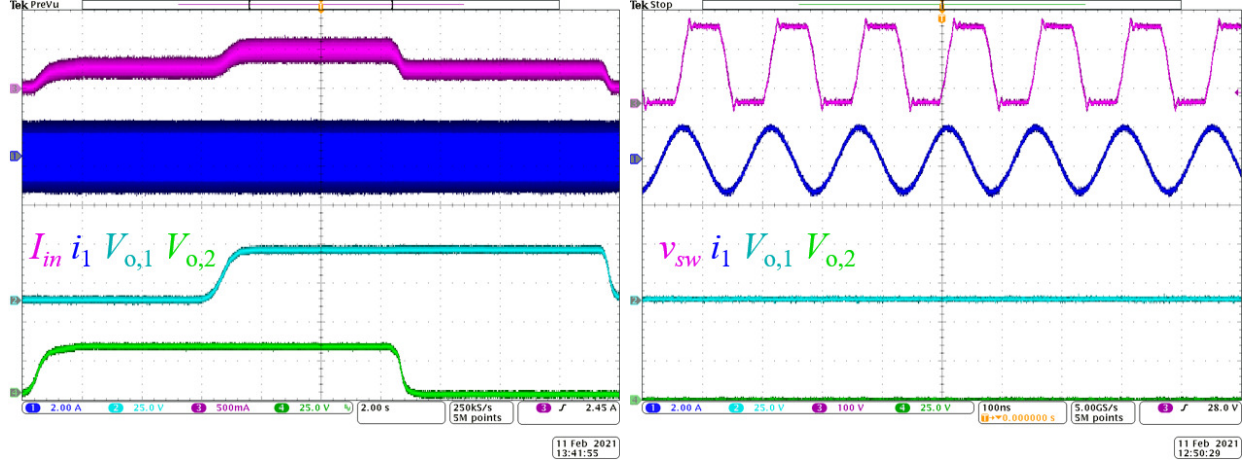
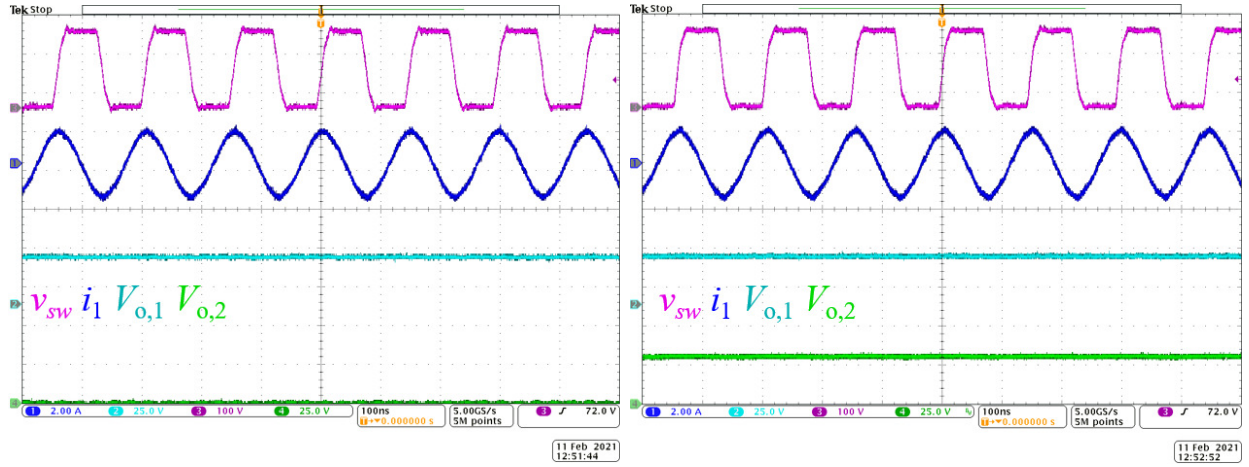


Figure 7.4: PCB design including the inverter, the IMN, the front-end PFC [73], sensors, and auxiliary power supply. The two boards will be implemented to a small box for 25 W/inch³ power density.



(a)

(b)



(c)

(d)

Figure 7.5: Working waveform showing limited cross-channel disturbance and constant field with changing load (a) overall waveform with changing load; (b) zoomed-in waveform of open load operation; (c) zoom-in waveform of 50 W operation; and (d) zoom-in waveform of 100 W operation.

Table 7.2: Comparison of measured and calculated circuit working parameters

Parameters	Variable	Design	Measure
Transmitter input voltage (V)	V_{dc}	200	200 (+0%)
Transmitter input power (W)	P_{in}	107.64	108.04 (+0.37%)
1st receiver output voltage (V)	$V_{o,1}$	31.5	31.72 (+0.7%)
1st receiver output power (W)	$P_{o,1}$	50	50.31 (+0.6%)
2nd receiver output voltage (V)	$V_{o,2}$	31.5	31.61 (+0.3%)
2nd receiver output power (W)	$P_{o,2}$	50	49.96 (-0.1%)

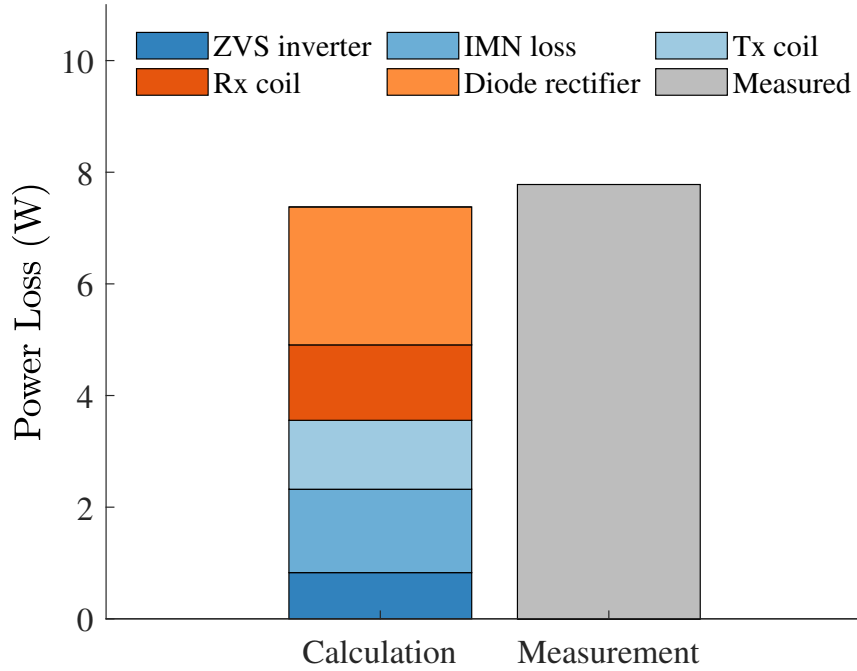


Figure 7.6: Calculated loss components contributing to the total conversion losses of the prototype system at near 100 W output power.

where users move their device up in the air, or charge around the edge of the table. At each power level, the two receivers are tuned to have similar output power. The end-to-end efficiency vs. power level is shown in Fig. 7.7. The system efficiency is higher than 82 % at 20 % load, and increases gradually to 92.8 % at full load. The measured system efficiency matches with calculation.

7.3 Summary

The goal of the WPT system-level design is to achieve high end-to-end efficiency at full power. In this chapter, it is shown that the power losses of each individual stages are coupled. The low-loss design of one stage may lead to high loss design of its adjacent stages. Thus, the system-level design requires considering all stages simultaneously. Using the proposed method, a WPT station was designed and fabricated. The prototype tests agrees with the designed result in both circuit operating point and power loss. The tested system efficiency is 92.8 %.

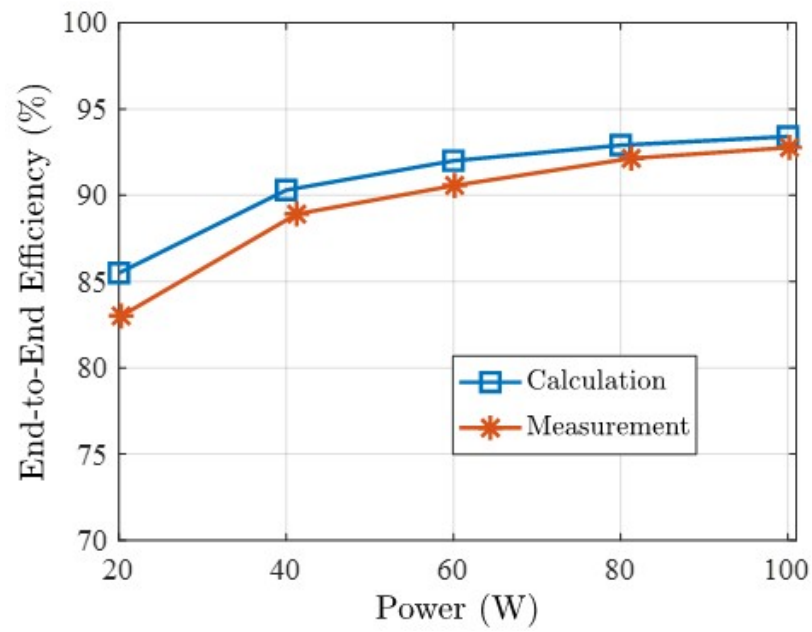


Figure 7.7: Efficiency vs. output power, from 20 % load to full load.

Chapter 8

Receiver Design in Consumer Electronics Application

In Chapter 6, the 50 W receiver for monitor application is designed at system-level and shows high efficiency. Fig. 8.1 is a picture showing that the receiver is planted in the plastic base of a computer monitor. The monitor was lighted up successfully with a 20 to 50 W dynamic power level.

This wireless charging monitor application has two characteristics:

- 1) target application is not sensitive to coil thickness. Thus, thick solid wire is used to enhance the conduction area and reduce power loss.
- 2) target application is not sensitive to leakage magnetic or electric field, as the neighboring object is plastic base and not interfering with the magnetic field.

In cellphone or laptop applications, new challenges arise. The first challenge is the height sensitivity, which prevents the use of thick solid copper wire to reduce power loss. The second challenge is that the target application is sensitive to leakage magnetic field and electric field. For example, the leakage magnetic field causes eddy currents in neighboring metal or PCB trace, potentially destroying the magnetic coupling with the transmitter side. As in Fig. 8.2, when the receiver was placed under a laptop, the induced voltage after rectification is only 1.3 V and the laptop cannot work. The electric field interacts with the device touch screen, causing “ghost finger” or false-triggering [28] issue. The leakage electric field also potentially



Figure 8.1: Prototype of a WPT monitor lighted up successfully.



Figure 8.2: Prototype of a laptop with metal case.

causes dielectric loss in neighboring materials, such as coil base, neighboring PCB, and ferrite shielding.

To protect magnetic coupling and reducing interference with neighbouring objects, ferrite shielding is necessary, which changes coil induced voltage and self-inductance and must be considered when designing the receiver coil. To reduce leakage electric field, discrete distributed capacitors may be used to compensate the voltage potential of each turns [28]. Fig. 8.3 is a picture showing the distributed connection of a commercial Airfuel receiver, where 6 capacitors are implemented. The repeated connection introduces additional parasitic ESR [50]. And the physical extent of the capacitor (typically 2x0.9 mm height for 0805 packages) is space-consuming.

This chapter presents the receiver coil and receiver side design for laptop applications. The ferrite impact is modeled and included. A novel self-resonant coil is proposed to achieve low profile, low E-field, and high Q . Modeling of the coil is proposed. The systematic design of the receiver is demonstrated. A prototype is presented that validates the coil function and modeling, and the systematic design result.

8.1 Proposed Coil Structure and Working Principles

Fig. 8.4 shows a 2-turn, proposed self-resonant coil schematic. The coil consists of two copper layers separated by one layer of dielectric material. The two copper layers are identical and stick to the dielectric laminate with no air gap between the three layers. The shape of the copper is carefully designed to provide resonant inductance. The surface area of the copper and the dielectric material is carefully designed to form the resonant capacitors between the two copper layers. Terminals a and b are connected to the ac source (e.g., WPT inverter).

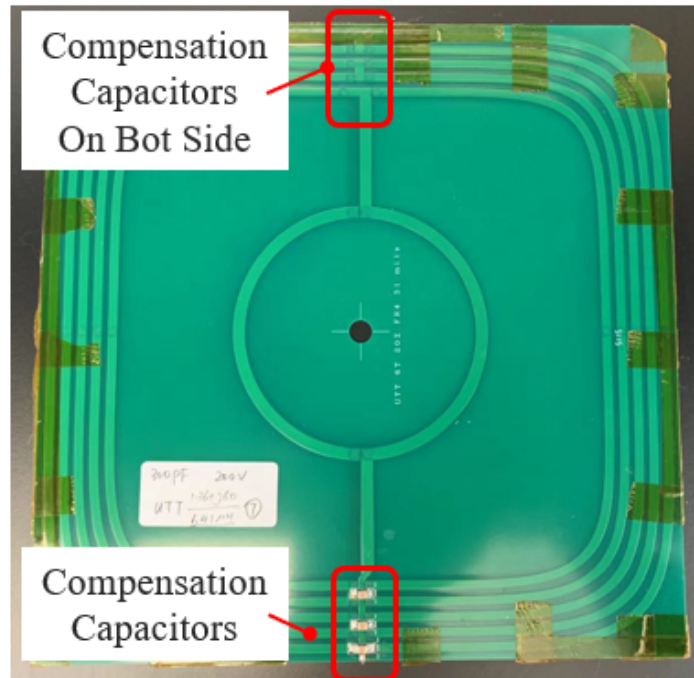


Figure 8.3: Picture of a commercial coil, showing the repeated distributed compensation.

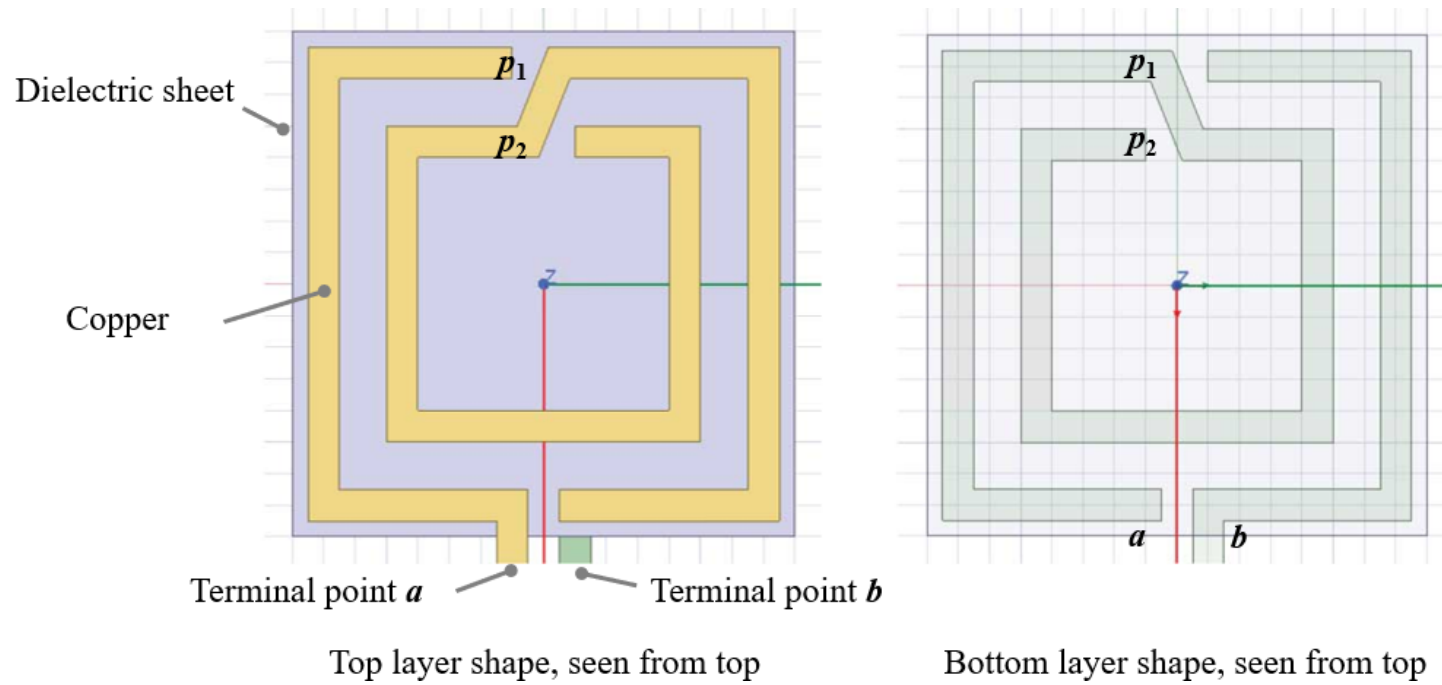


Figure 8.4: Diagram showing the structure of the proposed coil.

Due to the complete separation of the two conduction layers by the dielectric, the coil is an open circuit for a DC source. When an AC source is applied, current flows from one terminal to the other crossing through the dielectric a couple of times, resulting in a repeated series LC connection, and the total capacitance consists of three distributed parasitic capacitance in series. The three parasitic capacitor sections are: 1) point a to p_1 , 2) point p_1 to p_2 , and 3) point p_2 to b .

Similar to a parallel-plate capacitor, the current transients between two layers in the form of displacement current $J_d = \varepsilon_r \varepsilon_0 \partial E / \partial t$ where ε_r is the relative permittivity of the dielectric material. According to [56], the electric field between the two copper layers is constant within each capacitor section, which leads to the uniform distribution of J_d along the length and width of the trace in each section.

Due to the uniform J_d in each capacitor section, the input current linearly transitions from the input terminal a on the top to the end of the first half turn, to the bottom layer at point p_1 . At point a and p_1 , the entire coil current flows through one of the conductors, with zero in the opposite conductor. Then, the current linearly transitions from the p_1 on the bottom conductor to the end of the inner turn to the top layer at point p_2 . Then, the current linearly transitions from the p_2 on the top conductor to the end of the outer turn to the bottom layer at point b . At point p_2 and b , the entire coil current flows through one of the conductors, with zero in the opposite conductor. The proposed current distribution on the two conductor layers is shown in Fig. 8.5, showing distributed series LC characteristics.

Full-wave FEA simulation (Ansys HFSS) is used to verify the current distribution pattern. The results are shown in Fig. 8.6, which agrees with the analyzed current distribution.

In Fig. 8.4 each half turn is compensated by a parasitic capacitor forming a fully compensated self-resonant coil (FSRC). This structure potentially can be combined with conventional planar coil to form a new structure that compensates every other turn, forming a hybrid self-resonant coil (HSRC). Fig. 8.7 shows a three-turn example where the voltage potential of the outermost and innermost turn is compensated by parasitic capacitors, but the middle turn is not compensated.

FSRC has repeated serial capacitance. Thus the total capacitance is small and may be difficult to acquire the target capacitance given geometry constraint. This is a key factor

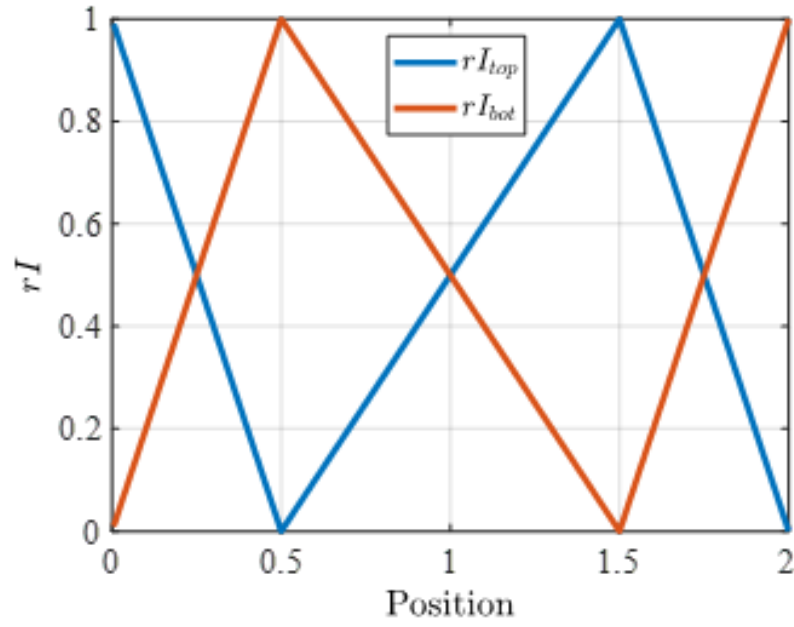


Figure 8.5: Diagram showing the current distribution along the longitude change of coil length.

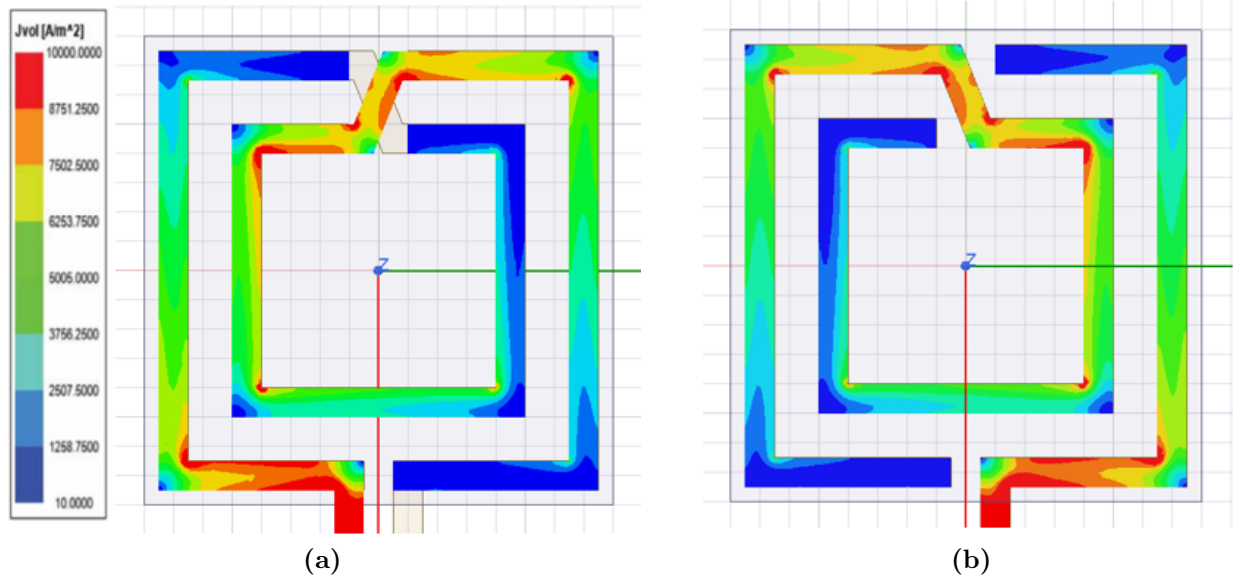


Figure 8.6: Simulated current density distribution (a) top view, and (b) bottom view.

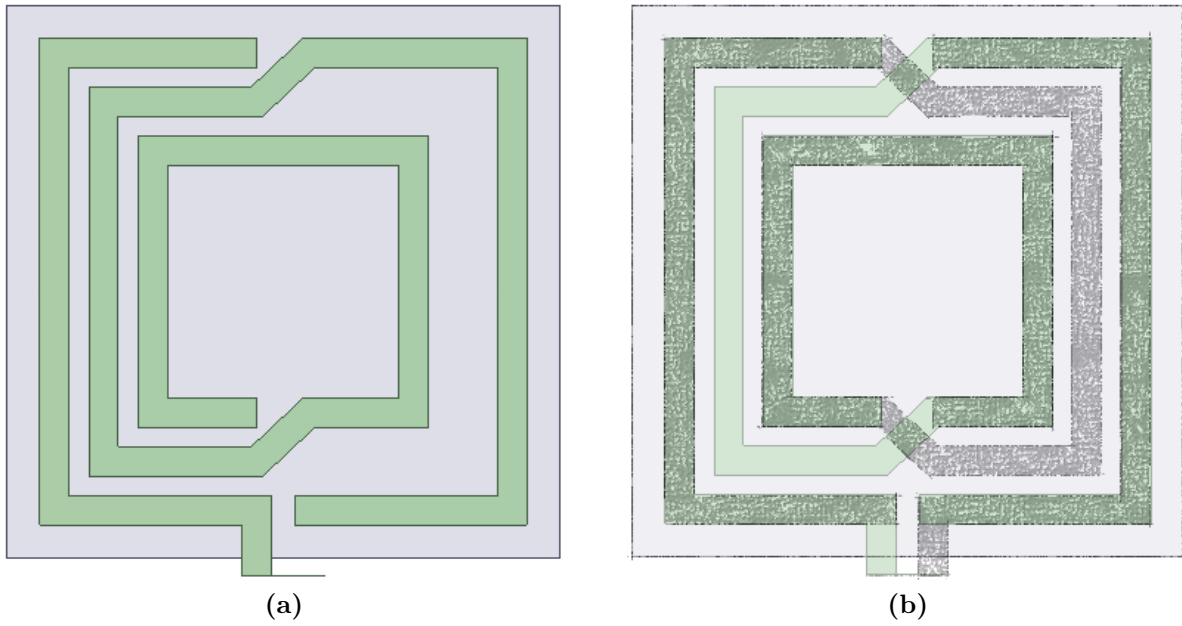


Figure 8.7: HSRC structure top view (a), and top view with transparent dielectric (b). The dashed grey trace represents the bottom layer seen through from the top.

why FSRC prefers thin dielectric compared to other self-resonant coils without repeated serial connections. Compared to FSRC, HSRC has a reduced number of serial connections, thus the maximum achievable capacitance is larger than FSRC with the same dielectric and geometry. Due to the reduced number of compensation, the leakage electric field for HSRC is larger than FSRC but is smaller than conventional coils.

It should be noted that the FSRC and HSRC could be designed in a circular shape, per various application requirements, which have the same working principle as the rectangular coil.

8.2 Coil LCR Modeling

The geometric parameters of the coil are shown in Fig. 8.8. w_r is the width of each turn, l_i is the inner length, and l_o is the outer length. n_s is the number of turns of one layer, and h_r is the thickness of the dielectric layer. t is the copper thickness. Additionally, the dielectric is described by Dk , h and tg .

To examine the performance capabilities, analytical models for the inductance, capacitance and resistance are developed based on results from the literature and FEA-assisted simulations. L is analyzed based on magnetic field simulation. C and R are analyzed based on comparison to existing self-resonant coils.

8.2.1 Inductance

The top and bottom layers of the proposed coil have identical current flow directions (i.e. from input terminal towards output terminal). Compared to a conventional PCB coil, the current flow in the proposed coil differs only in that it crosses vertically through the dielectric layer. Both layers have identical current flow direction (When the dielectric thickness is thin relative to the width, the magnetic flux distribution of the proposed self-resonant coil and the traditional PCB coil are nearly identical. Fig. 8.9 is a magnetic field simulation comparing the flux distribution between a conventional coil 8.9a and a self-resonate coil 8.9b, assuming both coils conduct 1 A 6.78 MHz current. Due to the identical flux distribution, the two coils have the same inductance.

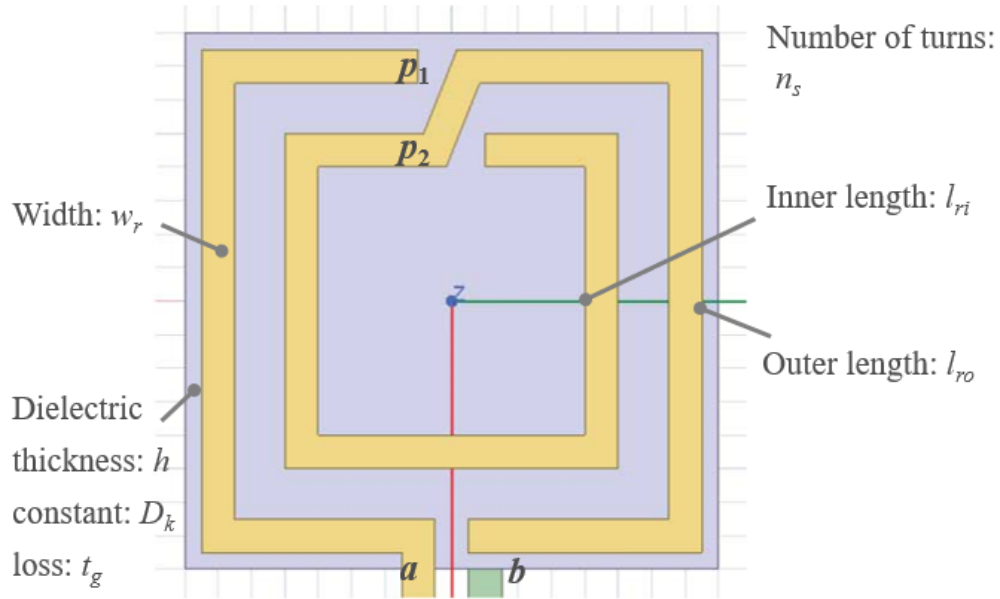


Figure 8.8: Geometry of the proposed coil.

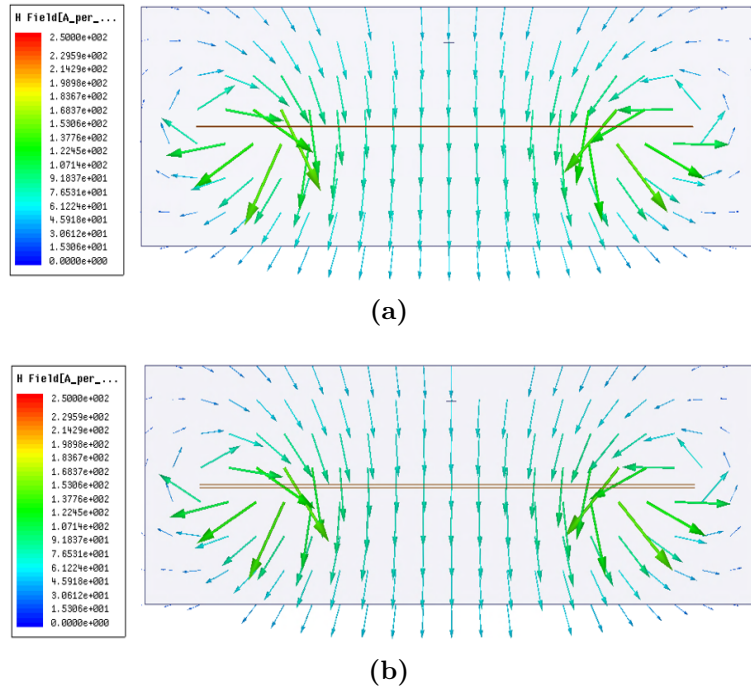


Figure 8.9: Flux distribution of traditional coil (a), and the proposed self-resonant coil (b) when conducting 1 A, 6.78 MHz current .

The inductance of a conventional planar coil was reported in [64] and exhibits an error of less than 3 %. The inductance equation is used to describe the proposed self-resonant coil

$$L_s = \frac{1.27\mu n^2 d_{avg}}{2} \left(\ln \frac{2.07}{k_u} + 0.18k_u + 0.13k_u^2 \right) \quad (8.1)$$

where μ is the magnetic permeability, d_{avg} is the average length $d_{avg} = (l_o + l_i)/2$, $k_u = (l_o - l_i)/(l_o + l_i)$. The inductance equation is re-written as

$$L_s = \frac{1.27\mu n^2 (l_i + l_o)}{4} \left(\ln \frac{2.07(l_i + l_o)}{l_o - l_i} + 0.18 \frac{l_o - l_i}{l_o + l_i} + 0.13 \left(\frac{l_o - l_i}{l_o + l_i} \right)^2 \right) \quad (8.2)$$

8.2.2 Capacitance

Compared to a conventional self-resonant coil, FSRC has the same parallel-plate structure in each capacitor section of each half-turn. Thus, the capacitance of each section is modeled using the same method as a CSRC, which has been detailed in [56].

The i th turn capacitance of a CSRC is [56]

$$C_{csr,i} = \frac{D_k \varepsilon_0 w_r 4l_i}{h_r} \left[1 + \frac{h_r}{\pi w_r} \ln \frac{2\pi h_r}{w_r} + \frac{h_r}{\pi w_r} \ln \left(1 + \frac{h_r}{2w_r} + 2\sqrt{\frac{t}{h_r} + \left(\frac{t}{h_r} \right)^2} \right) \right] \quad (8.3)$$

For FSRC, the total capacitance is the serial connection of each half turn and the innermost turn

$$C_{fsrc,s} = \frac{1}{\sum_{i=2}^{n_r} \frac{4}{C_{csr,i}} + \frac{1}{C_{csr,1}}} \quad (8.4)$$

For HSRC, the total capacitance is the serial connection of each half-turn that has a self-resonant structure, excluding the non-self-resonant turns. For HSRC, if the self-resonant turn is the odd number of turns, the capacitance is

$$C_{hsrc,s} = \begin{cases} \frac{1}{\sum_{i=2}^{n_r/2} \frac{4}{C_{csr,(2*i-1)}} + \frac{1}{C_{csr,1}}} & , \text{ } n_r \text{ is even number} \\ \pm \frac{1}{\sum_{i=2}^{(n_r+1)/2} \frac{4}{C_{csr,(2*i-3)}} + \frac{1}{C_{csr,1}}} & , \text{ } n_r \text{ is odd number} \end{cases} \quad (8.5)$$

8.2.3 Resistance

The total loss of the proposed self-resonant coil consists of copper loss and dielectric loss. The copper loss is modeled as skin-effect loss plus proximity effect loss [72]. The skin effect loss is calculated through the integration of the loss density over the whole coil. The proximity loss is through the calculation of the proximity field on each turn, and calculation of the proximity loss afterward.

As has been discussed, the input current linearly transitions from the top to bottom spiral over the whole length in each capacitor section, which is the same as in CSRC. Thus, skin-effect ESR of the i th turn is, if configured as a self-resonant structure [56]

$$R_{skin,i} = \frac{2\rho_{copper}4l_i}{3w_r\delta(1 - e^{-\frac{t}{\delta}})} \quad (8.6)$$

If the i th turn is configured as a conventional structure as in a HSRC, the skin-effect ESR is

$$R_{skin,i} = \frac{\rho_{copper}4l_i}{w_r\delta(1 - e^{-\frac{t}{\delta}})} \quad (8.7)$$

The total skin-effect ESR is $R_{skin} = \sum i = 1n_r R_{skin,i}$.

In addition to the skin effect, the time-varying H-field around the coil trace causes eddy current loss in the copper foil. Since the H-field in the FSRC coil has an almost identical H-field compared to a conventional coil (as shown in Fig. 8.9), the magnetic field distribution and the proximity related ESR are calculated using the same method as in a CSRC [56].

H_c is the H-field strength at the center point. H_{in} is the H-field strength at the innermost point. H_{out} is the outermost field. The H-field strength drop on each turn $dH = (H_{in} - H_{out})/n$. They are calculated by [56]

$$H_c = \frac{nI_{in}}{(l_o - l_i)} \ln \frac{\sqrt{l_o^2 + t^2} + l_o}{\sqrt{l_i^2 + h^2} + l_i} \quad (8.8)$$

$$H_{in} = H_c \exp^{\frac{l_i}{l_o} (0.4 + 0.15 \ln \frac{l_o}{4t + 2h})} \quad (8.9)$$

$$H_{out} = -H_c \left(0.4 + 0.08 \ln \frac{l_o}{4t + 2h} \right) \exp^{\frac{l_i}{l_o} (1 + 0.125 \ln \frac{l_o}{4t + 2h})} \quad (8.10)$$

After obtaining the field distribution information, the proximity effect loss is calculated using the standard formula for eddy-currents in a lamination [72]

$$P_{prox,i} = \frac{B_{avg,i}^2 w_s^2 t^2}{24\rho} Vol_i \quad (8.11)$$

where Vol_i is the copper volume of the i th turn, and $B_{avg,i}^2$ is the average peak flux density square of the proximal H-field

$$Vol_i = 4(l_{i,in}^2 - l_{i,out}^2)t \quad (8.12)$$

$$B_{avg,i}^2 = \int_0^{w_r} \frac{\mu(Hl(i) + dH/w_r dl)^2}{w_r} = \mu^2 (Hl(i)^2 + (Hl(i) - Hr(i))^2 / 3 + (Hl(i) - Hr(i))Hl(i)) \quad (8.13)$$

The proximity effect ESR of i th turn is

$$R_{prox,i} = \frac{2 \sum_{i=1}^n P_{prox,i}}{I_{in}^2 / 2} \quad (8.14)$$

The total proximity-effect related ESR is $R_{prox} = \sum i = 1 n_r R_{prox,i}$.

The dielectric loss is calculated based on the loss tangent D_k of the dielectric material

$$R_c = \frac{D_k}{2\pi f C_s} \quad (8.15)$$

Finally, the total equivalent series resistance (ESR) of the coil is

$$R_s = R_{skin} + R_{prox} + R_c \quad (8.16)$$

8.3 Magnetic Shielding Effect

As mentioned previously, the neighboring metal may destroy magnetic coupling and requires ferrite shielding, which impacts the receiver coil inductance and induced voltage. To facilitate the proposed self-resonant coil design, the FEA simulation (Ansys Maxwell 2D) method is used to quantify the impact. The simulation schematic is shown in Fig. 8.10, which includes the transmitter coil with a uniform magnetic field. Due to the uniform magnetic field, the coupling is constant if the receiver coil is placed on other positions on the transmitter surface,

or if the transmitter coil is configured with other geometries as long as the field is uniform. The ferrite is a high-frequency, low-loss material [6] with a permeability of 120. The metal is a 2 oz copper layer representing a 2-layer PCB. The length of the coil is 17.8 cm (7 inch). And the height limit is 1.02 mm (0.04 inch).

In the simulation, the geometry of the coil and the thickness of ferrite are swept to evaluate the shielding effect of various ferrite thicknesses. The inner radius is swept from 3 cm to 6 cm. The number of turns is swept from 2 to 3. And the width is designed for each geometry leaving a 2 mm gap between adjacent layers. The outer length is fixed at 8.9 cm (3.5 inch) for application requirement. The inductance and induced voltage are evaluated and compared for two situations: 1) with ferrite only, and 2) with ferrite+copper. The results are shown in Fig. 8.11

In Fig. 8.11a, the inductance ratio is the inductance with ferrite only L_f over the inductance of the coil without any shielding L_0 , or the inductance with ferrite and copper $L_{f_{cu}}$ over L_0 . In Fig. 8.11b, the voltage ratio is the induced voltage with ferrite only $V_{ind,f}$ over the inductance of the coil without any shielding V_{ind} , or the inductance with ferrite and copper $V_{ind,f_{cu}}$ over V_{ind} . Without the copper layer, the ferrite shielding enhances the magnetic field and thus increasing the coil inductance for all ferrite thickness. The ferrite also forms a low impedance loop that facilitates more magnetic flux penetrating the Rx coil, leading to an increased induced voltage. With the copper layer, the part of the magnetic flux that penetrates through the ferrite causes eddy currents and is canceled, potentially leading to a reduced inductance and induced voltage if the ferrite is thin. Increasing the ferrite thickness reduces the penetrating flux to the copper and thus increasing inductance and induced voltage. Considering that 1 mm ferrite is sufficient to reduce $L_{f_{cu}}/L_f$ to 5 % and yields a voltage ratio of 0.95, and the improvement becomes increasing slow after this thickness, a ferrite thickness of 1 mm is selected, which is also approaching the height limit of 1.02 mm. Both voltage and inductance ratios will be used in the systematic design.

8.4 Receiver Coil Design

The new receiver coil is designed following the proposed systematic design method.

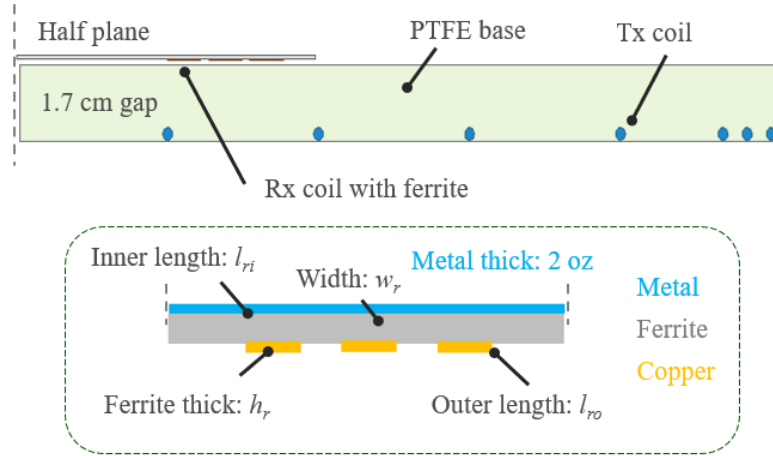


Figure 8.10: Schematic showing Maxwell ferrite simulation [20].

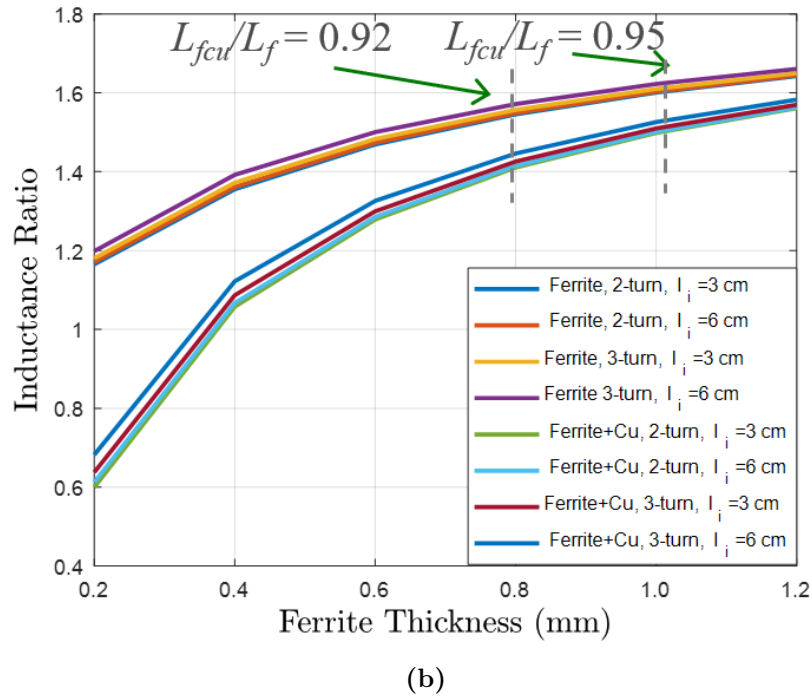
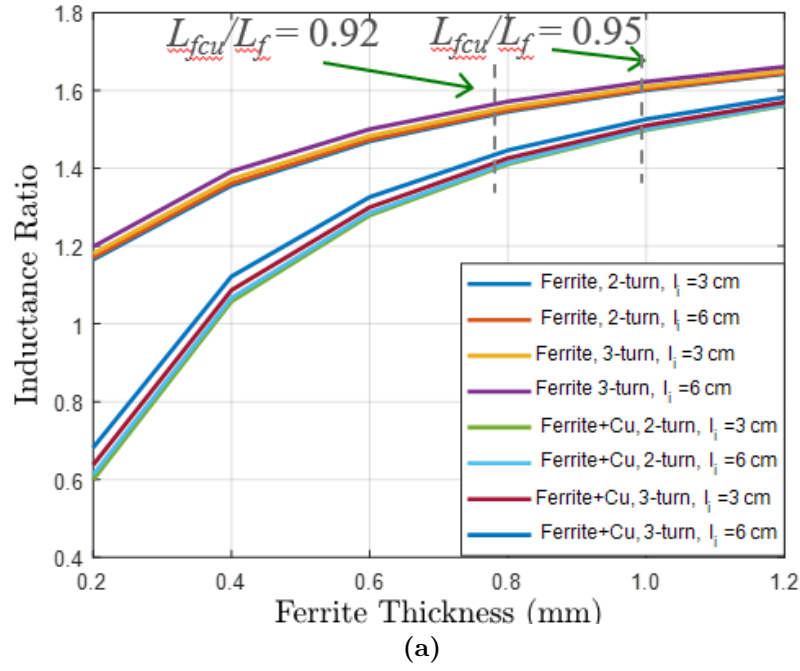


Figure 8.11: Simulated ferrite impact on inductance (a), and induced voltage (b).

In the internal stage design, geometric iteration is used to calculate coil circuit parameters under application geometric requirements. Four types of coils are compared 1) solid copper coil, 2) conventional self-resonant coil (CSRC), 3) HSRC, and 4) FSRC. The modeling of 1) has been detailed in Chapter 6. The modeling of 2) was reported in [56]. Sweeping geometries within application requirement, the designed circuit parameters for the solid coil are shown in Fig. 8.12. Note that the thickness is 0.2 mm.

Fig. 8.12a shows the LCR parameters, where R is represented by Q for a better sense of the coil quality. The dashed blue curve is 6.78 MHz LC resonance curve. The red curve is the LC design that cancels reactance for a 50 W rectifier using B340LB. The interested induced voltage area, which is also determined by the rectifier stage, is highlighted by the yellow box. Fig. 8.12a is the design result of LC vs A_{ind} , where A_{ind} is replaced by induced voltage when the coil is placed in 20 μ T magnetic field. The specific field value, device, and output voltage are shown with the same value as the optimal design in Chapter 7 to show that the SC coil has very low efficiency.

The design result for CSRC, HSRC, and FSRC are shown in Fig. 8.13, all resulting from geometric iteration.

The x symbols in Fig. 8.13 and Fig. 8.12 are designs with 1.60 μ H, 360 pF, and 36 V induced voltage, which is close to the fabricated monitor coil (1.65 μ H, 355 pF). A set of coil schematic corresponding to the x symbols are shown in Fig. 8.14.

With limited thickness, solid copper can only use thin wire and have limited conduction area, thus limiting the Q . CSRC configures the capacitors of every turn in parallel, requiring a limited capacitance from each turn, thus resulting in the thin width design. As a comparison, HSRC and FSRC configure the capacitance of each turn (if any) in serial, expanding the required capacitance of each turn, thus resulting in a wider trace than CSRC.

Combining four coils, the resulting LCR design space is shown in Fig. 8.15a only showing the minimum ESR. The type of coil is shown in Fig. 8.15b. Number 1 to 4 represents SC, CSRC, HSRC, and FSRC. The reduced design space will be used for a system-level design using the same method in chapter 6.

The system-level design space of the receiver coil for 50 W B340LB rectifier is shown in Fig. 8.16a, with projections on the XY plane, showing a similar trend as the previously

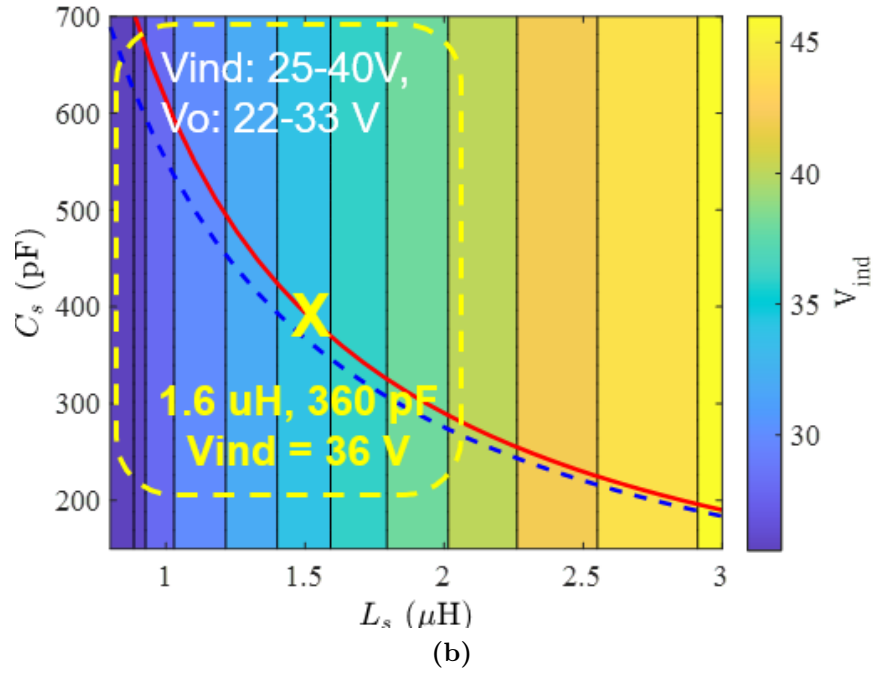
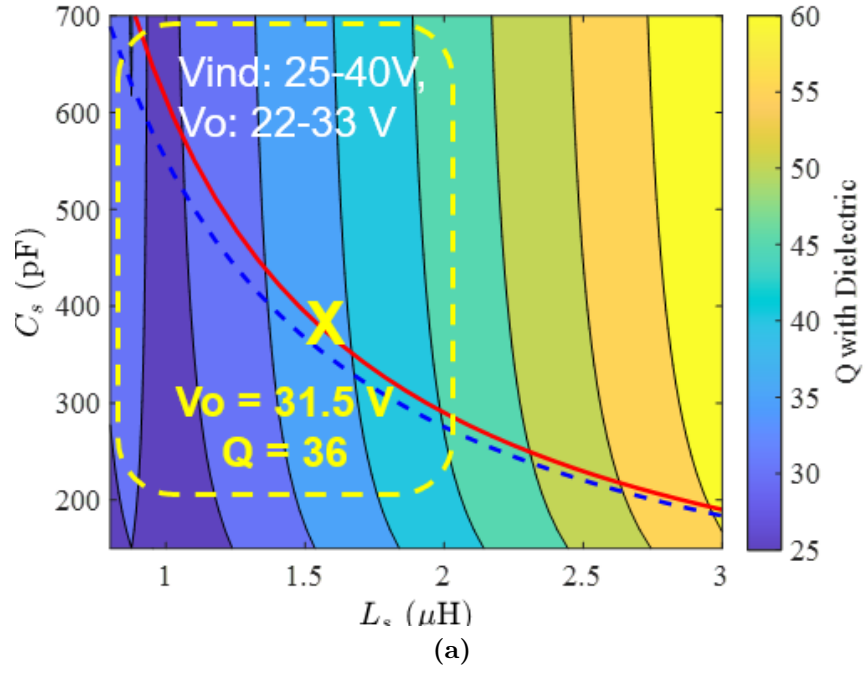


Figure 8.12: Calculated coil circuit parameters for SC, (a) LC vs R (represented by Q , and (b) LC vs A_{ind} (represented by V_{ind} at $20\text{ }\mu\text{T}$ field).

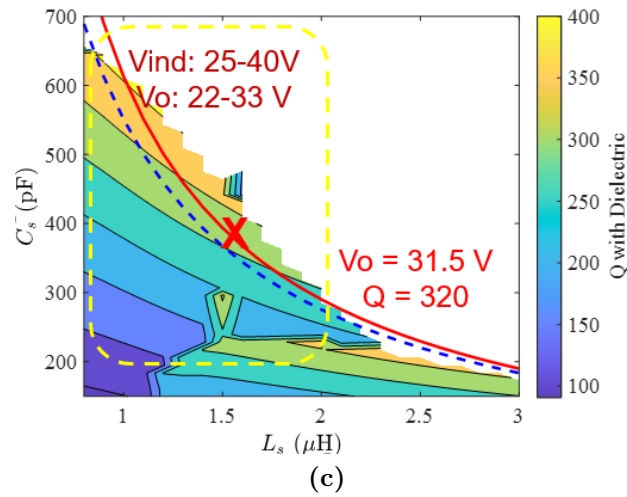
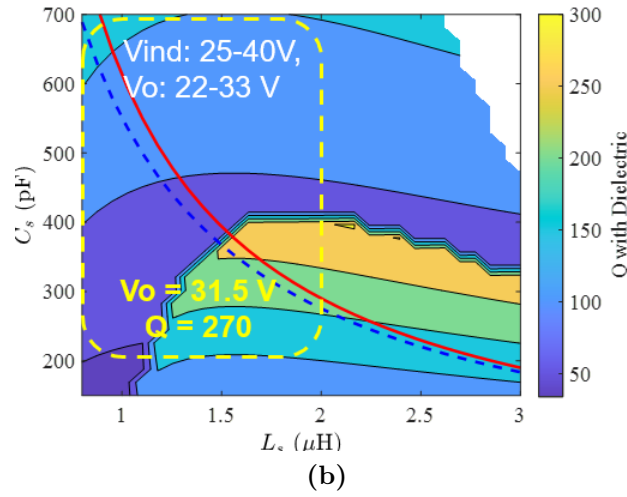
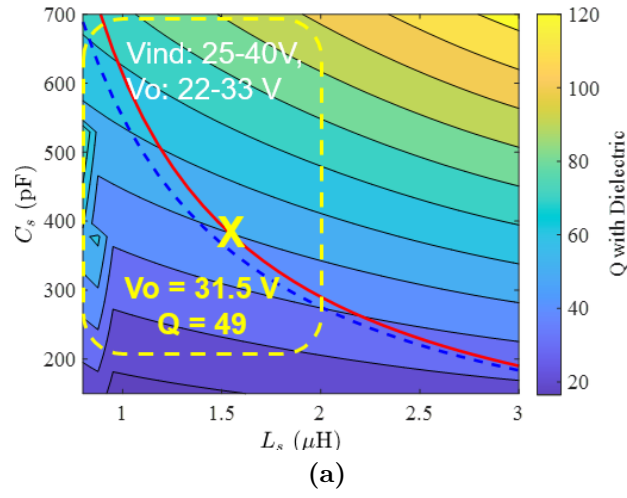


Figure 8.13: Calculated coil circuit parameters for (a) CSRC, (b) HSRC, and (c) FSRC.

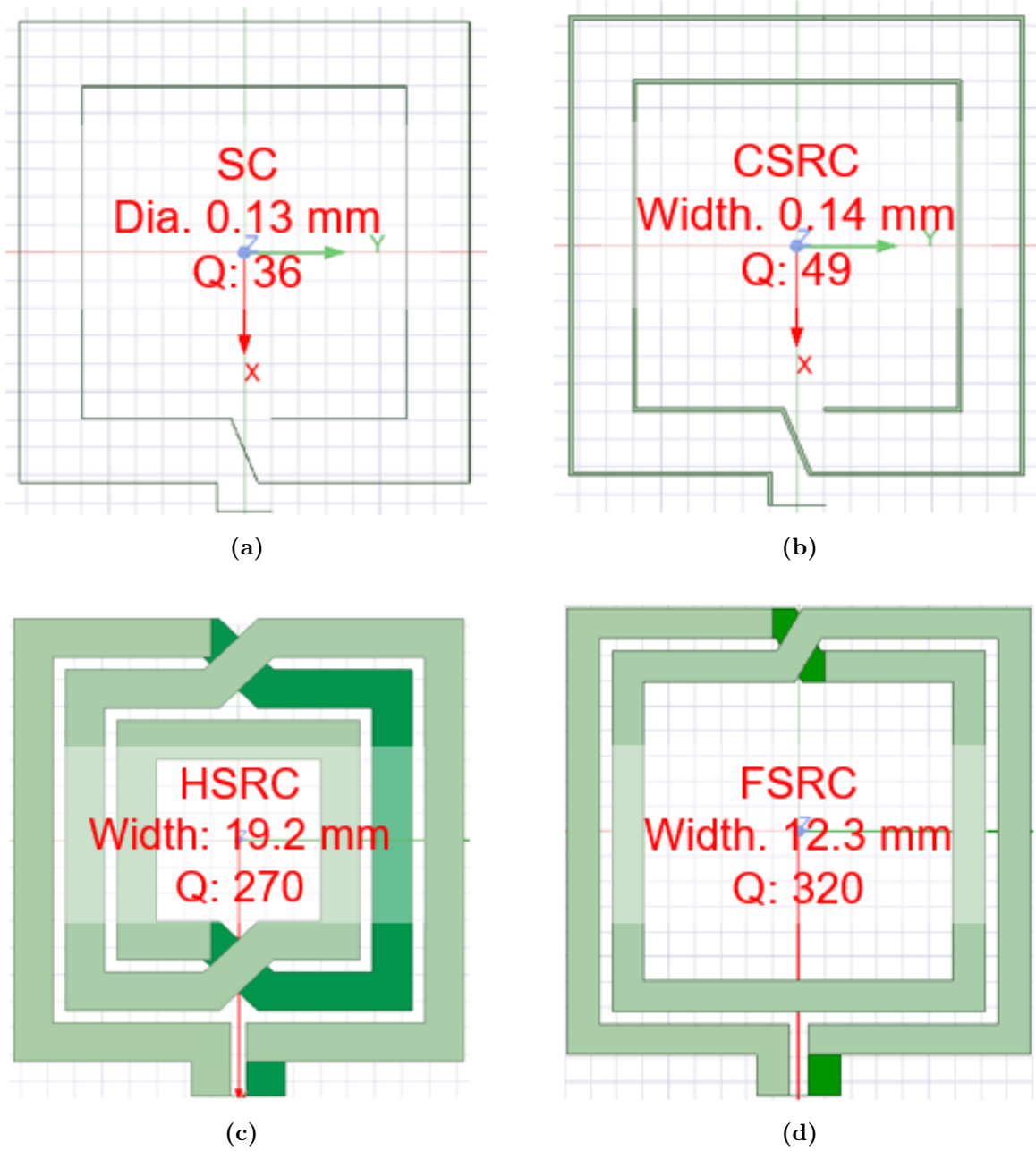


Figure 8.14: Schematic showing 4 coil structures corresponding to x marker, SC (a), CSRC (b), HSRC (c), and FSRC (d).

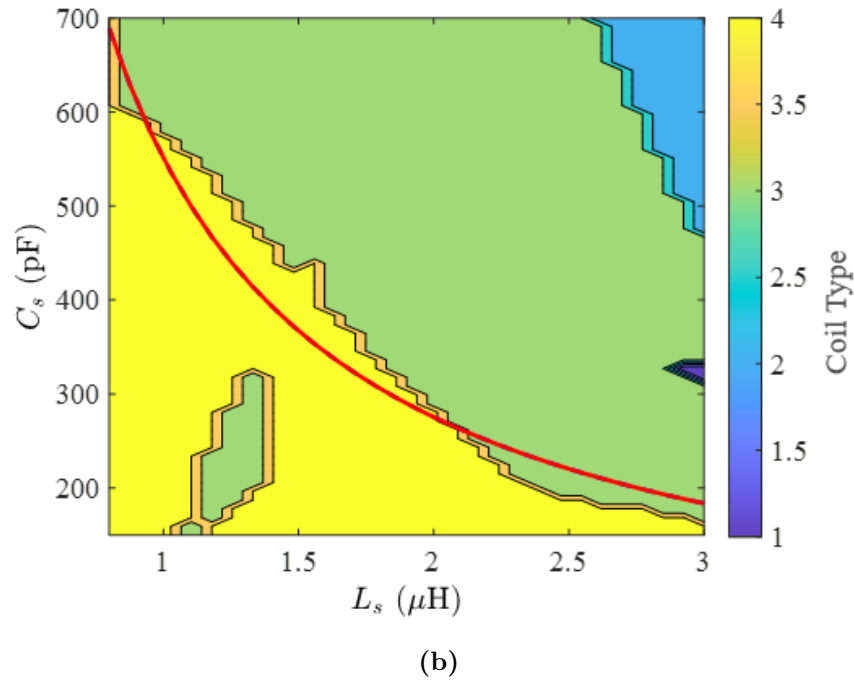
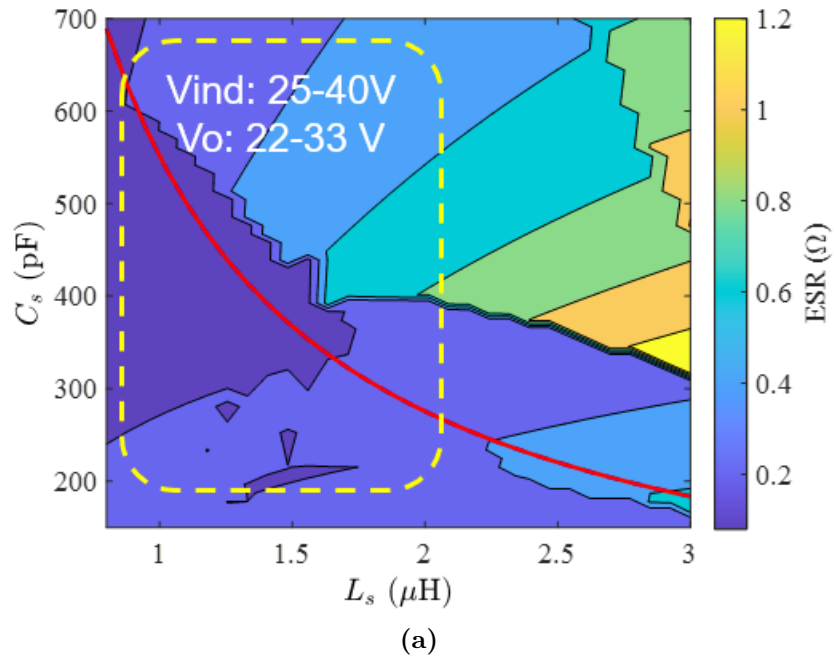


Figure 8.15: Minimum ESR for each LC combination (a), and the corresponding type of coil (b)

analyzed receiver without ferrite. The system-level design space including diode selection is shown in Fig. 8.16b, which will be used in systematic optimization design.

The receiver is design from system-level. The system-level design result showing the power loss on every individual stage is shown in Fig. 8.17, which follows a similar trend compared to the previous two monitor receiver design. The optimal design is $B_0 = 21 \mu\text{T}$ and yields a power loss of 6.94 W. Note that the fabricated transmitter is configured at $B_0 = 20 \mu\text{T}$ and results in a close power loss of 6.99 W. The receiver will be optimized for $B_0 = 20 \mu\text{T}$ considering the tiny difference to the optimal point.

The design result for the transmitter and monitor receiver are the same as the previously optimized result in Chapter 7. The design result for the laptop receiver are: $L_s = 1.6 \mu\text{H}$, $C_s = 360 \text{ pF}$, $\text{ESR} = 0.18 \Omega$, and $V_o = 31.5 \text{ V}$. B340LB is selected and the target V_o is 31.5 V. The receiver coil structure is selected to be FSRC, with $l_i=5.47 \text{ cm}$, $l_o = 8.89 \text{ cm}$, $w_r = 1.23 \text{ mm}$, and $n_r = 2$.

8.5 FEA Simulation Result

The proposed coils are simulated using Ansys HFSS. The top view of the schematic in the simulation is shown in Fig. 8.18. The bottom layer is the same as the top layer.

The impedance curve of the FSRC coil is shown in Fig. 8.19 directly imported from HFSS, showing serial LC. The y axis is the impedance. The x-axis is the frequency range from 5.5 to 8.5 MHz. The simulated resonant frequency is 6.60 MHz (-0.5 %). Another simulation was done at 1 kHz, extracting the capacitance to be 378 pF(+5 %). From the simulated resonant frequency and capacitance, the inductance is calculated to be $1.54 \mu\text{H}$ (-3.8 %), illustrating the accuracy of the modeling. The FSRC coil alone removing ferrite and copper is also simulated. The results are summarized in Table. 8.1, showing high accuracy. The ESR measurement of FSRC with ferrite is not accurate, due to ferrite loss.

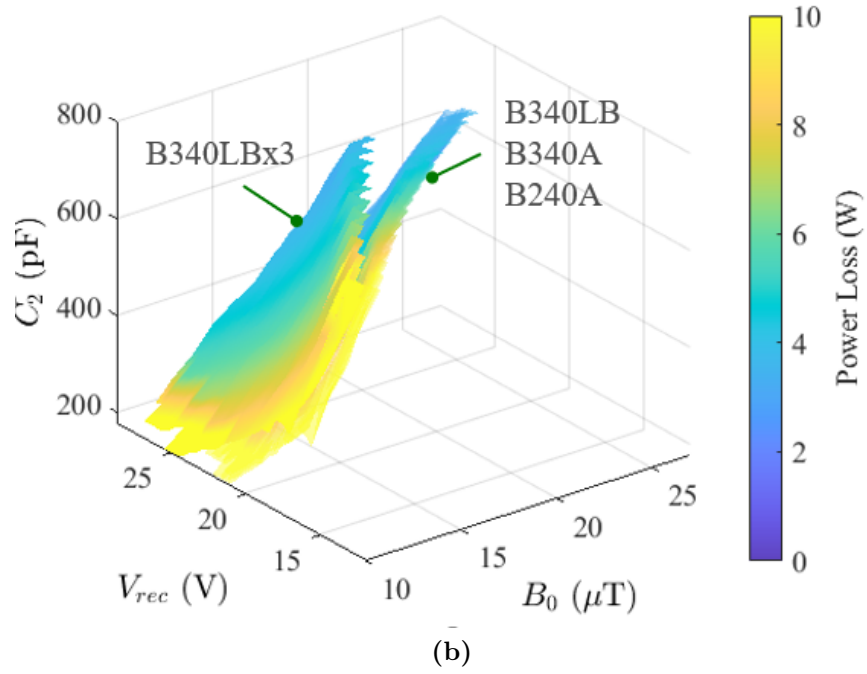
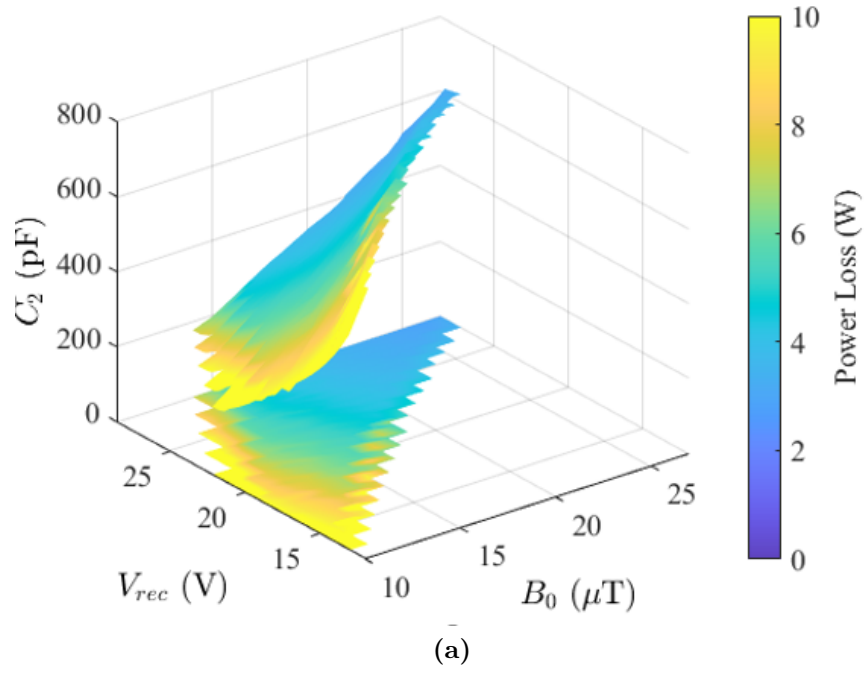


Figure 8.16: Receiver side design result with B340 (a), and multiple devices (b).

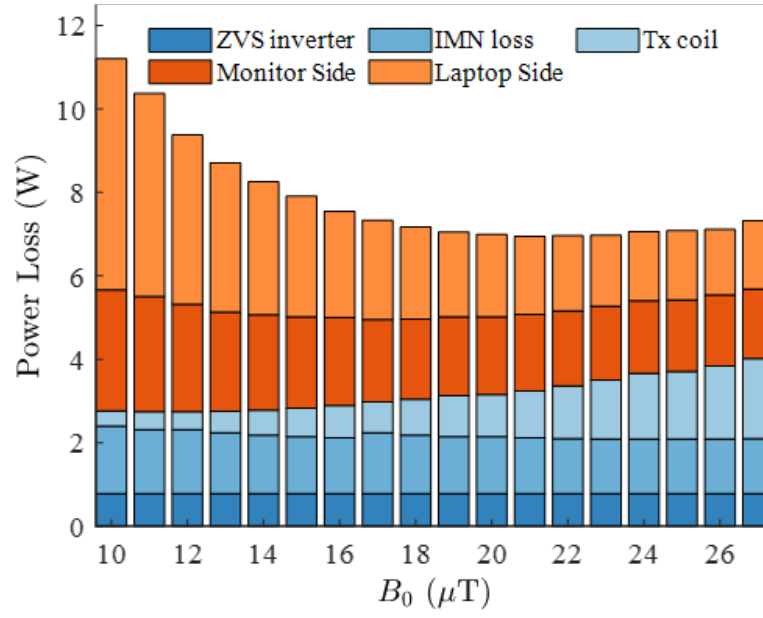


Figure 8.17: Systematic design showing loss breakdown of individual stages.

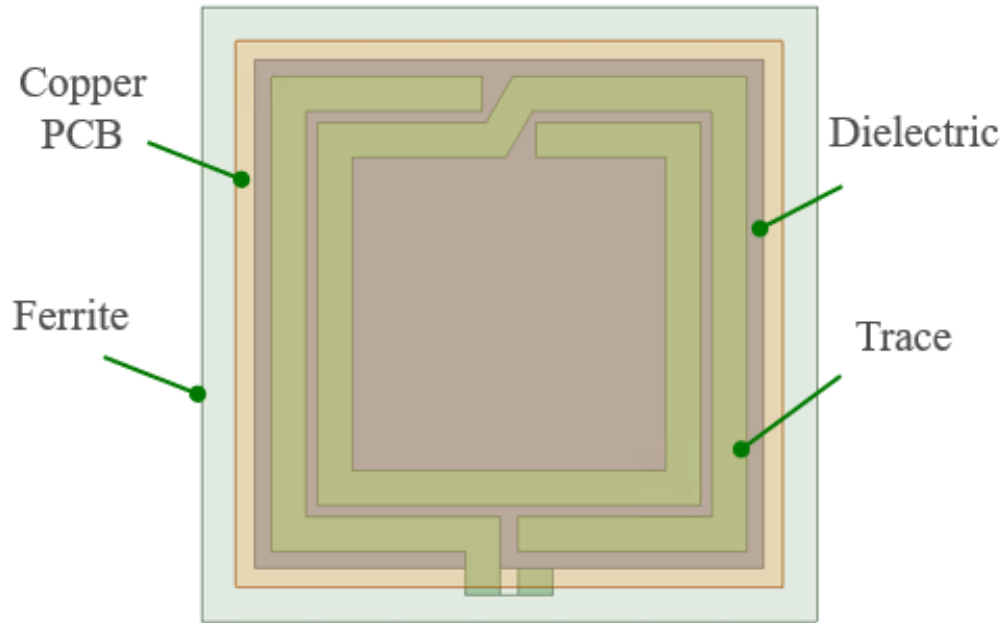


Figure 8.18: HFSS simulation schematic.

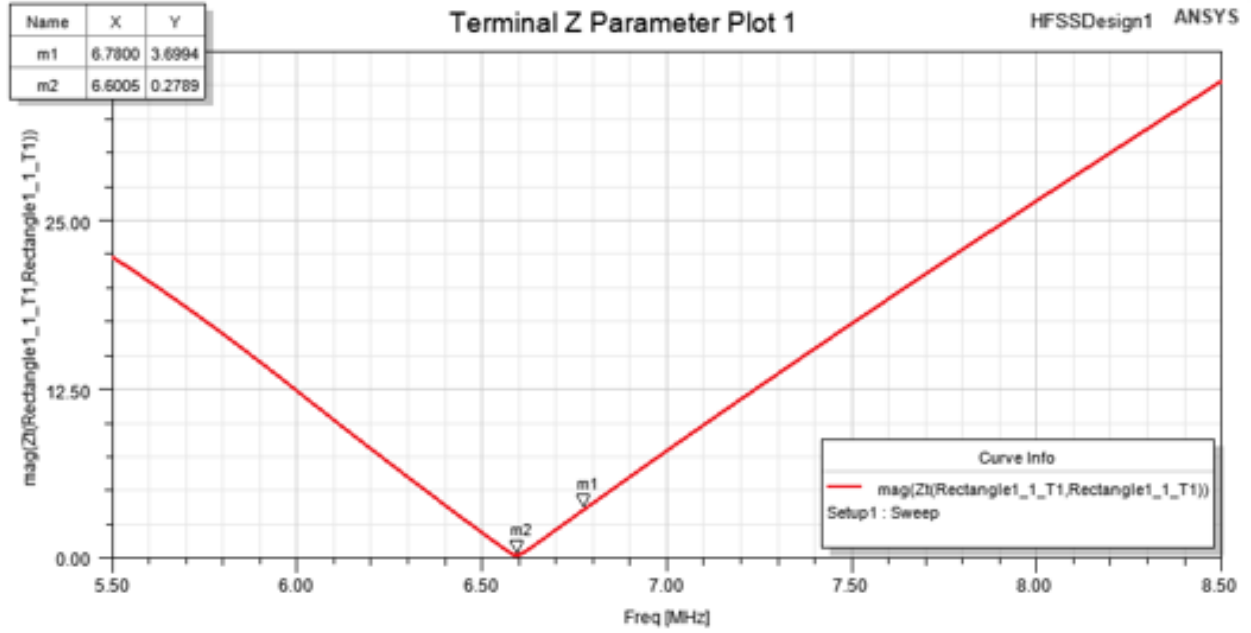


Figure 8.19: HFSS simulation of the impedance.

Table 8.1: Comparison of the measured and calculated circuit parameters.

Parameters	Variable	Design	FEA	Measurement
FSRC resonance with shield (MHz)	$f_{2,s}$	6.63	6.6 (-0.5%)	6.61(-0.3%)
FSRC inductance with shield (μH)	$L_{2,s}$	1.60	1.54 (-3.8%)	1.56 (-2.5%)
FSRC capacitance (pF)	$C_{2,s}$	360	378 (+5%)	372 (+5.7%)
FSRC ESR (Ω)	$R_{2,s}$	0.19	0.23 (+21%)	—
FSRC resonance w/o shield (MHz)	$f_{2,s}$	8.12	8.10 (-0.12%)	8.19 (-0.86%)
FSRC inductance w/o shield (μH)	$L_{2,ns}$	1.07	1.02 (-4.7%)	1.03 (-3.7%)
FSRC capacitance w/o shield (pF)	$C_{2,ns}$	360	380 (+5.6%)	370 (+3.3%)
FSRC ESR w/o shield (Ω)	$R_{2,ns}$	0.19	0.20 (+5.3%)	0.18 (-5.3%)

8.6 Experimental Validation

To verify the idea of the new coil structure and the accuracy of the modeling and proposed design method, an FSRC is fabricated following the steps in [56]. The fabricated coil prototype is shown in Fig. 8.20.

8.6.1 Coil Test in System-level

The FSRC is implemented with the proposed rectifier and is tested with the proposed monitor receiver and transmitter. Fig. 8.21 shows the experimental setup.

The system efficiency is defined as the total DC output power from all receivers divided by the transmitter side DC input power. Measured DC voltages and powers at the full load operating point are summarized in Table 8.2, together with a comparison to the model predictions. The measured power loss is 7.94 W compared to the calculation value of 6.99, proving the accuracy of the system modeling and design. The additional loss might be caused by ferrite. According to the manufacture, [6], the ferrite typically has a loss tangent smaller than 2 %. However, no detailed modeling of how the power loss changes with B was provided to accurately quantify ferrite loss. Another possible reason is that the ferrite increases FSRC coil ESR, as shown in the simulation. The measured system efficiency is 92.7 %. The loss breakdown at this operating point is shown in Fig. 8.22.

8.7 Summary

This chapter solves the receiver design issue in WPT laptop applications. A novel self-resonant coil structure is proposed that achieves a high-Q, low E-field, and thin profile is proposed. The impact of ferrite shielding is studied. The systematic design of a laptop charger with a monitor charger is presented and a complete prototype is presented. Experimental tests verify the coil function, modeling, and the systematic design method application in consumer electronics applications.

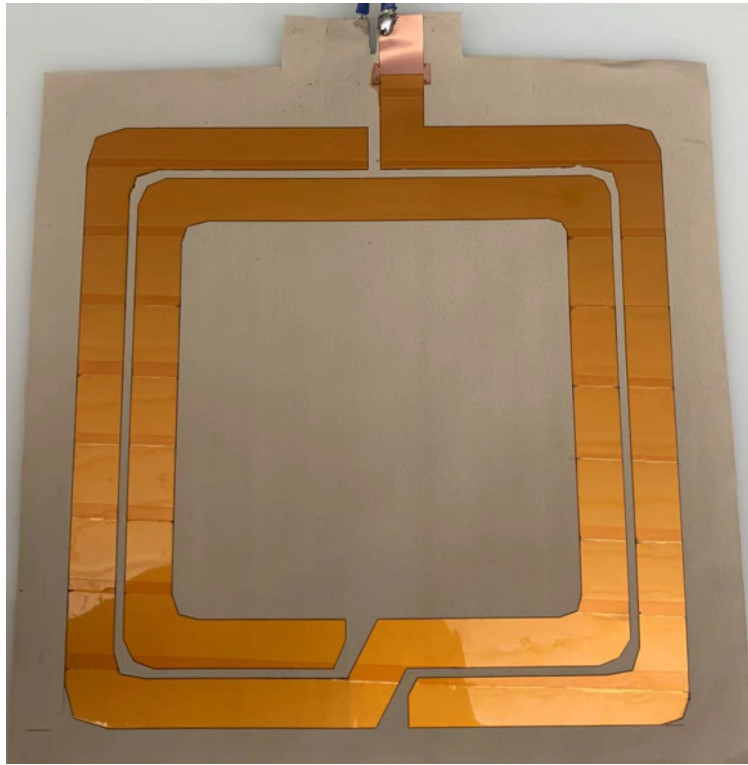


Figure 8.20: Fabricated FSRC coil prototype.

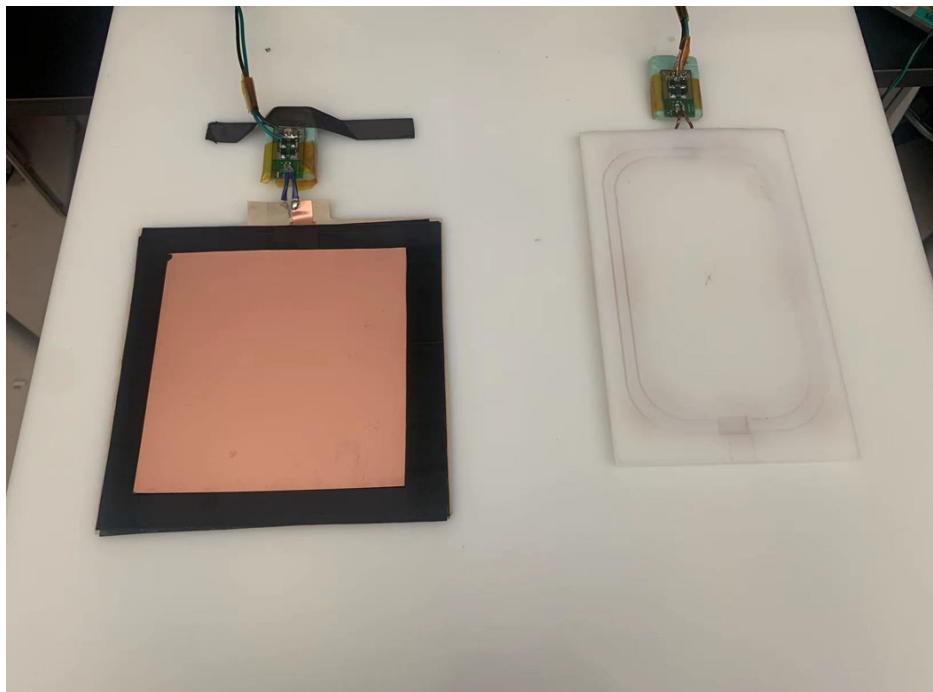


Figure 8.21: Experimental WPT system with FSRC.

Table 8.2: COMPARISON OF MEASURED AND CALCULATED CIRCUIT PARAMETERS

Parameters	Variable	Design	Measure
Transmitter input voltage (V)	V_{dc}	200	201 (+0.5%)
Transmitter input power (W)	P_{in}	107.69	108.54 (+0.37%)
1st receiver output voltage (V)	$V_{o,1}$	31.5	31.1 (-1.3%)
1st receiver output power (V)	$P_{o,1}$	50	47.93 (-4.1%)
New receiver output voltage (V)	$V_{o,2}$	31.5	32.1 (+1.9%)
New receiver output power (W)	$P_{o,2}$	50	52.68 (-5.4%)

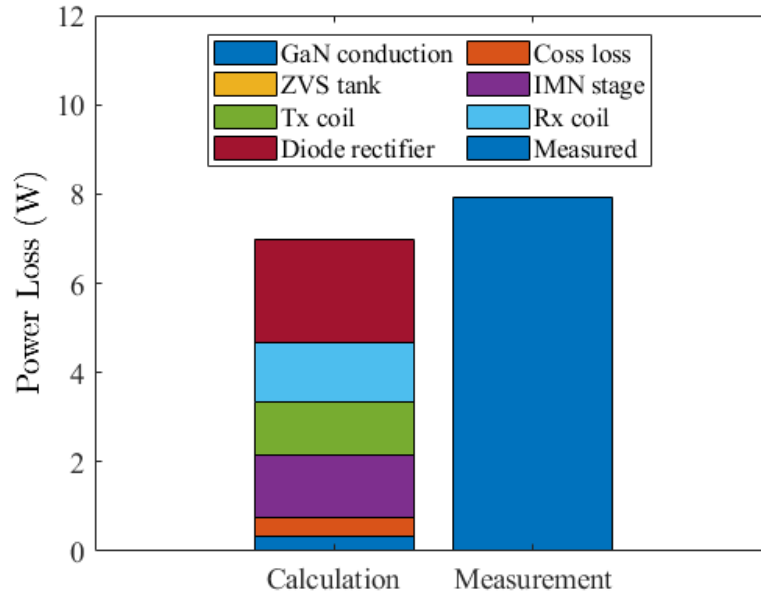


Figure 8.22: Calculated loss components contributing to the total conversion losses of the prototype system at near 100 W output power.

Chapter 9

Conclusion and Future Work

9.1 Conclusions

The promise of higher power and larger spatial freedom has driven the operating frequency of consumer electronics WPT systems into the MHz range. Motivated by the need for a WPT system that exhibits free-loading, free-positioning, and high efficiency, this dissertation details the analysis, design, and prototyping of a 100 W 6.78 MHz WPT system that achieves these features.

From the literature review, the current design method for free-positioning is insufficient for 6.78 MHz WPT covering a large area due to the standing wave issue. The current design method for free-positioning is insufficient or lossy and lacks a design method at the system level to avoid high loss. The current sequential design method for high efficiency is based on assumptions that are not generally true due to the neglect of ZVS requirements and diode rectifier reactance.

In the proposed work, free-positioning is achieved using the geometric and current distribution design of an interleaved coil. Compared to other work, the proposed coil has small field variation, high- Q , and no standing wave issue. Free-loading is achieved using an IMN for constant i_t . Compared to other work, a systematic design of IMN in a WPT system is proposed. High efficiency is achieved by accurate modeling of ZVS inverter with IMN, rectifier, coils, and the entire system, and the proposed system-level design method. The systematic design method parameterizes each stage using the coupling parameters

between adjacent stages. The first step is internal optimization, which leads to an optimal, reduced design space. The second step is system-wide iteration, which addresses loss inter-dependency.

Using the resulting parameters from system modeling and design, a 100 W, 6.78 MHz multi-load prototype is fabricated, achieving 15.9 % field variation, no cross-channel disturbance, and 92.8 % end-to-end efficiency. The proposed features of free-positioning, free-loading, and high efficiency have been verified experimentally.

In many applications, the receiver side is sensitive to the leakage magnetic field and electric field, and are sensitive to shielding thickness. Such applications include WPT charging pads, laptops, and cellphones, all of which are close to people's everyday life. For these applications, a novel self-resonant coil that has intrinsic distributed compensation, high-Q, and low profile are proposed and the co-design with ferrite are studied. Using the systematic design method, the fabricated prototype for a laptop application demonstrates 92.7 % efficiency given strict thickness requirements. The coil working principles, modeling, and the systematic design method are verified.

9.2 Future Work

In the future, several works can enhance the broader adoption of multi-load wireless charging stations.

In the proposed system architecture, the transmitter side always generates a constant magnetic field regardless of the receiver output power. In this operation mode, the power loss in the transmitter coil is constant at light load, which causes a low-efficiency issue. To increase light-load efficiency, the transmitter may be operated in burst-mode, which decreases the PWM signal pulse density to reduce the coil current and power loss, while maintaining soft switching. This control method was discussed in literature [52]. However, this operation mode and control method causes a current ripple in the transmitter coil, which may lead to an induced voltage ripple in the receiver side. This method also increases the transmitter coil power loss. Studying the control method to reduce light-load power loss while maintaining a constant induced voltage is a potential future work.

In consumer electronics applications, meeting the thickness requirement with high system efficiency is challenging. The ferrite shielding design and the receiver coil design are vital and have been discussed in Chapter 8. In this chapter, the ferrite thickness is selected to shield the magnetic field efficiently to avoid interference with neighboring devices, and achieve a relatively high induced voltage for the receiver side efficiency. In some lower power level applications where the receiver may need smaller induced voltage for high efficiency, a thinner ferrite may be selected. In this situation, an additional copper layer is usually added to shield the leakage magnetic after the ferrite to avoid interference with neighboring devices [29]. Studying the co-design of the receiver coil, ferrite shielding, and the additional copper shielding is a potential future work, especially for consumer electronics applications requiring a smaller induced voltage and power, such as a 10 W cellphone charging system.

From the power loss breakdown, the diode rectifier takes a significant portion of the overall loss. The synchronous rectifier is promising to reduce the diode power loss [24]. In the future, the synchronous rectifier may be included in the systematic design method proposed in this thesis, performing internal optimization and system-level parameter selection.

Bibliography

- [1] *BQ500101 datasheet*. Texas Instruments. [10](#)
- [2] *BQ501210 datasheet*. Texas Instruments. [10](#)
- [3] *BQ51050B datasheet*. Texas Instruments. [10](#)
- [4] *C0G (NP0) Dielectric General Specifications*. AVX Corporation. [41](#), [57](#), [58](#)
- [5] *EPC9083 product*. EPC. [10](#)
- [6] *Ferrite Shield DataSheet*. Würth Elektronik. [122](#), [134](#)
- [7] *MWPR1516: 15 Watt Wireless Charging Receiver ICs*. NXP. [5](#)
- [8] *NXQ1TXH5 datasheet*. NXP. [8](#), [10](#)
- [9] *PowerLife Wireless Charging Coils*. Molex. [10](#)
- [10] *SAE Standards*. SAE International. [7](#)
- [11] *T68-6 DataSheet*. Micrometals Inc. [72](#), [73](#)
- [12] (2021). *Qi Certificated Products*. WPC - Wireless Power Consortium. [5](#), [8](#)
- [13] Arteaga, J. M., Aldhaher, S., Kkelis, G., Yates, D. C., and Mitcheson, P. D. (2017). Multi-mhz ipt systems for variable coupling. *IEEE Transactions on Power Electronics*, 33(9):7744–7758. [40](#)
- [14] Beh, T. C., Kato, M., Imura, T., Oh, S., and Hori, Y. (2013a). Automated impedance matching system for robust wireless power transfer via magnetic resonance coupling. *IEEE Transactions on Industrial Electronics*, 60(9):3689–3698. [17](#)

- [15] Beh, T. C., Kato, M., Imura, T., Oh, S., and Hori, Y. (2013b). Automated impedance matching system for robust wireless power transfer via magnetic resonance coupling. *IEEE Transactions on Industrial Electronics*, 60(9):3689–3698. [36](#)
- [16] Berger, A., Agostinelli, M., Vesti, S., Oliver, J. A., Cobos, J. A., and Huemer, M. (2015). A wireless charging system applying phase-shift and amplitude control to maximize efficiency and extractable power. *IEEE Transactions on Power Electronics*, 30(11):6338–6348. [36](#)
- [17] Bosshard, R. and Kolar, J. W. (2016). Multi-objective optimization of 50 kW/85 kHz IPT system for public transport. *IEEE Journal of Emerging and Selected Topics in Power Electronics*, 4(4):1370–1382. [25](#), [27](#), [29](#), [86](#)
- [18] Bosshard, R., Kolar, J. W., and Mühlethaler, J. (2015). Modeling and η - α -pareto optimization of inductive power transfer coils for electric vehicles. *IEEE Journal of Emerging and Selected Topics in Power Electronics*, 3(1):50–64. [25](#), [61](#), [64](#)
- [19] Carobolante, F., Menegoli, P., Marino, F. A., and Jeong, N. S. (2017). A novel charger architecture for resonant wireless power transfer. *IEEE Journal of Emerging and Selected Topics in Power Electronics*, pages 1–1. [x](#), [17](#), [19](#), [20](#)
- [20] Chan, P. C. F., Lee, C. K., and Hui, S. Y. R. (2007). Stray capacitance calculation of coreless planar transformers including fringing effects. *Electronics Letters*, 43(23):1308. [xiv](#), [29](#), [31](#), [123](#)
- [21] Chen, K. and Zhao, Z. (2013). Analysis of the double-layer printed spiral coil for wireless power transfer. *IEEE Journal of Emerging and Selected Topics in Power Electronics*, 1(2):114–121. [x](#), [31](#), [32](#)
- [22] Cheng, Y. and Shu, Y. (2014). A new analytical calculation of the mutual inductance of the coaxial spiral rectangular coils. *IEEE Transactions on Magnetics*, 50(4):1–6. [50](#)
- [23] Cochran, S. and Costinett, D. (2017a). Modeling a 6.78 MHz synchronous WPT rectifier with reduced THD. In *14th Workshop on Control and Modeling for Power Electronics (COMPEL)*, pages 1–8. IEEE. [29](#), [36](#), [86](#)

- [24] Cochran, S. and Costinett, D. (2017b). Modeling a 6.78 MHz synchronous WPT rectifier with reduced THD. In *14th Workshop on Control and Modeling for Power Electronics (COMPEL)*, pages 1–8. IEEE. [139](#)
- [25] Costinett, D., Maksimovic, D., and Zane, R. (2014). Circuit-oriented treatment of nonlinear capacitances in switched-mode power supplies. *IEEE Transactions on Power Electronics*, 30(2):985–995. [67](#)
- [26] de Miranda, C. M. and Pichorim, S. F. (2017). A self-resonant two-coil wireless power transfer system using open bifilar coils. *IEEE Transactions on Circuits and Systems II: Express Briefs*, 64(6):615–619. [31](#)
- [27] Deng, C., Zhu, G., and Lorenz, R. D. (2017). MHz frequencies, kW, 30 cm gap wireless power transfer with low air gap flux density and high efficiency using surface spiral winding coils. In *Applied Power Electronics Conference and Exposition (APEC), 2017 IEEE*, pages 1606–1613. IEEE. [35](#)
- [28] Elkhoully, E. and Yang, S. (2016). Transmitter coil design for resonant wireless power transfer. pages 1–5. [x](#), [17](#), [19](#), [20](#), [109](#), [111](#)
- [29] Feng, J., Fu, M., Li, Q., and Lee, F. C. (2017a). Resonant converter with coupling and load independent resonance for omnidirectional wireless power transfer application. pages 2596–2601. [139](#)
- [30] Feng, J., Li, Q., and Lee, F. C. (2017b). Omnidirectional wireless power transfer for portable devices. In *2017 IEEE Applied Power Electronics Conference and Exposition (APEC)*, pages 1675–1681. IEEE. [x](#), [17](#), [20](#), [22](#)
- [31] Feng, J., Li, Q., Lee, F. C., and Fu, M. (2019). Transmitter coils design for free-positioning omnidirectional wireless power transfer system. *IEEE Transactions on Industrial Informatics*, 15(8):4656–4664. [17](#), [77](#)

- [32] Florian, C., Mastri, F., Paganelli, R. P., Masotti, D., and Costanzo, A. (2014). Theoretical and numerical design of a wireless power transmission link with gan-based transmitter and adaptive receiver. *IEEE Transactions on Microwave Theory and Techniques*, 62(4):931–946. [29](#), [35](#), [40](#)
- [33] Fotopoulou, K. and Flynn, B. W. (2011). Wireless power transfer in loosely coupled links: Coil misalignment model. *IEEE Transactions on Magnetics*, 47(2):416–430. [29](#), [35](#)
- [34] Fu, M., Yin, H., Liu, M., and Ma, C. (2016a). Loading and power control for a high-efficiency class e pa-driven megahertz wpt system. *IEEE Transactions on Industrial Electronics*, 63(11):6867–6876. [34](#)
- [35] Fu, M., Yin, H., Liu, M., and Ma, C. (2016b). Loading and power control for a high-efficiency class e PA-driven megahertz WPT system. *IEEE Transactions on Industrial Electronics*, 63(11):6867–6876. [40](#)
- [36] Fu, M., Yin, H., Zhu, X., and Ma, C. (2015a). Analysis and tracking of optimal load in wireless power transfer systems. *IEEE Transactions on Power Electronics*, 30(7):3952–3963. [40](#)
- [37] Fu, M., Zefan, T., Liu, M., Ma, C., and Zhu, X. (2015b). Full-bridge rectifier input reactance compensation in megahertz wireless power transfer systems. In *Emerging Technologies: Wireless Power (WoW), 2015 IEEE PELS Workshop on*, pages 1–5. IEEE. [xi](#), [36](#), [38](#), [86](#)
- [38] Garnica, J., Chinga, R. A., and Lin, J. (2013). Wireless power transmission: From far field to near field. *Proceedings of the IEEE*, 101(6):1321–1331. [2](#), [4](#)
- [39] Grandi, G., Kazimierczuk, M. K., Massarini, A., and Reggiani, U. (2002). Stray capacitances of single-layer air-core inductors for high-frequency applications. In *Industry Applications Conference, 1996. Thirty-First IAS Annual Meeting, IAS '96., Conference Record of the 1996*, pages 1384–1388. IEEE. [29](#), [31](#)
- [40] Green, A. W. and Boys, J. (1994). 10 khz inductively coupled power transfer-concept and control. [2](#)

- [41] Hui, S. (2013). Wireless power transfer: A brief review & update. In *2013 5th International Conference on Power Electronics Systems and Applications (PESA)*, pages 1–4. IEEE. [8](#)
- [42] Hui, S. Y. R., Zhong, W., and Lee, C. K. (2014). A critical review of recent progress in mid-range wireless power transfer. *IEEE Transactions on Power Electronics*, 29(9):4500–4511. [2](#), [4](#), [5](#), [15](#), [17](#), [23](#), [25](#)
- [43] Hurley, W. and Kassakian, J. (1979). Induction heating of circular ferromagnetic plates. *IEEE Transactions on Magnetics*, 15(4):1174–1181. [2](#)
- [44] ICNIRP (2010). Guidelines for limiting exposure to time-varying electric and magnetic fields (1 Hz to 100 kHz). *Health physics*, 99(6):818–836. [4](#), [44](#)
- [45] Jang, Y. and Jovanovic, M. M. (2003). A contactless electrical energy transmission system for portable-telephone battery chargers. *IEEE Transactions on Industrial Electronics*, 50(3):520–527. [2](#)
- [46] Jiang, L. and Costinett, D. (2018). A high-efficiency GaN-based single-stage 6.78 MHz transmitter for wireless power transfer applications. *IEEE Transactions on Power Electronics*, 34(8):7677–7692. [23](#), [25](#), [80](#)
- [47] Jiang, L., Costinett, D., Fathy, A., and Yang, S. (2017). A single stage AC/RF converter for wireless power transfer applications. In *Applied Power Electronics Conference and Exposition (APEC), 2017 IEEE*, pages 1682–1688. IEEE. [29](#), [34](#), [36](#)
- [48] Ke, Q., Luo, W., Yan, G., and Yang, K. (2016). Analytical model and optimized design of power transmitting coil for inductively coupled endoscope robot. *IEEE Trans Biomed Eng*, 63(4):694–706. [29](#), [31](#)
- [49] Kurs, A., Karalis, A., Moffatt, R., Joannopoulos, J. D., Fisher, P., and Soljacic, M. (2007). Wireless power transfer via strongly coupled magnetic resonances. *Science*, 317(5834):83–6. [2](#), [20](#)

- [50] Kyaw, P. A., Stein, A. L. F., and Sullivan, C. R. (2017). High-Q resonator with integrated capacitance for resonant power conversion. In *Applied Power Electronics Conference and Exposition (APEC), 2017 IEEE*, pages 2519–2526. IEEE. [x](#), [20](#), [21](#), [29](#), [31](#), [34](#), [35](#), [111](#)
- [51] Lee, S.-H. and Lorenz, R. D. (2013). Surface spiral coil design methodologies for high efficiency, high power, low flux density, large air-gap wireless power transfer systems. In *Applied Power Electronics Conference and Exposition (APEC), 2013 IEEE*, pages 1783–1790. IEEE. [x](#), [27](#), [29](#), [30](#)
- [52] Li, H., Fang, J., Chen, S., Wang, K., and Tang, Y. (2017). Pulse density modulation for maximum efficiency point tracking of wireless power transfer systems. *IEEE Transactions on Power Electronics*, pages 1–1. [23](#), [29](#), [34](#), [36](#), [86](#), [138](#)
- [53] Li, H., Li, J., Wang, K., Chen, W., and Yang, X. (2015a). A maximum efficiency point tracking control scheme for wireless power transfer systems using magnetic resonant coupling. *IEEE Transactions on Power Electronics*, 30(7):3998–4008. [17](#), [23](#), [25](#), [27](#), [34](#), [36](#), [77](#)
- [54] Li, H., Wang, K., Fang, J., and Tang, Y. (2018a). Pulse density modulated ZVS full-bridge converters for wireless power transfer systems. *IEEE Transactions on Power Electronics*, pages 1–1. [xi](#), [34](#), [35](#), [36](#), [37](#), [40](#), [70](#)
- [55] Li, H., Wang, K., Huang, L., Li, J., and Yang, X. (2015b). Coil structure optimization method for improving coupling coefficient of wireless power transfer. pages 2518–2521. [17](#), [20](#)
- [56] Li, J. and Costinett, D. (2018). Analysis and design of a series self-resonant coil for wireless power transfer. In *2018 IEEE Applied Power Electronics Conference and Exposition (APEC)*, pages 1052–1059. IEEE. [35](#), [114](#), [119](#), [120](#), [125](#), [134](#)
- [57] Li, L., Liu, H., Zhang, H., and Xue, W. (2018b). Efficient wireless power transfer system integrating with metasurface for biological applications. *IEEE Transactions on Industrial Electronics*, 65(4):3230–3239. [2](#)

- [58] Liu, M., Fu, M., and Ma, C. (2016). Parameter design for a 6.78-MHz wireless power transfer system based on analytical derivation of class E current-driven rectifier. *IEEE Transactions on Power Electronics*, 31(6):4280–4291. [34](#), [77](#), [86](#)
- [59] Liu, M., Fu, M., and Ma, C. (2017). Low-harmonic-contents and high-efficiency class E full-wave current-driven rectifier for megahertz wireless power transfer systems. *IEEE Transactions on Power Electronics*, 32(2):1198–1209. [35](#), [40](#)
- [60] Lope, I., Carretero, C., Acero, J., Alonso, R., and Burdio, J. M. (2014). Frequency-dependent resistance of planar coils in printed circuit board with litz structure. *IEEE Transactions on Magnetics*, 50(12):1–9. [27](#), [29](#)
- [61] Lu, X., Wang, P., Niyato, D., Kim, D. I., and Han, Z. (2015). Wireless charging technologies: Fundamentals, standards, and network applications. *IEEE Communications Surveys Tutorials*, 18(2):1413–1452. [2](#), [4](#)
- [62] Lu, X., Wang, P., Niyato, D., Kim, D. I., and Han, Z. (2016). Wireless charging technologies: Fundamentals, standards, and network applications. *IEEE Communications Surveys Tutorials*, 18(2):1413–1452. [5](#)
- [63] López-Alcolea, F. J., d. Real, J. V., Roncero-Sánchez, P., and Torres, A. P. (2020). Modeling of a magnetic coupler based on single- and double-layered rectangular planar coils with in-plane misalignment for wireless power transfer. *IEEE Transactions on Power Electronics*, 35(5):5102–5121. [52](#), [54](#)
- [64] Mohan, S. S., del Mar Hershenson, M., Boyd, S. P., and Lee, T. H. (1999). Simple accurate expressions for planar spiral inductances. *IEEE Journal of Solid-State Circuits*, 34(10):1419–1424. [29](#), [119](#)
- [65] Pantic, Z. and Lukic, S. (2013). Computationally-efficient, generalized expressions for the proximity-effect in multi-layer, multi-turn tubular coils for wireless power transfer systems. *IEEE Transactions on Magnetics*, 49(11):5404–5416. [5](#), [29](#), [31](#)

- [66] Pham, P., Cochran, S., Costinett, D. J., and Tolbert, L. M. (2020). Active rectifier design and synchronization control for 6.78 mhz wireless power transfer. In *2020 IEEE Energy Conversion Congress and Exposition (ECCE)*, pages 5501–5508. IEEE. [25](#)
- [67] Regensburger, B. (2020). *Capacitive Wireless Power Transfer Systems for Electric Vehicle Charging*. PhD thesis, Cornell University. [4](#)
- [68] Sample, A. P., Meyer, D. A., and Smith, J. R. (2011). Analysis, experimental results, and range adaptation of magnetically coupled resonators for wireless power transfer. *IEEE Transactions on Industrial Electronics*, 58(2):544–554. [29](#), [35](#), [40](#)
- [69] Schuler, J. (1961). High-level electromagnetic energy transfer through a closed chest wall. *IRE International Conv. Rec. pt 9*, 9:119–126. [1](#)
- [70] Son, H. C., Kim, J., Kim, D. H., Kim, K. H., and Park, Y. J. (2011). Self-resonant coil with coaxial-like capacitor for wireless power transfer. In *Microwave Conference Proceedings (APMC), 2011 Asia-Pacific*, pages 90–93. IEEE. [xi](#), [31](#), [33](#)
- [71] Sullivan, C. R. (2001). Cost-constrained selection of strand diameter and number in a litz-wire transformer winding. *IEEE Transactions on Power Electronics*, 16(2):281–288. [29](#)
- [72] Sullivan, C. R. and Beghou, L. (2013). Design methodology for a high-Q self-resonant coil for medical and wireless-power applications. In *14th Workshop on Control and Modeling for Power Electronics (COMPEL)*, pages 1–8. IEEE. [xi](#), [29](#), [31](#), [32](#), [120](#), [121](#)
- [73] Sun, J., Gui, H., Li, J., Huang, X., Strain, N., Costinett, D., and Tolbert, L. G. M. (2021). Mitigation of current distortion for gan-based crm totem-pole pfc rectifier with zvs control. *IEEE Open Journal of Power Electronics*, pages 1–1. [xiii](#), [44](#), [76](#), [103](#), [104](#)
- [74] Surakitbovorn, K. and Rivas-Davilla, J. (2018). Design of a gan-based wireless power transfer system at 13.56 mhz to replace conventional wired connection in a vehicle. In *2018 International Power Electronics Conference (IPEC-Niigata 2018-ECCE Asia)*, pages 3848–3854. IEEE. [40](#)

- [75] Tesla, N. (1914). Apparatus for transmitting electrical energy. US Patent 1,119,732. [1](#)
- [76] Tseng, R., von Novak, B., Shevde, S., and Grajski, K. A. (2013). Introduction to the alliance for wireless power loosely-coupled wireless power transfer system specification version 1.0. In *Wireless Power Transfer (WPT), 2013 IEEE*, pages 79–83. IEEE. [7](#)
- [77] Tuan, T. M. and Choi, W. (2019). Design and implementation of a constant current and constant voltage wireless charger operating at 6.78 mhz. In *2019 10th International Conference on Power Electronics and ECCE Asia (ICPE 2019-ECCE Asia)*, pages 1–6. IEEE. [40](#)
- [78] Waffenschmidt, E. (2015). Homogeneous magnetic coupling for free positioning in an inductive wireless power system. *IEEE Journal of Emerging and Selected Topics in Power Electronics*, 3(1):226–233. [x](#), [17](#), [20](#), [21](#)
- [79] Wang, S., deRooij, M. A., Odendaal, W. G., vanWyk, J. D., and Boroyevich, D. (2005). Reduction of high-frequency conduction losses using a planar litz structure. *IEEE Transactions on Power Electronics*, 20(2):261–267. [x](#), [xi](#), [27](#), [29](#), [30](#), [33](#)
- [80] Yang, Y., Liu, F. X., and Chen, X. L. (2017). A maximum power point tracking control scheme for magnetically coupled resonant wireless power transfer system by cascading SEPIC converter at the receiving side. *2017 Thirty Second Annual IEEE Applied Power Electronics Conference and Exposition (Apec)*, pages 3702–3707. [5](#), [17](#)
- [81] Zhao, C., Wang, Z., Du, J., Wu, J., Zong, S., and He, X. (2014). Active resonance wireless power transfer system using phase shift control strategy. In *Applied Power Electronics Conference and Exposition (APEC), 2014 IEEE*, pages 1336–1341. IEEE. [36](#)
- [82] Zulauf, G., Park, S., Liang, W., Surakitbovorn, K. N., and Rivas-Davila, J. (2018). C_{oss} losses in 600 V GaN power semiconductors in soft-switched, high-and very-high-frequency power converters. *IEEE Transactions on Power Electronics*, 33(12):10748–10763. [xii](#), [69](#), [70](#), [71](#), [76](#)

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