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**INTEGRATED CIRCUITS DESIGN AND
CONTROL SOLUTIONS FOR WIRELESS
POWER TRANSFER APPLICATIONS**

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ABSTRACT

In the past years, extensive research has been carried out on wireless power transfer (WPT), since it offers customers a convenient charging technique compared to wires which can be inconvenient and allows charging of portable devices in a wide power range. WPT technology has been observed to have some remarkable technological advantages due to its spatial freedom. Owing to modern power electronics technology and high-power semiconductors, WPT technology is highly integrated, and many smart WPT devices have been developed for different applications such as biomedical devices, electric vehicles (EV), and portable consumer electronic devices. A burst in the WPT market for consumer electronics is expected in the coming five years, due to the ubiquitous usage of 3G/4G wireless networks and continually evolving technologies for media and entertainment. WPT systems for consumer devices can be classified into two main types: inductive power transfer (IPT) and magnetic resonant coupling (MRC). The former works at 100-200 kHz, whereas the latter works at 6.78 or 13.56 MHz. The bottleneck of the IPT technology is to design a low cost but high performance WPT system since the IPT mainly targets on the application with less requirement of distance freedom like home appliance or some industrial applications with short charging distance. The main targets of MRC technology is to provide consumers a flexible WPT solution with higher distance freedom. The main concern of this technology is that 1) it is challenging to design high efficiency circuits for MHz application, especially for rectifier; 2) the control of the MRC is challenging since the output power would be effected remarkably by distance, which is known as splitting frequency phenomenon.

This research develops circuits and control algorithms for both types of WPT systems. Firstly, a single-stage AC-DC voltage regulator is developed. A non-charging configuration is introduced to regulate the output voltage at a desired level with a pulse width modulation (PWM) control algorithm. In this way, there is no need for the DC-DC stage after the rectifier and additional wireless communication circuit blocks. In addition, a full-active cross-coupled structure is utilized to reduce the power loss of the regulator. Experimental results of the proposed regulator based on a $0.35\ \mu\text{m}$ technology and PCB circuit demonstrate a maximum output power of 15 W and peak efficiency of 92 %.

Another contribution of this dissertation is a design of a full active rectifier working at 6.78 MHz. An adaptive time delay (ATD) circuit is used to maximize the conduction interval of the gallium nitride (GaN) switch in order to minimize the power loss due to the forward voltage drop of the diode. The proposed control algorithm also eliminates the reverse leakage current of the rectifier. Except the power devices, all the other circuits have been taped out with a $0.18\ \mu\text{m}$ CMOS process. The experimental results of the proposed rectifier shows that it can output a maximum output current of 3 A at 5 V with a 6.78-MHz AC input voltage with a peak power efficiency over 90%.

The last contribution is a novel control method for a 6.78-MHz MRC system. In this part, detailed analysis of MRC, especially on the relationship between operating frequency, transfer efficiency, output power, and coupling coefficient are presented. Contrary to the traditional research, this research proves that the MRC system is able to work with sufficient output power and efficiency within a wide coupling coefficient range, relaxing the system from the bound of the key coupling coefficient limitation. A hybrid control method with frequency/phase-shift tuning with zero voltage switching (ZVS) Class-D amplifier is proposed to provide the load with a constant output voltage. The testing results shows that it is able to transfer 10 W over a 5 cm with an overall efficiency (end to end) of 71.8 %, and over a 1 cm with an overall efficiency of 73.1%.

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LIST OF ABBREVIATIONS

WPT	wireless power transfer
IPT	inductive power transfer
MRC	magnetic resonant coupling
AC	alternating current
DC	direct current
EV	electric vehicles
CMT	coupled mode theory
ZPA	zero phase angle
M	mutual inductance
L_1	self-inductance of the transmitting coil
L_2	self-inductance of the receiving coil
ZVS	zero voltage switching
PCE	power conversion efficiency
DC	direct current
PWM	pulse width modulation
PFM	pulse frequency modulation
EMI	electro magnetic interference
PSM	pulse skipping modulation
CCM	continuous conduction mode
DCM	discontinuous conduction mode
PVT	process, voltage and temperature
THD	total harmonic distortion

GaN	gallium nitride
CMOS	complementary metal-oxide semiconductor
PCB	printed circuit board
RFID	radio-frequency identification
LCL	inductor-capacitor-inductor
LCCL	inductor-capacitor-capacitor-inductor
MOSFET	metal-oxide-semiconductor field-effect transistor
FPGA	field-programmable gate array
PID	proportional-integral-derivative
ATD	adaptive time delay
SS	series-series
SP	series-Parallel
PS	parallel-series
PP	parallel-parallel
VA	voltage-ampere
EM	electromagnetic

Chapter 1

Introduction

1.1 Background

The idea of wireless power transfer can be dated back to 1901, when Nikola Tesla first attempted to carry out his famous Wardencllyffe Tower experiment (Figure 1.1) near Long Island, for transmitting power wirelessly [1]. Unfortunately, due to the limited radio-wave technology available in the early 1900s, his experiment to prove the feasibility of WPT was not successful. With the development of electromagnetic (EM) wave technology, WPT via EM has drawn researchers attention since the late 1970s [2] [3] [4] [5] [6]. In an EM WPT system, a transmitter emits EM waves in a specific industrial, scientific, and medical (ISM) band (850-950 or 902-928 MHz) [6] and a receiver with the same frequency band catches and stores the EM power (Figure 1.2(a)). However, the trade-off between efficiency and transmission directionality hinders the development of this far-field approach [3] , [7].

The near-field IPT [8] [9] [10] [11] [12] [13] is becoming popular as a result of the fast development of portable consumer devices. The operation principle of an IPT system is EM induction, which is similar to that of a transformer. It is an efficient WPT system for short-distance applications because it is non-radiative so the energy is not wasted in the medium. IPT technology is widely used for mid-power home appliances such as mobile phones (Figure 1.2(b)) or electric

Figure 1.1: Nikola Tesla's experiment known as Wardenclyffe Tower in 1901

Figure 1.2: WPT technologies for three applications: (a) the EM wave WPT for unmanned plane receives power beamed from earth 21, (b) IPT WPT for a mobile phone 22 and (c) MRC WPT system developed by Intel using flat coils 23

toothbrushes, and efficiency of over 70% is possible for these applications. For high-power applications of up to several kilowatts, the efficiency of IPT technology could be as high as 90 %. The operating frequency range of IPT is generally from 20 KHz to 1 MHz.

Despite the advantages of IPT technology, one problem is that its efficiency decreases as the distance or the position mismatch between the two coils becomes larger. A new technology, known as MRC WPT, has been developed to solve this problem. In 2007, an research group from Massachusetts Institute of Technology first conducted experiments using this technology to transfer 60 W over 100 cm with an efficiency of over 90 % [14] [15]. In this experiment, two resonant coils with a Q-factor nearly as high as 1000 form a powerful resonator

and thus enhance the efficiency of WPT over longer distances. Compared to EM WPT, resonant WPT is less sensitive to the environment [16]. Numerous research works have been carried out to accelerate the development of resonant WPT. Figure 1.2(c) shows a new resonant coupled WPT developed by Intel in 2008. In this system, a pair of planar coils are used, as they can be implanted into consumer devices more easily than helix coils [15]. Several single-transmitter to multi-receiver systems based on resonant WPT technology have been developed in recent years [17]. The operating frequency of this technology is from 1 MHz to nearly 200 MHz [18]. A summary of these three kinds of WPT technologies is given in Table 1.1

Table 1.1: Summary of three main wireless power transfer technology

Types	Advantages	Disadvantages	Applications
EM wave	<p>The size of receiver is tiny</p> <p>The transfer efficiency is high over a wide power and load variations.</p>	<p>Sensitive to distance</p> <p>Only for low-power applications</p>	<p>Wireless sensor network</p> <p>Unmanned plane</p>
IPT	<p>High safety</p> <p>High efficiency in short transfer distance</p> <p>High power transfer capability in short transfer distance</p>	<p>Short transfer distance</p> <p>Sensitive to alignment</p>	<p>Portable electronic devices such as cell phones and laptops</p> <p>Electric Vehicles</p>
MRC WPT	<p>High efficiency for mid-range power transfer</p> <p>Insensitive to the environment</p>	<p>Sensitive to alignment</p> <p>Large size of coil</p>	<p>Portable electronic devices such as cell phones and laptops</p> <p>Implantable devices</p>

1.2 Motivation

The progress of technology through the history of human kind has continually changed people's way of life. At the end of the 20th century, the development of personal computers brought us into the information age. At the beginning of the 21st century, with the rapid growth of wireless communication, we are entering a new era of wireless technology. However, people still use wires to charge their wireless devices, which ironically cause wireless devices to become non-wireless. Wireless Power Transfer (WPT) is a suitable solution for this problem. With continuous efforts in improving the process of modern semiconductor industry, it seems more attractive than ever to develop a high performance WPT system with smart control and affordable cost.

Moreover, due to the widespread usage of technologies for entertainment such as high-definition video and virtual reality games, the escalating power consumption of electronic devices has become a growing concern in both the industrial and academic fields because the duration of device operation is quite limited. One solution is to increase the storage capacity of the battery, but this is limited to chemistry and materials science due to their current state of development. In that case, a flexible power transfers solution, like WPT, is considered to be appealing for consumers for the following reasons: 1) this technology expand the spatial freedom of charger and devices for users; 2) it is an effective power transfer method for unreachable loads, like several applications in the bio-medical application and automation industry; 3) without exposed conductors, it can be considered a safer power transfer process. Despite these advantages, it yet remains as a challenge task to design a high-performance WPT. It have a larger power loss compared to traditional plug-in chargers due to the power consumed in the extra circuits (like power amplifier, rectifiers, voltage regulator) and transmission coils. Besides, a WPT system consists of two separate parts without any physical contact so the power flow control becomes difficult. These advantages

and challenges motivate the work in this thesis.

1.3 Objectives and Research Scope

The objectives of this research project are listed as follows:

- To improve the design and performance of circuits in WPT system targeting IPT WPT and MRC WPT by utilizing different circuit topologies;
- To develop smart, low cost, high performance output voltage control methods for IPT wireless power transfer system @ kHz;
- To develop a strategy for MRC technology @ MHz targeting at high-end WPT application to maintain the performance of the entire system within a large dynamic range of different parameters of the system, such as transferring distance, load conditions and so on;

In this research, WPT system and circuits are designed for consumer electronic devices. The power of the system and circuits is limited to 15 W and the output voltage is limited to below 10 V. However, the applications of the system structure, the circuit structure, and the control algorithm are not limited to this power range if the benefits proven to be attractive for high-power applications and if the process can handle the required voltage and current. EM WPT systems with operating frequencies above 100 MHz are not included in this thesis due to their low efficiency, low output power and safety issues.

1.4 Thesis Contributions

The main contributions of this thesis are as follows:

- Design of a single-stage 125-kHz IPT wireless power transfer receiver with a Pulse Width Modulation (PWM)-controlled active power stage. The output voltage is controlled at a constant value, such as 5 V in a mobile

phone charger. Unlike the traditional two-stage alternating current (AC)-direct current (DC) regulator or voltage regulation at the transmitter by using wireless communication, a non-charging configuration is introduced to regulate the output voltage at a constant voltage with a PWM control algorithm. With this method, there is no need for the DC-DC stage after the rectifier and wireless communication circuit blocks. Besides, a full active cross-coupled rectifier with a high-speed comparator effectively reduces the power loss of the regulator. This solution is suitable for those low cost IPT application.

- Design of a high-efficiency full active rectifier for MRC wireless power transfer system at 6.78-MHz with ATD control. For this design, an ATD control circuit is developed to expand the conduction interval of the GaN switch in order to reduce the power loss due to the forward voltage of the diode. The proposed control algorithm also eliminates the reverse leakage current of the rectifier, thus improving the power conversion efficiency of the rectifier. Meanwhile, the proposed rectifier is designed with the capability to self-start in the initial stage when the voltage level is low. The controller is implemented in a high-voltage complementary metal-oxide semiconductor (CMOS) process. GaN transistors are assembled with the controller on a printed circuit board (PCB). The proposed rectifier can output a maximum output current of 3 A at 5 V with a 6.78-MHz AC input voltage. The proposed rectifier can achieve a peak power efficiency of 90 %.
- Design of a hybrid control method for 6.78-MHz MRC wireless power transfer system with varying transmission distance with a ZVS class-D power amplifier. The efficiency (coil to coil) of the WPT system decreases as the distance increases. So a 6.78-MHz MRC technology is used to enhance the efficiency. However, traditional MRC WPT has to work at resonant frequency over a key coupling coefficient to maintain the efficiency

and output power. In this thesis, detailed analysis of MRC WPT, especially on the relationship between operating frequency, transfer efficiency, output power, and coupling coefficient are presented. Contrary to the traditional research, this thesis proves that the system is able to work with sufficient output power and efficiency within a wide coupling coefficient range. A hybrid control method with frequency/phase-shift tuning with ZVS Class-D amplifier is proposed to provide the load with a constant output voltage with high efficiency in a wide range of charging distance.

1.5 Organization of the Thesis

This thesis consists of six chapters. Chapter 1 introduces the background of WPT and lists the research objectives and scope. Chapter 2 reviews the published literature with regard to the following: (i) the structures of WPT for different applications, (ii) several high-efficiency rectifiers for WPT and radio-frequency identification (RFID) energy-harvesting applications with full active structures, (iii) several high-performance DC to AC converters with zero voltage switch (ZVS) or zero current switch (ZCS) technology, and (iv) several WPT power control methods. A novel single-stage WPT receiver suitable for low-cost IPT WPT operation at 125-kHz is presented in Chapter 3, followed by the design and discussion of a high-efficiency rectifier for 6.78-MHz MRC WPT application in Chapter 4. In Chapter 5, a 6.78-MHz MRC WPT system with hybrid control of frequency-shifting and phase-shifting is proposed and verified. Finally, Chapter 6 concludes the thesis and also provides suggestions for future work.

Chapter 2

Literature Review

2.1 Basic concepts and principle of WPT

2.1.1 Magnetic coupling

Magnetic coupling in a WPT system means transfer of power through a magnetic field. The theory behind magnetic WPT is the Ampere's and Faraday's laws. According to Ampere's law, the displacement current can be neglected in an ideal case:

$$\oint_{C_1} H dl = \oint_{S_2} J dS \quad (2.1)$$

An AC through the primary coils will generate an alternating magnetic field over a closed contour C_1 that encloses the transmitting coils. Assuming that the currents through all turns of the transmitting coils are the same, Equation (2.1) can be simplified to give:

$$\oint_{C_1} H dl = N_1 I_1 \quad (2.2)$$

Equation (2.2) indicates that the magnetic field intensity depends on the number of turns of the transmitting coil and the transmitting current. Faradays law states that when the receiving coil is exposed to an alternating magnetic field, an induced voltage will be created at the receiving coil terminals:

$$\oint_{C_2} E dl = -\frac{\partial}{\partial t} \oint_{S_2} B dS \quad (2.3)$$

Thus power is further delivered to the load. It is clear that the basic concept is simple, but transferring an adequate amount of power from the transmitter to the load efficiently, controllably, and safely is much more difficult. In the text below, ways to achieve this target will be discussed.

2.1.2 Coupled-Mode Theory

coupled mode theory (CMT) is a basic theory for analysing the power exchange between two resonant objects [19]. Based on CMT, $a_1(t)$, and $a_2(t)$ are defined as the amplitudes of two resonant objects in the time-domain field containing energies of $|a_1(t)|^2$ and $|a_2(t)|^2$, respectively. A first-order differential equation between two resonant objects is used to describe the power exchange [20] [21]:

$$\frac{da_1(t)}{dt} = -(j\omega_1 + \Gamma_1)a_1(t) + jK_{12}a_2(t) + F_S(t) \quad (2.4)$$

$$\frac{da_2(t)}{dt} = -(j\omega_2 + \Gamma_2 + \Gamma_L)a_2(t) + jK_{12}a_1(t) \quad (2.5)$$

where ω_1 and ω_2 are the eigenfrequencies of the two resonant objects, and Γ_1 , Γ_2 are the intrinsic decay rates due to the power loss of the coil (ohmic loss) and propagation medium (radiative loss), Γ_L is the decay rate of the power consumed by the load resistance connected to the receiving coil, $F_S(t)$ is the excitation applied to the transmitter, which can be treated as a power source, and K_{12} is the coupling rate between the two resonant objects. Generally, in a WPT system, the two resonant objects should be tuned to the same resonant frequency, and thus $\omega_1 = \omega_2 = \omega$.

In steady-state analysis, it is assumed that $F_S(t)$ is a sinusoidal signal that is equal to $A_S \cdot e^{-j\omega t}$, so that $a_1(t) = A_1 \cdot e^{-j\omega t}$ and $a_2(t) = A_2 \cdot e^{-j\omega t}$. It can be derived

from Equation (2.4) and Equation (2.5) that

$$\frac{A_2}{A_1} = \frac{jK_{12}}{\Gamma_2 + \Gamma_L} \quad (2.6)$$

The ohmic losses of the two resonant objects are $P_1 = 2\Gamma_1|A_1|^2$ and $P_2 = 2\Gamma_2|A_2|^2$, while the power transfer to the load is $P_L = 2\Gamma_L|A_2|^2$. The total power generated from the source $F_S(t)$ is thus $P_S = P_1 + P_2 + P_L$. Therefore, the power transfer efficiency of the two resonant objects can be calculated as the ratio of power delivered to the load to the power from the source.

$$\eta = \frac{P_L}{P_S} = \frac{\Gamma_L|A_2|^2}{\Gamma_1|A_1|^2 + \Gamma_2|A_2|^2 + \Gamma_L|A_2|^2} \quad (2.7)$$

2.1.3 Maximum Power Transfer Principle and Maximum Transfer Efficiency Principle

Figure 2.1: (a) Equivalent circuit of the power transfer system, (b) variations of output power and transfer efficiency as a function of R_L/R_S [22].

Regardless of the structure of the power transfer system, it can be described by an equivalent circuit, as shown in Figure 2.1 (a). The maximum power transfer occurs when the source and load impedance match each other, which means that $R_S = R_L$ and $X_S = X_L$. The transfer efficiency in this condition can be derived as:

$$\eta_{P_{max}} = \frac{i^2 R_L}{i^2 R_L + i^2 R_S} = \frac{R_L}{R_L + R_S} = 50\% \quad (2.8)$$

The variation of the output power and transfer efficiency as a function of the

ratio R_L/R_S is shown in Figure 2.1 (b). For fixed R_S , it is obvious that the transfer efficiency is proportional to R_L/R_S . So the maximum transfer efficiency occurs when equivalent load R_L reaches its maximum allowed value (R_{Lmax}), given by

$$\eta_{max} = \frac{i^2 R_{Lmax}}{i^2 R_{Lmax} + i^2 R_S} = \frac{R_{Lmax}}{R_{Lmax} + R_S} \quad (2.9)$$

From the Equation (2.8) and Equation (2.9), it is concluded that the maximum output power and maximum efficiency do not have the same conditions. In a real design, the tradeoff between these two conditions is important to the designer, and will be discussed later.

2.1.4 Topologies of WPT

2.1.4.1 Two-coil WPT

In WPT research over recent years, a widely adopted structure is the two-coil system consisting of a transmitting coil and a receiving coil, where magnetic power is generated and exchanged between coils. Different compensation structures are used to tune the coils so that they are resonant. Figure (2.2)(a)(d) shows four basic compensation topologies, named series-series (SS), series-Parallel (SP), parallel-series (PS), and parallel-parallel (PP). Here, the relationships between the voltage and current are listed as follows:

Here, the relationships between the voltage and current are listed as follows:

For SS:

$$V_1 = I_1(R_1 + j\omega L_1 + \frac{1}{j\omega C_1}) + I_2 j\omega M \quad (2.10)$$

$$0 = I_1 j\omega M + I_2(R_2 + R_L + j\omega L_2 + \frac{1}{j\omega C_2}) \quad (2.11)$$

For SP:

Figure 2.2: Topologies of four basic compensation network.

$$V_1 = I_1(R_1 + j\omega L_1 + \frac{1}{j\omega C_1}) + I_2 j\omega M \quad (2.12)$$

$$0 = I_1 j\omega M + I_2(R_2 + \frac{\frac{R_L}{j\omega C_2}}{\frac{1}{j\omega C_2} + R_L}) \quad (2.13)$$

For PS:

$$I_1 = V_1 j\omega C_1 + \frac{(V_1 - I_2 j\omega M)}{j\omega L_1} \quad (2.14)$$

$$0 = I_1 j\omega M + I_2(R_2 + R_L + j\omega L_2 + \frac{1}{j\omega C_2}) \quad (2.15)$$

For PP:

$$I_1 = V_1 j\omega C_1 + \frac{(V_1 - I_2 j\omega M)}{j\omega L_1} \quad (2.16)$$

$$0 = I_1 j\omega M + I_2(R_2 + \frac{\frac{R_L}{j\omega C_2}}{\frac{1}{j\omega C_2} + R_L}) \quad (2.17)$$

where the mutual inductance (M) indicates the coupling strength of the coils, and the self-inductance of the transmitting coil (L_1), self-inductance of the re-

ceiving coil (L_2) indicates the energy store capabilities of both coils. The WPT system works in resonance, which means the operation frequency (ω) equals the resonant frequency of both the primary side (ω_1) and the secondary side (ω_2), which is expressed as:

$$\omega = \omega_1 = \omega_2 = \frac{1}{\sqrt{L_1 C_1}} = \frac{1}{\sqrt{L_2 C_2}} \quad (2.18)$$

If the system works in resonant mode and the ESR of the coils are so small that can be omitted, the relationships between the voltage and current for the secondary side are as listed below:

For series compensation:

$$V_2 = I_1 j\omega M \quad (2.19)$$

For parallel compensation:

$$I_2 = I_1 \frac{M}{L_2} \quad (2.20)$$

According to Equations (2.19, 2.20), we can find that the series compensation has a constant-voltage capability, which means that if we keep the primary current constant, the voltage across R_L will also have a constant value, and the parallel compensation also has a current constant capability, which means that if we keep the primary current constant, the current through R_L will also have a constant value. To decrease the voltage-ampere (VA) rating of the WPT system, designers tend to tune the input voltage and current in phase; in other words, a zero phase angle (ZPA) is expected.

According to Equation (2.10) to (2.17), the input impedance Z_{in} seen from the AC source V_1 can be found as follows:

For SS:

$$Z_{in_SS} = R_1 + \frac{\omega^2 M^2}{R_2 + R_L} \quad (2.21)$$

For SP:

$$Z_{in_SP} = R_1 + \frac{\omega^2 M^2}{R_2 + j\omega L_2 + \frac{R_L}{1+j\omega C_2 R_L}} \quad (2.22)$$

For PS:

$$Z_{in_PS} = \frac{1}{\frac{1}{R_1 + j\omega L_1 + \frac{\omega^2 M^2}{R_2 + R_L}} + j\omega C_1} \quad (2.23)$$

For PP:

$$Z_{in_PP} = \frac{1}{\frac{1}{(R_1 + j\omega L_1) + \frac{\omega^2 M^2 (1 + j\omega C_2 R_L)}{(R_2 + j\omega L_2)(1 + j\omega C_2 R_L) + R_L}} + j\omega C_1} \quad (2.24)$$

From the above equations, it can be concluded that only SS structure's ZPA conditions are not related to the mutual inductance and the load resistance. Z_{in} for the SP structure varies with M. For PS and PP, Z_{in} is greatly affected by both the M and R_L conditions.

Figure 2.3: The LCL and LCCL compensation topologies.

Besides these four basic compensation techniques, [23] presents an inductor-capacitor-inductor (LCL) compensation structure (shown in Figure 2.3(a)). An SP-tuned LCL is used in the circuit to achieve a unity power factor and thus to decrease the VA rating and power loss in the coil winding, metal-oxide-semiconductor field-effect transistor (MOSFET) switches, and source resistance. Compared to

the traditional PP compensation technique, the new LCL structure can increase the efficiency by 20%-30% (coil to coil), as demonstrated by the measured results illustrated in Figure 2.4.

Figure 2.4: The experimental results of [23].

An SP inductor-capacitor-capacitor-inductor (LCCL) compensation (see Figure 2.3(b)) technique for WPT is presented in [24]. The author analyses the conditions for both ZVS and ZPA operation with the proposed technique. The proposed ZCS LCC compensation technique has the advantages of lower shut-down losses, lower switching losses, and faster switching transients. The results are verified with a 750-W, 40-kHz LCC compensated wireless charging system.

A brief analysis of the efficiency of the WPT system is described based on the SS compensation model. The analysis adopts a frequency-domain equivalent circuit with only the fundamental component for simplicity.

Based on Equations (2.10) and (2.11), as shown in Figure 2.5, the whole receiver could be coupled into the transmitter side as a so-called reflected impedance Z_r , which is equal to:

$$Z_r = \frac{\omega^2 M^2}{Z_2 + R_L} \quad (2.25)$$

where Z_2 is the impedance of the whole secondary circuit, given by:

$$Z_2 = j\omega L_2 + \frac{1}{j\omega C_2} + R_2 \quad (2.26)$$

Figure 2.5: The equivalent circuit for SS model.

The power transfer efficiency is defined by considering only the real power. When the system works under resonant mode, the efficiency of the receiver side can be simply given by:

$$\eta_2 = \frac{R_L}{R_2 + R_L} \quad (2.27)$$

and the efficiency of the transmitter is:

$$\eta_1 = \frac{\Re(Z_r)}{R_1 + \Re(Z_r)} \quad (2.28)$$

in which $\Re(Z_r)$ is the real part of Z_r .

The efficiency of the whole system is the product of η_1 and η_2 , which is:

$$\eta = \eta_1 \eta_2 = \frac{R_L}{R_1 \frac{(R_2 + R_L)^2}{\omega^2 M^2} + R_2 + R_L} \quad (2.29)$$

2.1.4.2 Four-coil WPT

Unlike the conventional two-coil system, two intermediate multi-turn coils with high Q values are placed between the transmitting and receiving coils. The two intermediate coils work as a repeater in the transformer and were first re-

ported and proved at the MIT laboratory in 2007 [15]. Due to the potential of this new structure to expand the spatial freedom of the WPT system with high efficiency, it has been extensively researched [[14], [25], [26], [27], [28], [29], and [30]]. The main disadvantage of the four-coil structure is the large space cost of the additional two coils. The high Q factor requires more turns of the coil, leading to a bigger space requirement, especially for helical coils.

Figure 2.6: Schematic of four-coil WPT system. [31]

Figure 2.6 shows a schematic diagram of a four-coil WPT system consisting of a transmitting coil ($Coil_1$), receiving coil ($Coil_4$), and two repeaters ($Coil_2$, $Coil_3$). L_2 is coupled with the transmitting coil through a high-frequency AC source that provides power to the WPT system, and L_3 , coupled with the receiving coil, supplies power to the connected devices as a load. The compensation capacitors (C_1 , C_4) are added with the transmitting and receiving coils to make them resonant at the same frequency as the repeaters. k_{12} , k_{23} and k_{34} are the coupling coefficients between neighbouring coils, given by

$$k_{ij} = \frac{M_{ij}}{L_i L_j} \quad (2.30)$$

The couple between non-neighbouring coils is negligible in our analysis. The power transfer efficiency of the WPT system mainly depends on the coupling coefficient between the two repeaters since they are normally located a significant distance away from each other. Figure 2.7 illustrates the equivalent circuit model of the four-coil WPT system. Here, L_2 and L_3 represent the self-inductances of

Figure 2.7: Equivalent circuit model of four-coil WPT system. [14]

the repeaters, *Coil₂* and *Coil₃*, respectively, with their parasitic capacitances C_2 and C_3 . Similarly, the transmitting and receiving coils are represented by L_1 and L_4 , respectively. C_1 and C_4 are the compensation capacitors for L_1 and L_4 . All the coils are electromagnetically coupled to each other by coupling coefficients k_{12} , k_{23} and k_{34} . The state equations are derived and are given by:

$$V_s = I_1(j\omega L_1 + \frac{1}{j\omega C_1} + R_1) + I_2 j\omega \frac{k_{12}}{L_1 L_2} \quad (2.31)$$

$$0 = I_2(j\omega L_2 + \frac{1}{j\omega C_2} + R_2) + I_1 j\omega \frac{k_{12}}{L_1 L_2} + I_3 j\omega \frac{k_{23}}{L_2 L_3} \quad (2.32)$$

$$0 = I_3(j\omega L_3 + \frac{1}{j\omega C_3} + R_3) + I_2 j\omega \frac{k_{23}}{L_2 L_3} + I_4 j\omega \frac{k_{34}}{L_3 L_4} \quad (2.33)$$

$$0 = I_4(j\omega L_4 + \frac{1}{j\omega C_4} + R_4 + R_L) + I_3 j\omega \frac{k_{34}}{L_3 L_4} \quad (2.34)$$

The efficiency of the four-coil system is derived in [14] and is given by:

$$\eta_{4-coil} = \frac{(k_{12}^2 Q_1 Q_2)(k_{23}^2 Q_2 Q_3)(k_{34}^2 Q_3 Q_{4L})}{[(1 + k_{12}^2 Q_1 Q_2)(1 + k_{34}^2 Q_3 Q_{4L})][1 + k_{23}^2 Q_2 Q_3 + k_{34}^2 Q_3 Q_{4L}]} \frac{Q_{4L}}{Q_L} \quad (2.35)$$

2.2 Control Circuits and Methods for WPT

Designing a suitable control for WPT is important and challenging. For consumer devices, most standards [32] [33] require a constant voltage output. Some other works also control the output power, phase angle of the input voltage/current and so on. The WPT control strategies can be classified into two main categories: control at the transmitter side and the receiver. The former requires wireless communication to transfer information of control target from the receiver to the transmitter. The transmitter then can adjust inputs, like DC supply, operating frequency, phase-shift, to control the targets.

2.2.1 Control at the transmitter side

Qi standard is one of the mainstream WPT standards. Figure 2.8 shows the basic concepts of the control in the Qi standard. This control strategy is based on backscattering communication, which is realized by changing the load of the receiver. It can be seen from Figure 2.8 that the current in the transmitter varies with the change of the load, indicating digital bit "0" or "1". As shown in Figure 2.9, the realization of load changing can be resistive or capacitive, as applied in [34].

The communication circuit sends the difference between the output voltage and reference voltage back to the transmitter. The transmitter filters the signal and uses a proportional-integral-derivative (PID) control algorithm to adjust either the

Figure 2.8: Qi wireless power transfer system [34].

Figure 2.9: Two different kinds of communication: (a) resistive (b) capacitive [32].

duty or the frequency of the inverter, as shown in Figure 2.10.

Figure 2.10: PID control algorithm [32].

A4WP is another main WPT standard. The control of the A4WP is shown in Figure 2.11. Although a DC-DC converter is used to realize fast voltage control, the A4WP system contains a blue-tooth communication module, which monitors the rectifier output for performance tracking and efficiency optimization. Since both the rectifier output and voltage across the load are constant, such control is still considered as transmitter control.

Figure 2.11: A4WP wireless power transfer system [33].

In some applications, multiple-target control is required, such as in grid-to-EV WPT. Reference [12] presents a two-way WPT system for grid-EV applications. The power can be controlled bidirectionally. A theoretical model is developed to analyse both the amount and direction of power flow between EVs and grids by carrying out the control according to either the phase or the magnitude modulation of the voltages of the DC input. Figure 2.12 shows an equivalent circuit model of such a system. In this reference, the relationship between the output power of the n_{th} load and the phase angle of V_{son} and V_{pi} is derived and a control system based on the equation is designed. The testing results of a 1.5-kW prototype bidirectional WPT system with a 4-cm air gap prove the validity of the design.

2.2.2 Control at the receiver side

Compare to the control at the transmitter side, control at the receiver side has advantage of a faster response. The most common control circuit in the receiver side is the DC-DC converter. With regard to power management circuits on chip, there are three basic topologies for switching-mode DC-DC converters, called buck, boost, and buck-boost DC-DC converters, respectively.

- The buck converter (Figure 2.13 (a)) is utilized to provide an output voltage lower than the input voltage for applications where step-down voltage

Figure 2.12: Wireless power transfer system in [12].

conversions are required.

- The boost converter (Figure 2.13 (b)) is utilized to provide an output voltage higher than the input voltage for applications where step-down voltage conversions are required.
- The output voltage of a buck-boost converter (Figure 2.13 (c)) can be either higher or lower than the input voltage for applications where both step-up and step-down voltage conversions are required.

We take the buck converter as an example to analyze the transfer function of the DC-DC converter. During the phase when S_1 is on and S_2 is off, the inductor stores energy at a rate of $(V_{IN}-V_{OUT})$ for a duration of D ; in the phase where S_1 is off and S_2 is on, the inductor releases energy at a rate of V_{OUT} for a duration of $(1-D)$. At steady state, the energy exchange of the inductor should be equal to zero in each cycle, which leads to:

$$D \bullet (V_{in} - V_{OUT}) - (1 - D) \bullet V_{OUT} = 0 \quad (2.36)$$

Figure 2.13: Basic DC-DC Topologies.

From the above equation, the transfer function of buck DC-DC converter is given by:

$$\frac{V_{OUT}}{V_{IN}} = D. \quad (2.37)$$

The transfer functions of the boost converter and buck-boost converter can be derived in the same way, and are given by Equation (2.38) and (Equation 2.39), respectively.

$$\frac{V_{OUT}}{V_{IN}} = \frac{1}{1-D} \quad (2.38)$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{D}{1-D} \quad (2.39)$$

For many applications, a stable output voltage is required in a large dynamic range of loads. So a controller is always employed. There are several control methods for DC-DC: voltage mode PWM control, current mode PWM control, pulse frequency modulation (PFM) control, and hysteretic control. Generally, the DC-DC converter works in the continuous conduction mode (CCM) and discontinuous conduction mode (DCM).

The voltage mode PWM control adjusts the duty cycle of the DC-DC converter to maintain the output voltage with a fixed frequency. Figure 2.14 shows a typical voltage mode PWM-controlled DC-DC converter. The duty cycle can be started either by a clock signal or by the output of the comparator, as shown in Figure 2.14.

Figure 2.14: Voltage mode PWM buck DC-DC converter [35].

The current PWM-control schematic is shown in Figure 2.15. This control method can be further classified into peak current control and valley current control. Figure 2.15 shows the two different kinds of current control mode topologies. The difference between them is that the duty cycle is initiated by the clock signal in peak current mode, whereas in the valley current mode the duty cycle is initiated by the comparator signal. Generally, peak current mode is adopted in applications where the duty cycle is large (>0.5), since the current sensor needs a longer response time; similarly, valley current mode is used when the duty cycle is smaller than 0.5. A slope compensation ramp is used for both modes to reduce the sub-harmonic distortion. In addition to these two modes, the average current control mode and the charge control mode are also widely used [36].

In PFM control mode, the switching frequency of the DC-DC converter varies. It is quite effective for improving the efficiency of the converter in light load con-

Figure 2.15: Current mode PWM buck DC-DC converter:(a) peak current control and (b) valley current mode control [35].

ditions. Figure 2.16 depicts a fixed on-time PFM mode buck converter [37]. A fixed on-time means that in each switching period, the on-time of the main switch is the same, while the switching period changes due to the difference between V_{IN} and V_{OUT} . The other widely used PFM control strategies are fixed off-time PFM, single-pulse PFM, and multi-pulse PFM.

PFM control DC-DC has a much better conversion efficiency under light loads. However, a prominent problem of PFM mode converters is that electro magnetic interference (EMI) problems occur due to the changing switching frequency. The unpredictable frequency adds to the design complexity of the input filter. pulse skipping modulation (PSM) is thus developed to solve the existing problem. An example of this sort of converter is shown in Figure 2.17 [38]. In most PSM applications, the switching is fixed or predictable, making EMI design easier.

A hysteretic mode DC-DC converter is a special PFM. Although this kind of converter also has variable frequency, it is introduced separately here from PFM because its frequency is normally not load dependent and it is widely applied in

Figure 2.16: PFM control buck converter with fixed on-time control [35].

Figure 2.17: PSM control buck converter in [38].

CCM operation.

Figure 2.18 [39] shows a structure for voltage mode hysteretic control. The difference between V_{b1} , V_{b2} and V_C decides the mode in which the converter works.

Several DC-DC converters have been developed for WPT systems in recent years.

Figure 2.18: Hysteretic control DC-DC converter in [39].

Figure 2.19: An average-current mode DC-DC with PSM for a 6 W WPT in [40].

Figure 2.19 [40] shows an average current-mode converter for a 6-W output WPT application. Average current mode is adopted due to its better noise immunity, absence of extra slope compensation, accurate regulation, and good performance in both CCM and DCM modes. These benefits make this kind of converter a good choice for WPT with a DC input from the rectifier. To improve the performance in light load conditions, the PSM control mode is adopted. As a result, the efficiency of the converter is over 90% for the whole range of load conditions.

Reference [41] presents a DC-DC hysteretic buck converter with a Phase Locked Loop (PLL) (Figure 2.20). The implanted PLL helps the DC-DC to maintain the operating frequency in a large dynamic range of process, voltage and temperature (PVT) variations. Besides, with adaptive control of the PWM/PFM mode, the efficiency of the converter can be maintained from full current condition to light current condition. The maximum efficiency of the converter can reach 92.7% and the maximum of DC-DC converter is 6 W.

There are also some other receiver voltage control circuits except for DC-DC converter. Reference [42] presents an active reconfigurable voltage recti-

Figure 2.21: The schematic of active reconfigurable voltage rectifier/doubler rectifier in [42].

is to control the output voltage of the diode bridge up and down every eight resonant cycles with integral cycle mode, where on or off switching is determined on the output voltage. The efficiency of the circuit reaches 86% at an output power of 3.4 W. The idea of the proposed 3R is to control V_{XH} in Figure 2.22 between the on-duty state and the off-duty state. During the on-duty state of the circuit, M_2 is turned off and both M_1 and M_3 are turned on. For this state, the circuit is configured as a full-bridge rectifier with a smoothing capacitor consisting of C_{FL} and C_{OUT} connected in parallel. During the off-duty state, M_2 is turned on and both M_1 and M_3 are turned off. So the load of the circuit is switched to C_{FL} in series with $C_{OUT} \parallel$ load. The energy and current of the receiver change consequently. PWM voltage control is used to control the circuit operating between the two states to maintain the output voltage at 5 V.

Reference [44] reported a 13.56 MHz WPT receiver consisting of a 1X/2X reconfigurable rectifier with voltage regulation for biomedical implants. The highlight of this research is that the receiver control the voltage in two control loops: 1) a PWM loop at the receiver side controls the duty cycle of the receiver between the so called 1X and 2X modes; and 2) an outside control loop with wireless communication with the information on the ratio of 1X to 2X, and thus the power of the transmitter is adjusted according to load and coupling variations. Two dif-

Figure 2.22: The schematic of the 3R rectifier in [43].

ferent communication techniques with backscattering method are presented for a fast response and energy-saving purpose. The transmitter and receiver of the WPT are fabricated in $0.35\ \mu\text{m}$ CMOS process. The digital circuits are implemented using field-programmable gate array (FPGA). The measured maximum output power is 102 mW and the maximum receiver efficiency is 92.6 %. For load transients, the overshoot and the undershoot are approximately 110 mV and the settling times are less than $130\ \mu\text{s}$.

Figure 2.23: The circuit model of the 1X/2X reconfigurable rectifier in [44].

Except voltage control in the receiver side. In some applications, the phase angle is controlled to tune the WPT system at resonance. Reference [45] proposes an active tuning method with a tri-state boost converter for a parallel-compensated secondary side circuit. Both active power and reactive power can be controlled by discontinuous current PWM modulation with the novel converter by synchronizing the switching period within the resonant period of the WPT system. A detailed circuit model of the design is shown in Figure 2.24.

Figure 2.24: The circuit model of the tri-state boost converter in [45].

2.3 High efficiency rectifier design

The rectifier is a circuit that converts AC voltage to a DC voltage. In a wireless power transfer system, the rectifier regulator is mainly used in the receiver to change the induced voltage to supply the post-stage circuits. The efficiency of the rectifier is a vital consideration in WPT system design since it is related to the heat emission of the portable electronic devices, which should be restricted in a strict range. A traditional diode bridge rectifier suffers from loss caused by the forward voltage of the diodes, so it is not considered suitable for applications discussed in this thesis. Full-active rectifier, which means that the diodes would be replaced by the transistor and the controller and driving circuits. Since the on-resistance of a transistor is usually small, the power loss on the transistor is much smaller than

the diode. However, the active rectifier has an extra loss source-reverse current. As shown in Figure 2.25, N_1 - N_4 are four transistors of the rectifier, in the positive half period, N_1 and N_4 are turned on. If the N_1 and N_4 are failed to turned off in time. The current will flow from V_{REC} to V_{AC} , which incur reverse leakage loss. Several kinds of rectifiers have been presented by researchers to improve the efficiency by reducing or eliminating the reverse current.

Figure 2.25: The explanation of reverse current.

Figure 2.26: The circuit model of [46].

Reference [46] presents an active cross-connection full wave bridge rectifier for wireless power transfer in 13.56 MHz RFID applications. The circuit consists of two cross connection PMOS and two symmetrically placed NMOS active diodes (shown in Figure 2.26) which are dynamically biased and controlled by a fast voltage comparator with reverse-current control to decrease the loss. The comparator with a 2-ns delay is a four-input circuit with a common gate structure input stage that can drive the NMOS switch working as a diode without forward

voltage.

The proposed rectifier with an output power of 5.7 mW has an efficiency as high as 85 %. The circuit is fabricated in a $0.35\ \mu\text{m}$ CMOS process with an active area of $0.0055\ \text{mm}^2$.

Figure 2.27: The schematic of active rectifier with speed-up latch in [47].

In [47], a highly efficient rectifier application with active MOSFETs using a speed-up common-gate-type cross-coupled latched comparator is presented for biomedical implant applications. Compared to the design in [46], the proposed rectifier introduces an unbalanced transistor and an additional comparator to accelerate the turn-on/turn-off speed of the MOSFETs. As shown in Figure 2.27, when the voltage potential of RF+ drops below DCout, M_{P2} is turned ON, while M_{N2} is turned OFF, thus pulling the OUT_{PM} to DCout. An additional pMOSFET M_{P3} is connected parallel to M_{P2} for to enhance the pull-up effort, leading to a short response time, which minimize the leakage current.

To accelerate the transient response of the comparator, a speed-up comparator is placed parallel to the main comparator, and are coupled to the main comparator with C_{C1} , C_{C2} , C_{C3} and C_{C4} . when RF+ increases above DCout, M_{P5} is ON, whereas M_{N4} is OFF, pulling the node "X" to high voltage, which speeds up the turn-on action of M_{N2} and turn-off action of M_{P2} . A similar process happens when RF+ drops below DCout, turning M_{N4} ON while turning M_{P5} OFF, pushing "X" to ground. Such actions help to turn M_{P2} ON harder, contributing to a quicker

pull-up of OUT_{PM} . The designed rectifier achieves a maximum measured PCE of 81.9 % at 13.56 MHz under conditions of a low $1.5\text{-}V_{pp}$ RF input signal with a $1\text{-k}\Omega$ output load.

However, the solution in [47] cannot eliminate the reverse current since the comparator will not respond until the input AC voltage goes across the output DC voltage. A rectifier with unbalanced-biased comparators has been designed in [48]. The circuit schematic of the rectifier is shown in Figure 2.28.

The highlight of the research in [48] is that the trigger voltage for turning off the nMOSFETs (M_{N1} and M_{N2}) of 0 V is replaced with $-\Delta V$. This design helps the comparator to turn off the transistors before the AC input drops below DC output, eliminating the reverse current as a result. Figure 2.28(b) depicts the scheme of the unbalanced comparator, which comprises two pairs of common-gate current amplifiers CGA1 and CGA2, a current mirror (M_9 and M_{13}), and a gate driver (M_{14} , M_{15} , M_{16} , and M_{17}). The unbalanced biasing is created by setting $I_{S1} > I_{S2}$ and ΔV is given by:

$$\Delta V = \frac{\sqrt{I_{S1}} - \sqrt{I_{S2}}}{2\sqrt{\alpha}}. \quad (2.40)$$

where α is given by:

$$\alpha = (1/2)\mu_n C_{OX}(W/L). \quad (2.41)$$

and (W/L) is the size of the transistors in CGA₁ and CGA₂. When V_{in2} rises above $-\Delta V$, the current of M_7 is mirrored to M_9 , while current of M_{10} is mirrored to M_8 , generating a push-pull effect at node "A". In a conclusion, by carefully designing I_{S1} and I_{S2} , M_{n2} can be fully turned off at 0 V, thus removing the reverse leakage current. The proposed circuit is able to provide a maximum output current of 20 mA and operate properly with inputs of different amplitudes and frequencies. With a 1.5 MHz input of 1.2 V amplitude, the proposed rectifier can achieve the peak voltage conversion ratio of 95% and the power efficiency of at least 82%.

Figure 2.28: The schematic of the rectifier with unbalanced biasing design in [48].

Reference [41] reported a high-efficiency active rectifier with a delay locked loop (DLL) for 6.78 MHz WPT applications. The schematic is shown in Figure 2.29.

The idea of a high-speed comparator is suitable only for those applications with small output currents, for example in the range of tens of millamperes to hundreds of millamperes. For wireless power transfer in mobile phones or tablets, the large output current needs be handled by large-sized power transistors, which lead to large input capacitance of the transistors. As a result, the comparator cannot turn off the upper-side transistor in a short given time due to the large capacitance. Consequently, the power loss due to such a leakage cur-

Figure 2.29: The schematic of a high-efficiency active rectifier with DLL in [41].

rent may bring serious problems. The proposed active rectifier uses the DLL to compensate for the delay caused by the gate driver delay and other internal circuits. Moreover, to accelerate the shutting-off process, a negative impedance circuit was utilized to accelerate the circuit. When the magnitude of the AC input voltage is 8.95 V of 6.78 MHz, the maximum efficiency of the proposed active rectifier is 91.5 % .

Based on the references quoted in the previous section, it is apparent that the active rectifier with cross-couple structure with fast comparator design had drawn the attention of researchers ever since the 13.56 MHz RFID energy harvesting implementation. However, it is challenging to transfer such technology to the megahertz WPT for high power application since the extra power leads to larger transistor sizes. The increased sizes further incur larger delay of the comparator causes reverse current to happen, which deteriorate the efficiency of the rectifier. Ref [3] indicates that the delay time control would be suppressed for megahertz WPT application. However, the DLL would be slow when a load transition happens and the replica unit in [41] also caused area overhead in the design. In Section 4 of Chapter 4, the author proposed a rectifier with adaptive time delay control for the 6.78 MHz WPT with 91.8 % peak efficiency and 15 W output power. Detailed analysis and design description would be given in Chapter 4. A short summary is given at Table 2.1, the advantages and problems of each structure are listed in this table as well.

Table 2.1: A summary about the rectifier

	[49]	[48]	[42]	[43]	[44]	[40]	[41]
Structure	Active Diode with RCC	Fast Comp.	Dual Mode Rect.	3R Rect.	Dual Mode Rect.	Half Active Rect.	DLL control Recti.
V_{out} (V)	3.5	1	3.1	5	3.6	5	8.6
$P_{out_{Max}}$ (mW)	6.8	5	6	6000	102	6000	6000
Resonant f. (MHz)	13.56	1.5	13.56	6.78	13.56	6.78	6.78
Rectifier Peak eff.	87%	87.5%	77%	86%	92.6%	89.2%	91.5%
Pros.	Almost zero reverse current	Zero reverse current	Zero reverse current	Zero reverse current	Almost zero reverse current	Zero reverse current	Zero reverse current
Problems for high power WPT	Reverse current due to high power	Reverse current due to high power	Diode loss	Diode loss	Reverse current due to high power	Diode loss	Extra area and Low Speed

2.4 Frequency Splitting

Frequency splitting is a widely known phenomena in resonant wireless power transfer system. The concept of frequency splitting is from a mechanics model of two pendulums connected by a spring, which can be analogous to two LCR resonator connected by magnetic coupling [50]. In this single oscillating mechanic example, the two pendulums can have the largest oscillate amplitude in two frequencies, one of higher and one of lower frequency, which will be called as peak upper and lower peak frequency in the rest of this thesis. More importantly, the frequency separation of the peak frequencies is dependent on the coupling of the spring. As this coupling decreases, the frequency separation also decreases until the two peak frequencies converge to the resonate frequency of the system. This indicates that a coupling system can have peak output amplitude in different frequencies, which is a function of coupling. When it comes to WPT system,

coupling is dependent on the distance between the transmitting and receiving coils.

[25] gave a detailed analysis on frequency splitting of MHz resonant WPT system based on a four-coils model shown in Figure 2.6. [25] takes the scattering parameters (S21), which is the forward voltage gain of the system, as the indicator of the figure of merit (FOM) of the system and call S21 as efficiency. According to both simulation and experimental experiment, shown in Figure 2.30, the critical coupling is defined as coupling between coil2 and coil3 when the system has the maximum S21, expressed as:

Figure 2.30: The experiment results of critical coupling of 5-10 MHz WPT in [25](a) Comparison of (black dots) experimental data to the calculation results of the system; (b) Measured and Predicted critical coupling.

$$k_{critical} = \frac{1}{Q_{coil}} + (k_{lc})^2 Q_{loop}. \quad (2.42)$$

in which it is assumed that the coupling between coil1 and coil2, coil3 and coil4 is fixed, defined as k_{lc} .

Although this paper presents a deep understanding of the underlying principles of coupled WPT system with a simple circuit model, a derivation of the optimum of forward voltage gain based on the system transfer function and a detailed explanation of frequency splitting, this paper takes S21 as efficiency incorrectly. The definition of the efficiency should be the ratio of the output power consumed on the load and the transmitted power from the AC source. Based on

the conclusion of [25], [51] [52] further discussed the efficiency of a two-coil wireless power transfer system, which is expressed as:

$$\eta = \frac{T_Q^2}{(1 + S_M)(1 + L_M) + T_Q^2 + \frac{1+S_M}{1+L_M} F_D^2} \frac{L_M}{1 + L_M}. \quad (2.43)$$

where

$$\begin{aligned} S_M &= \frac{R_S}{R_0} & L_M &= \frac{R_L}{R_0}. \\ T_Q &= \frac{\omega M_{12}}{R_0} & T_Q &= \frac{1}{R_0} \left(\omega L_0 - \frac{1}{\omega C_0} \right). \end{aligned} \quad (2.44)$$

The optimal L_M is calculated by taking $\partial \eta / \partial L_M$, giving that:

$$L_{opt} = \sqrt{\frac{1 + S_M + T_Q^2}{1 + S_M}}. \quad (2.45)$$

Although these papers derive the condition where the system can achieve the maximum efficiency, the authors failed to considered the output power in such condition. Actually, the output power at such condition would be very low when the resonant frequency is several megahertz. The discussion of the trade-off of the output power and efficiency would be expanded in the chapter 5 in this thesis.

2.5 Power Amplifier

2.5.1 Class D Amplifier

Class D DC-AC RF resonant power amplifier was firstly reported by Baxandall in 1959 [53], and have been developed in different sorts of applications to convert DC power to AC power. A typical Class D half-bridge amplifier is shown in Figure 2.31.

It is composed of two active switches S_1 and S_2 and a load resistor with a LC resonator. S_1 and S_2 are turned on and off alternately with a 50% duty cycle or a little less, so the output voltage is switched from V_I and ground, realizing a DC to AC conversion.

Figure 2.31: A typical Class D half-bridge amplifier:(a) Schematic.(b)Equivalent circuits.

Figure 2.32: A Class D amplifier with power control in [54].

Reference [54] proposes a variable duty cycle power control algorithm for a fixed frequency class D amplifier for induction heating applications. The schematic of the circuit and control flow is shown in Figure 2.32. The difference between the load current i_{Leq} and the desired power $P_{desired}$ determines D_{power} . After synthesis with a fixed high frequency to reduce spikes in low frequency, the signal

triggers the gate drive to control the output power of the amplifier.

Figure 2.33: The analytical model in [55].

To improve the efficiency of the Class D amplifier, [55] presents an analytical model. In this reference, the circuit transitions during different working situations are analysed in details, as illustrated in Figure 2.33(b) and (c). After the analysis, detailed equations of the current and voltage of each transition are listed in this paper. This model accelerates the design procedure of a Class D amplifier. In addition, this reference also developed adaptive control emulation, which helps to improve the performance of the amplifier.

Reference [56] gives analytical expressions of a Class D amplifier in steady state for the ZVS condition in any duty cycle and proves them with a practical circuit. The output power capability, voltage and current waveforms, switch voltage, and current stress are shown in this paper. The main contribution of this reference is that the analysis is based on a variable compensation capacitor, which can be seen in Figure 2.34.

2.5.2 Class E Amplifier

Class E amplifiers are suitable for high-frequency DC-AC power conversion because they can easily achieve either ZVS or ZCS. A schematic of a basic ZVS Class E DC-AC converter is shown in Figure 2.35.

Figure 2.34: Circuits and Waveforms in [56].

Figure 2.35: A typical Class E amplifier: (a) Schematic. (b) Equivalent circuits.

The basic design idea is that the current and voltage of the transistor are non-overlapping during the operation, so the switching loss of the converter should theoretically be zero, leading to a very high efficiency. The amplifier comprises a power transistor, an LC resonator, a shunt capacitor C_1 , a choke inductor L_f , and a load resistor R . Usually the choke inductor should be sufficient large so that it can be treated as a constant current source in the circuit. When the switch is off, the DC power source will charge the LC resonator and the load; when the switch is on, the energy stored in the LC resonator releases the power to the C_1 , and since the resonant frequency is different between the LC resonator and the L, C, C_1 series network, the parameters of the circuit should be designed to make sure that the voltage of C_1 will be exactly zero at the end of the switch-off mode,

resulting in a ZVS action.

Figure 2.36: Analysis of Class E amplifier at different loads and frequency in [57].

Reference [57] presents an example of a Class E amplifier application for a WPT system. This reference analyses the optimized conditions for a Class E amplifier at different loads and frequencies, as shown in Figure 2.36. Equations for the optimization have been derived for both transmitter and receiver. An experimental platform with an output power of 3.7 W and efficiency of 66% has been established to prove the design.

Reference [58] applied the Class E technique in both the transmitter and the receiver. The simulation results showed that the circuit could achieve ZVS/ZDS conditions for both sides with an output power of 5 W and an overall efficiency of 65.9%. In this reference, reflected load theory (RLT) has been used to combine the receiver with the transmitter when analysing the optimized conditions for the Class E amplifier, while combining the transmitter with the receiver when analysing the optimized condition for the Class E rectifier. It shows that for the amplifier, the optimized condition relies on the equivalent reflected impedance as shown in Figure 2.37(c), while for the receiver, it depends on the equivalent impedance of the rectifier and the load.

Reference [59] extended the analysis of a CLASS Class E amplifier not only to ZVS, ZDVS, ZCS and ZDCS conditions, but also to an application without a

Figure 2.37: Equivalent circuits model Class E- Class E WPT system in [58].

tedious tuning process. The main contribution of this reference is that it takes the coil turns as a main parameter of the whole design procedure. Meanwhile, the Q value of the transmitter has been used to analyse the total harmonic distortion (THD) and the output power of the system. A detailed example is shown in Figure 2.38. It can be seen that the high-frequency harmonics need to be considered as the Q_1 decreases. The resonator consisting of the coil and capacitor works as a filter, whose the performance of which deteriorates as the Q decreases. The validity of the analysis has been confirmed from the simulation and experimental results.

Figure 2.38: Design curves with the parameter Q_1 of [59].

2.6 Summary

In summary, the WPT technology has a lot of research directions , which include:

- WPT topologies design. In this field, the research mainly focuses on the maximum efficiency design or maximum output power design.
- High performance circuits design. In this field, for the transmitter design, the research mainly focuses on the ZVS/ZCS conditions design of the power amplifier; for the receiver design, high-efficiency rectifier and DC-DC design are important in the WPT system.
- Smart control design for WPT. In this field, the main objects of control include the voltage, amount and direction of power, and phase angle between the voltage and current.

In the next chapters, a single-stage AC-DC regulator for 125-kHz WPT, a full-active rectifier for 6.78-MHz WPT and a frequency/phase control design will be presented.

Chapter 3

Design of A Single-Stage AC-DC

Voltage Regulator for 125 kHz

Wireless Power Transmission

3.1 Introduction

A basic WPT system consists of a transmitter and a receiver, as shown in Figure 3.1.

Figure 3.1: Wireless power transfer (WPT) system for consumer devices.

In the transmitter, the power amplifier (PA) transforms input DC voltage to a

high frequency AC voltage that drives the transmitting coil to generate AC magnetic fluxes. In the receiver, the coupled AC voltage is rectified into DC voltage (V_{REC}) by a rectifier. In real-life WPT applications, the coupling coefficient of the two coils varies with the distance and the alignment between them. Moreover, the output current changes with time depending on the capacity of the battery. Both of these factors cause the V_{REC} to be unstable. Since most WPT standards for consumer electronic devices require the charger to be able to provide a constant voltage, like Qi standard [32], output voltage regulation is a key function of the system. The regulation can be achieved at either the transmitter [34] or the receiver [42]. For the former, wireless data communication blocks are required in the system, which essentially increase the design complexity and power loss. For the latter, the conventional approach uses a two-stage regulator consisting of a rectifier and a step-down converter, which can either be a low dropout regulator (LDO) with low efficiency at high voltage drop or a DC-DC converter with an inductor, but it is nonetheless difficult to be incorporated into an integrated circuit (IC). The resonant regulating rectifier (3R) structure was reported in [43] to address the aforementioned challenges; however the passive rectifier reduces the efficiency of the receiver.

In this chapter, a novel structure of AC-DC voltage regulator is proposed. The whole system structure is shown in Figure 3.2. This regulator consists of four power MOSFETs and a controller. The details of the circuit will be described in the following sub-sessions. As opposed to the existing systems mentioned earlier, this method does not need any wireless communication blocks and a step-down DC-DC converter. The regulator with a maximum output power of 15 W is able to provide over 90 % efficiency over a large output power range.

The rest of the chapter is organized as follows. A detailed analysis of the feasibility and advantages of the design is presented in Section 3.2. In Section 3.3, the operating principle of the single-stage regulator with PWM control is described and the relationship between the output voltage and switching duty

Figure 3.2: Proposed WPT system with proposed single-stage regulator.

ratio is derived. In Section 3.4, several important circuits and related techniques are presented and analyzed. Measurement results are presented and discussed in Section 3.5, followed by the concluding remarks in Section 3.6.

3.2 Feasibility and Performance Analysis of the Novel Regulator

As discussed before, voltage regulation at the transmitter incurs additional silicon cost as a result of complicated digital communication circuit blocks. Furthermore, the response of the regulation is not sufficiently fast as it is limited by the low data transfer rate of wireless communication technology. Since charging the consumer portable device requires rapid response, the regulation has to be included in the receiver, whereby a step-down DC-DC switch mode converter with a bulk inductor is to be used to regulate the voltage with high efficiency but the inductor is difficult to be integrated into a chip.

The idea of the proposed design originates from the basic operating principle of switched-mode power converter, which works on two basic configurations by switching storage elements (e.g. inductor) into different electrical states.

Take a step-down DC-DC converter for example, as shown in Figure 3.3. During the first configuration when S_1 is on while S_2 is off, the inductor L_1 builds up energy from the DC source. During the second configuration when S_1 is off while S_2 is on, the DC input voltage source is disconnected, and the inductor release the energy to the load in order to maintain the output voltage, which is

usually called a freewheel mode [60], [61], [62], [63] and [64].

Figure 3.3: Operation of a traditional buck DC-DC Topologies.

In most cases, a PWM control algorithm is used to alternate the operation of the converter between the two configurations to achieve an optimal and accurate voltage regulation [65], [66], [67], [68] and [69]. The bulk inductor as a passive element plays a key role to store and provide the power for both configurations. There are in fact several other passive elements such as coils and capacitors, which can be utilized in WPT system.

In this work, we exploit the operating characteristics of the dual-configuration power converter as the seed idea for our proposed voltage regulated WPT system. During the charging configuration, the DC input transfers the power to the load and passive elements. During the non-charging configuration, no power is dissipated and the energy stored in the passive elements maintains the output voltage. It is crucial to design a high efficiency non-charging configuration since it determines the overall efficiency of the entire system.

3.2.1 The Design of Non-Charging Configuration

The non-charging configuration should have two important criteria: 1) The power from the transmitter should not reach the load; 2) The power loss should be minimal. One option of non-charging configuration, as shown in Figure 3.4 (a), is to disconnect the load in the receiver so as to cut off any power transmission from the transmitter, and the power loss in the receiver is close to zero. However, in this case, the current in the transmitter is substantially large since the coil and compensated capacitor (C_2) form a short circuit loop at resonant frequency at the

transmitter end, which in turn leads to substantially high current in the transmitter, potentially jeopardizing the operation of the transmitter. It is imperative to limit and control the transmitter current when the load is not being charged.

Figure 3.4: Two different non-configurations in WPT system (a) Disconnection mode; (b) Freewheeling loop mode.

To achieve low transmitter current for the non-charging configuration, a new topology is proposed and presented in Figure 3.4 (b). In this configuration, the receiving coil and its compensated capacitor form a short circuit loop with the load and the DC capacitor is disconnected. The voltage-current equations of this configuration is as follows:

$$\begin{cases} V_{in} = I_1(R_1 + j\omega L_1 + \frac{1}{j\omega C_1}) + I_2 j\omega M \\ 0 = I_2(R_2 + j\omega L_2 + \frac{1}{j\omega C_2}) + I_1 j\omega M \end{cases} \quad (3.1)$$

In Equation 3.1, V_{in} is the input equivalent AC voltage, L_1 , L_2 , C_1 and C_2 , are the inductance of coils and capacitance of compensated capacitors respectively. R_1 and R_2 are the equivalent series resistance of coils. I_1 and I_2 are currents of the transmitter and receiver. In a resonant WPT system, ideally, the transmitter and receiver resonant frequencies are identical as given by

$$\omega = \frac{1}{2\pi\sqrt{L_1C_1}} = \frac{1}{2\pi\sqrt{L_2C_2}} \quad (3.2)$$

From Equation 3.1 and Equation 3.2, we can get the value of I_1 and I_2 are given by

$$\begin{cases} I_1 = \frac{V_{in}}{R_1 + \frac{\omega^2 M^2}{R_2}} \approx \frac{V_{in}}{\frac{\omega^2 M^2}{R_2}} \\ I_2 = \frac{I_1 j \omega M}{R_2} \approx \frac{V_{in}}{\omega^2 M^2} \end{cases} \quad (3.3)$$

On contrary to the disconnection mode, it is apparent from Equation 3.3 that both I_1 and I_2 are finite values. In that case, the power loss in the non-charging configuration could be controlled by carefully design the value of I_1 and I_2 . To further elaborate the difference between the non-connection mode and free-wheeling mode, the design equations and current of both transmitter and receiver are listed in Table. 3.1 And during the process of optimize the I_1 and I_2 , it is important to keep the output voltage at the load adequately high for a 5 volt application. The definitions and criterions of the high efficiency operation area (HEOA) and the safe operation area (SOA) will be given and discussed in the next sections.

Table 3.1: Design equations of Non-Connection and Free wheeling Mode

Mode	Non-Connection	Free wheeling
Equations		
Transmitter Current	$I_1 = V_{in} \setminus R_1$	$I_2 = 0$
Receiver Current	$I_1 = (V_{in})R_2 \setminus (R_1R_2 + \omega^2 M^2)$	$I_2 = (V_{in})\omega M \setminus (R_1R_2 + \omega^2 M^2)$

3.2.2 High Efficiency Operation Area Analysis

For the charging configuration, a prior condition of having sufficient output voltage across the load when the output current is at its maximum is required.

The voltage-current equation of charging configuration is as follows

$$\begin{cases} V_{in} = I_1(R_1 + j\omega L_1 + \frac{1}{j\omega C_1}) + I_2 j\omega M \\ 0 = I_2(R_2 + R_L + j\omega L_2 + \frac{1}{j\omega C_2}) + I_1 j\omega M \end{cases} \quad (3.4)$$

Since there is a limitation on the voltage level of the input for safety considerations, the voltage gain is required to be high enough for a 5 V application. Based on Equation (3.2) and (3.4), the voltage gain of the system during charging is derived as

$$Gain = \frac{I_2 R_L}{V_{in}} = \frac{\omega M R_L}{R_1(R_2 + R_L) + \omega^2 M^2} \quad (3.5)$$

For the non-charging configuration, all power consumption should be considered as power loss, which is listed in Equation 3.6. This power loss should be an acceptably small fraction of the output power.

$$P_{Loss} \approx \frac{V_{in}^2}{(R_1 + \frac{\omega^2 M^2}{R_2})^2} R_1 + \frac{V_{in}^2}{(\frac{\omega^2 M^2}{R_2})^2} R_2 \quad (3.6)$$

In the design of the non-charging circuit configuration, a sufficiently large voltage gain and a minimal power loss are two key requirements of the system. Based on the measurement data of the coil (R_1 , R_2 and M), it can be considered that the term “ $\omega^2 M^2$ ” of Equation (3.5) is the dominant factor. As a result, the voltage gain of WPT system is inversely proportional to the mutual inductance (M); while Equation (3.6) suggests that the power loss is inversely proportional to M . The voltage gain and power loss trade-off makes the value of M a crucial design parameter.

For voltage gain, the recommended input DC voltage by Qi standard is 19 V and the output voltage is usually 5 V for most consumer electronic devices. In this regard, the voltage gain should not be less than 0.26. For power loss, a maximum of 1 W, which is 5 % of the output power, is set for the proposed design. And this is the definition of the high-efficiency operation area (HEOA).

Table 3.2: Mutual Inductance from 4 mm to 8 mm

Distance (mm)	M (μ H)
4	8.36
5	7.93
6	7.58
7	7.24
8	6.91

The experimental results of the charging coil show that the resistors of coils at 125 kHz are around 0.06 ohms. In this design, a 5V/3A output is required, so the R_L is set to 1.6 ohms. The Gain vs. M and Power Loss vs. M curves are plotted in Figure 3.5. In the range of 0 μ H - 8.4 μ H, the voltage gain is larger than 0.25; in the range of 6.8 μ H-12 μ H the power loss is less than 0.8 W. So the HEOA of the WPT system is limited to within the red dotted rectangle, which is from 6.8 μ H to 8.4 μ H.

Figure 3.5: Voltage Gain vs. M and Power Loss vs. M.

For kilo Hertz WPT applications, the reasonable charging distance is around 5 mm. The mutual inductance of this distance range is tested with a pair of coils of the same size and with a self-inductance of 11 μ H. Table 3.2 provides the experimental results for the required mutual inductance for the distance range of

4 mm - 8 mm. It shows that in this range, the mutual inductance is located inside the HEOA, which corroborates the feasibility of this design.

3.2.3 Safe Operation Area Analysis

The two basic safety considerations in WPT system are currents through the coils, switches and voltages across the compensated capacitors [70], [71] and [72], the equations for the current in the coils are the same as Equation (3.3) and equations of the voltages cross the capacitors are given by

$$\begin{aligned} V_{C1} &= I_1 \frac{1}{j\omega C_1} \\ V_{C2} &= I_2 \frac{1}{j\omega C_2} \end{aligned} \quad (3.7)$$

Based on the values of self-inductance of the charging coils, the values of V_{C1} and V_{C2} can be calculated. The simulation results of I_1 , I_2 , V_{C1} and V_{C2} are shown in Figure 3.6. The current is in the range of 0 A to 5.5 A, while the voltage is in the range of 0 V to 90 V. This range is considered to be safe for both coils and capacitors.

Figure 3.6: (a) I_1 and V_{C1} vs. M (b) I_2 and V_{C2} vs. M .

3.3 Proposed Single-Stage Voltage Regulator for Wireless Power Transfer

As mentioned in 3.2.1, if the receiver works in the configuration as depicted in Figure 3.4(b), it would provide a low power loss and safe non-charging configuration. In the proposed single-stage voltage regulator, four active switches are used to replace the diodes and a PWM controller is used to make these four switches to work between the charging and non-charging configuration to regulate the voltage. The detailed operation principle and the relationship between output voltage and the duty cycle of the charging configurations are analyzed with a phasor transformation [73], [74] and [75] in this section.

3.3.1 Structure and Operation of the Proposed Single-Stage Voltage Regulator

The basic structure of the regulator is shown in Figure 3.7 (a). The four active switches (S_1 , S_2 , S_3 and S_4) are controlled by a PWM controller. When the controller gives a positive voltage (V_{ctrl}), the four switches work in the charging configuration as an active rectifier, which is shown in Figure 3.7 (b). In this charging configuration, when the input AC voltage is in its positive half period, switches S_1 and S_4 are closed so the current flows through them to the load, while in the negative half period the current flows through S_2 and S_3 . If the PWM controller gives a negative voltage, the regulator works in non-charging configuration as shown in Figure 3.7 (c), in which S_1 and S_4 are opened and S_2 and S_3 are closed. The closed S_2 and S_3 form a short circuit together with the receiver coil and the compensated capacitor, as discussed above. Unlike the work published in [43], which is a structure of a passive rectifier followed by a PWM controlled capacitor converter, it can be seen from Figure 3.7(a) that the circuits between the compensation capacitor and the load are only a power stage consisting of four active switches, which replaces the four diodes of a full-wave

rectifier, and controlled by a PWM controller. The merger of four active switches and PWM controller can be considered as a single-stage AC-DC converter

The control period T equals 16 times the periods of the charging periods [76]. We define the duty of the duration of the charging configuration as D and the duration of non-charging as $(1-D)$. The waveform of the current in the receiver coil is shown in Figure 3.7(d). It can be seen that in the charging configuration, the current I_2 slowly rises up to the charging current and becomes stable. In the non-charging configuration, I_2 slowly reduces to the value that can be calculated from (3) and becomes stable.

3.3.2 Equivalent Model WPT System with Single-Stage Voltage Regulator

In this section, the relationship between the duty-cycle of the charging configuration and the output voltage is analysed in detail. First, the WPT system is simplified to Figure 3.8 (a). The equivalent series resistance and capacitance of the coil are removed for the analysis. For simplicity, it is assumed that the switches in the power stage are ideally loss-less in Figure 3.8 (a). In the charging configuration, the switches works as an active full-wave rectifier so that the power directly goes to the load while in the non-charging configuration, the switches short the receiver, which means it is the same as Figure 3.7 (c). The equivalent circuit model of a mutual inductance is used to further simplify Figure 3.8 (a) to Figure 3.8(b). Since the current in the proposed WPT system is continuous, a phasor transformation method [73], which is useful to analyze the characteristics of a series resonant converter quantitatively, could be used to turn a rotational system into a stationary system. With this transformation, the inductors and capacitors in Figure 3.8 (b) are simplified to Figure 3.8 (c) [73]. And an equivalent V_{REG1} , R_{DC1} , lossless switch and C_{DC1} are shown in Figure 3.8 (c) to replace the rectifiers. As discussed before, the control switching frequency is much lower than the resonant charging frequency of the WPT system. Therefore in Figure 3.8 (d), the equivalent inductors and capacitor could be removed since the system could

Figure 3.7: Operation principle of the proposed regulator (a) overall structure (b) charging configuration (c) non-charging configuration (d) current of the receiving coil.

be considered as a DC equivalent circuit. A Norton transformation is applied to transfer Figure 3.8(d) to Figure 3.8 (e). It could be concluded that after transformation, the model is simplified to a switched DC current to voltage converter in accordance with [73]. The details of the theory of the phasor transformation of L , C and R are given in the appendix.

In the charging configuration, the switch is open so that the current I_{eq} flows directly into C_{DC1} and R_{L1} , and the charging current of the capacitor C_{DC1} is equal to $I_{eq} - I_{out1}$. In the non-charging configuration, the I_{eq} flows through the switch so

Figure 3.8: Model of the WPT system with the proposed single-stage voltage regulator.

that the discharge current of C_{DC1} equals I_{out1} . The D - V_{REG1} relationship can be derived as follows:

$$(I_{eq} - I_{out1}) \times D = I_{out1} \times (1 - D) \quad (3.8)$$

Since I_{out1} equals V_{REG1} over R_{L1} , we have

$$V_{REG} = V_{REG1} \times I_{eq} \times R_1 \times D \times G_v \quad (3.9)$$

where G_v is the voltage gain of a full-bridge rectifier, which equals $\pi^2/8$ [77] [73].

The linearity of the $D - V_{REG}$ is simulated in a Simulink model and shown in

Figure 3.9, which proves the proposed model.

Figure 3.9: Simulation result of V_{REG} vs. D .

3.4 Circuit Structure and Key Blocks

The proposed voltage regulator with all peripheral circuit blocks is displayed in Figure 3.10. To meet the basic requirements of the WPT system, the regulator is designed to support a 15 W output at 5 V for fast charging of cellphones or tablets. The resonant frequency of the WPT system is set to 125 kHz like many other Qi standard circuits. To control the regulator, the frequency of PWM signal is set to 7.8 kHz, which means the control period is 16 integral cycles per charging period.

Due to the limitation of the CMOS process, the power stage is not included in this chip design. Other sub-circuits like comparators, ramp generator, error amplifier and digital controllers are fabricated in a 0.35 μm AMS technology, which is in red dash lines in Figure 3.10 (a). When the PWM signal is set to “1” by the feedback loop, the regulator works in the charging configuration, as depicted in Figure 3.10 (b), which comprises a crossed-connected rectifier [49], [47], [78], [79] and [80]. The operation of this circuit for the positive half period can be explained as follows: when V_{HL} is zero, all four MOSFETs

Figure 3.10: Circuit implementation of the proposed voltage regulator (a) Complete blocks of the regulator ; (b) Circuits for charging configuration; (c) Circuits for non-charging configuration.

are turned off; subsequently, V_{HL} increases gradually as the AC input voltage increases; when $|V_{REG} - V_L|$ becomes larger than the threshold voltage of a PMOS-FET ($|V_{tp}|$), P_2 gradually turns on, connecting V_H to V_{REG} ; following that, the increasing AC voltage V_{HL} continuously pushes V_L until it drops below VSS . As a result, the CMP_2 turns on N_2 so that the charging current flows through the branch of P_2 and N_1 thereby charging towards V_{REG} ; after V_{HL} reaches the peak, it begins to decrease so V_L starts to rise; CMP_2 turns off the N_2 when V_L is above VSS ; V_L keeps rising and P_1 is turned off when $|V_{REG} - V_L|$ is smaller than $|V_{tp}|$. The operation of the circuit for the negative half period is similar to its positive counterpart.

When the PWM signal is set to "0" by the feedback loop, the regulator works in the non-charging configuration, as portrayed in Figure 3.10 (c). The two

PMOSFETs are disabled and two NMOSFETs are turned on to form a freewheeling loop. In this case, a short circuit is formed in the receiver end so that V_{REG} gradually decreases. According to the RLT (reflected load theory) [13], [81], [82] and [83], a short circuit reflects large equivalent impedance to the transmitter, which dramatically decreases the current of the transmitter. The limited transmitter current brings a small induced voltage to the receiver, thereby limiting the current and the power loss of the regulator as was analysed in previous sections. The DC current to the load is provided by the smoothing capacitor C_{DC} . Therefore, it is inevitable for the C_{DC} to be a large capacitor of 320 μF in this design and it is the only component which cannot be integrated onto the chip (Note: Even though the power transistors are also not integrated in our current prototype, they can be readily integrated if BCD process is available). Nonetheless, the main feature of the proposed single-stage AC-DC voltage regulator is that it significantly reduces the design complexity of the entire WPT system and at the same time enhances the power efficiency by doing away with DC-DC converter or communication modules. The main building blocks are discussed in detail in the following subsections.

3.4.1 Main Comparator Design

The main comparator (CMP_1 and CMP_2 in Figure 3.10 (a)) compares the voltage difference between the AC input voltage (V_H or V_L) and the DC output voltage (V_{REG}) to drive the NMOSFETs. As such, the comparator has to turn on/off the NMOSFETs with very little delay as (V_H or V_L) crosses (V_{REG}), or else the reverse current causes large extra power loss. Meanwhile, since the input dynamic range of the comparator has to cover the full voltage range of V_H - V_L , the comparator also has to handle the low voltage. Since the common-gate comparator is a fast and low-power-loss structure [47], [84] and [85], it is used in this design and described in Figure 3.11 .

Take CMP_2 for example, when the AC voltage is larger than the DC voltage,

Figure 3.11: Schematic of Main comparator (CMP_2).

which means V_H goes beyond V_{REG} , the gate of M_3 and M_1 increases to a certain level that turns off M_4 while turning on M_2 . As a result, M_7 is turned on while M_8 is turned off. Additional drivers are added to drive the large input capacitance of the power MOSFETs N_2 . When the V_H goes below V_{REG} , M_4 is turned on and M_2 is turned off, so the driver turns off N_2 . To guarantee low reverse leakage current, the turn-off operation of power MOSFETs is the main concern. So in this design, the size of M_4 is a bit larger than M_3 to slightly strengthen the pull-up action, avoiding the reverse leakage current. The power consumption of the comparator is about 1.25 mW.

The simulation results of the comparator and the AC input voltage are shown in Figure 3.12. From this figure, it can be seen that for the turn on action of the switch, the driving signal (V_{CMP}) has a several nano seconds delay. Since, the reverse leakage does not happen during switch on action, such a delay does not decrease the efficiency. For the turn off action, the MOSFET can be turned off much faster, the leakage happens at the overlap area of the green line and red line, which is small.

Figure 3.12: The leakage of the MOSFET.

3.4.2 Ramp Generator

The ramp generator of the PWM controller generates a sawtooth wave that determines the control frequency of the voltage regulator. In the proposed voltage regulator, the control frequency of 7.8 kHz is not high, hence the feedback loop has high tolerance against the jitter performance of the control signal [86], [87], [88] and [89]. In that case, a phase-locked-loop (PLL) does not to be included.

The circuit of the ramp generator is shown in Figure 3.13. In this circuit, the sizes of MP_1 and MP_2 are the same while the size of MN_2 is nine times that of MN_1 . So the rising slope of the Ramp is generated by charging a capacitor, C , with a constant current, I_{BIAS} . Once the Ramp rises above V_+ , the comparator, CM_1 , triggers the R-S latch and sets the S_{Dchg} to high voltage level. The S_{Dchg} then turns off MP_3 and turn on MN_3 , which discharges C with a constant current

Figure 3.13: Schematic of Ramp generator.

which is equal to nine times that of I_{BIAS} . Once the Ramp falls below V_- , CM_2 triggers the R-S latch to turn off MN_3 and turn on MP_3 . Switches MN_4 and MP_4 provide the current path when the current source transistors MP_2 and MN_2 are disconnected from MP_3 and MN_3 respectively so as to reduce the transient on-off glitches for the circuit [90].

The simulation results are shown in Figure 3.14. From this figure, the error voltage is very small, only 70 mV.

Figure 3.14: Simulation results of the ramp generator.

3.4.3 Start-up Circuit

The four internal diodes across the MOSFETs N_1 , N_2 , P_1 and P_2 work as a traditional passive full-bridge rectifier to regulate the AC input voltage to charge the V_{REG} up at the beginning. However, until the V_{REG} rises up to a certain value, the PWM control circuits cannot work accurately, which causes the failure of the

regulator. A startup circuit is used to solve this problem. As shown in Figure 3.15, initially V_{REG} is small, hence V_{fb2} is lower than the V_{th} . So the start-up transistor is off and V_{start} is clamped at V_{REG} by a pull-up resistor R_{pull} . V_{start} forces the regulator to work in the charging configuration. When V_{fb2} is higher than V_{th} , the NMOS is turned on to pull V_{start} down to zero. After that, the regulator goes into steady state and the PWM controller takes charge to control the power stage. The power consumption of the start-up circuit is about 2 mW.

Figure 3.15: Schematic of Startup Circuit.

3.4.4 Digital Controller

The digital controller is shown in Figure 3.16. The first function of the digital controller is to determine the state of the regulator. If the “ V_{start} ” is a high-level voltage, the OR gate sets “Charging” to “1” and “Non-charging” to “0”; in this case, the digital controller forces the power stage to work in the charging configuration. If the “ V_{start} ” is low, the digital controller let the PWM controller control the power stage.

The second function of the digital controller is to desynchronize the control signal and the driving signal of the main comparator (CMP_1 and CMP_2 in Figure 3.10). As shown in Figure 3.10 (a), if the control switches (SN_{11} , SN_{12} , SN_{21} and SN_{22}) and power transistors (N_1 and N_2) are turned on/off at the same time, there may be conflicting decisions. In this case, we do not want the control signals to switch at the same time as the main comparator, which turns on/off the power

Figure 3.16: Schematic of Digital Controller.

MOSFETs at the rising/falling edge of the V_H or V_L . In the proposed regulator, a latch generation is used to generate the control signal, which exclude the rising/falling edge of the V_H and V_L . In that case, the conflicting actions would be avoided.

3.5 Experimental Results

The single-stage WPT voltage regulator (the power stage is not included) is fabricated in a $0.35 \mu\text{m}$ CMOS process. The die photo is shown at the top of Figure 3.17 and the whole regulator that is fabricated on a PCB is shown at the bottom of Figure 3.17 (a). The power stage including four MOSFETs is highlighted. The measurement setup is shown in Figure 3.17 (b). With the proposed voltage regulator, a Class-D power amplifier operating at 125 kHz with a constant 19 V DC input voltage and a WPT coil pair. The parameters of the charging coils are shown in Table 3.3. The WPT system works with the charging distance for the range between 4 mm and 8 mm.

The charging frequency of the system is 125 kHz. The control switching

Figure 3.17: (a)Die photo and whole regulator on PCB.(b) Test bench

Table 3.3: Coil specifications

	Diameter (cm)	Inductance (μ H)	Q. at 125 kHz
Transmitting coil	8.36	11	144
Receiving coil	8.36	11	144

frequency was designed to be 7.8 kHz. The V_{REG} is designed to 5 V to meet the requirement of most portable devices.

The load transition and the charging distance transition are shown in Figure 3.18. The waveforms for the load transition from 15 W to 2 W and vice versa at a charging distance of 4 mm are shown in Figure 3.18 (a) and (b) respectively. The waveform at a charging distance of 8 mm is shown in Figure 3.18 (c).

With regard to the regulator's performance, power conversion efficiency (PCE) is the most important index. It is defined as the ratio of P_{OUT} to P_{IN} of the regulator, where the former is the DC power consumed at R_L while the latter is the AC power at the input to the regulator. The AC power is defined as the average value of product of the voltage V_{HL} and I_{ac} in Figure 3.10. The test is conducted with R_L changing from 1.6 ohms to 25 ohms at a charging distance of 4 mm. The result shown in Figure 3.19 indicates a 92 % PCE at about 10 W. At light load, the efficiency drops below 70 %. A pie chart in Figure 3.19(b) and (c) shows the simulation result of power loss including non-charging configuration loss, conduction loss, switching loss and power consumption of the controller circuit. It can be seen that at the full load condition, the conduction loss due to the on-resistance of the power device occupies most of the power loss while in the light

Figure 3.18: Oscilloscope waveform: (a) Measured waveform with a load transient of between 15 W and 2 W at 4 mm; (b) Measured waveform with a load transient of between 2 W and 15 W at 4 mm; (c) Measured waveform with a load of 15 W at 8 mm.

load the non-configuration loss is the main power loss.

The load regulation, line regulation and the charging distance regulation are shown in Figure 3.20.

It can be seen from these figures that the voltage variation of the output voltage within reasonable range. For load regulation, from the figure we can see that

Figure 3.19: Measured efficiency of the regulator.

Figure 3.20: (a) Load regulation from 25 (1W) to 1.6 (15 W) at 20V input and 4 mm distance; (b) Line regulation from 18 V to 24 V at 15 W at 4mm; (c) Distance regulation from 4 mm to 8mm at 20 V input and 15 W.

for the entire range of the output current, the voltage variation is within 2 %; for line regulation, it can be seen from an input range of 19 V-26 V, the voltage varies within 2%, which means it can provide enough voltage however when the input voltage is only 18 V, the regulator is not able to provide the desired voltage; for the transmission distance, if it is within 4 mm to 8 mm, the voltage regulation is within 2%. The voltage ripple is around 150 mV to 200 mV.

Table 3.4 summarizes the performance comparison with other WPT regulation system [10], [12] and [20]. This single-stage regulator has the largest 15 W output power for a wireless charger and it can regulate the output voltage without a communication circuit. Furthermore, the receiver efficiency is comparable with other works [34], [43] and [91].

The proposed architecture is verified in a 0.35 μm standard CMOS process together with the power transistors on the PCB. External power transistors are used as BCD process is not available in author's research lab. This regulator is

suitable for low cost portable applications requiring efficient robust WPT with a constant output voltage.

Table 3.4: Performance comparison with the proposed regulator

	[34]	[43]	[91]	[92]	[93]	This Work
Regulation	Wire -less Comm.	3R circuit	Wire -less Comm.	Two Stages	Impedance Matching	Single Stage Regulator
V_{in} (V)	19	N.A.	N.A.	6-12	5-12	19
V_{out} (V)	5	5	3.3	5	12	5
P_{out} Max (W)	5	6	0.2	5	24	15
Resonant f. (kHz)	125	6780	100 -150	144	200	125
Receiver Peak eff. (%)	94	86	Around 78	86	88	92
Process	BCD	BCD 0.35 μm	PCB	IC N.A.	PCB	CMOS 0.35 μm (off-chip transistors)

3.6 Conclusions

A novel AC-DC voltage regulator achieving low cost, high PCE has been presented. This WPT system has a power transfer capability of 15W and an independently regulating ability without a step-down DC-DC converter and communication models. A cross-connected rectifier structure significantly increases the efficiency of the regulator. The maximum output power reaches 15 W and the peak efficiency of the receiver board is 92 %. The proposed architecture is verified in a 0.35 μm standard CMOS process. This regulator is suitable for portable applications requiring efficient and intelligent WPT system.

Chapter 4

Design of A Full Active Rectifier with Adaptive Time Delay Control for 6.78 MHZ Wireless Power Transmission

4.1 Introduction

Figure 4.1: Wireless power transfer (WPT) system.

Although the kHz wireless power transfer are popular. In order to extend its power transfer capability to higher power range (tens of Watts) and distance (e.g. several centimetres), the wireless power transmission system operating at the range of megahertz has recently been developed. It is particularly suitable to be used in tablets or video game consoles. The system topology of a WPT system is redrawn here, as shown in Figure 4.1. As can be inferred from this figure, the overall efficiency can be expressed as:

$$\eta_{tot} = \eta_{PA} \times \eta_{coil} \times \eta_{rec} \times \eta_{DC} \quad (4.1)$$

In line with Equation (4.5), it is discernible that rectifiers with a high efficiency can improve the overall performance. In other words, the device will then have a shorter charging time. Also note that the efficiency of a rectifier has a great impact on the thermal emission of a portable device. This means that a low power-loss rectifier has lower chances of becoming overheated by excessive resistive power loss. Nevertheless, while it is crucial to have a highly efficient rectifier, designing such rectifier with high output current for megahertz applications is a challenging task. This is because a rectifier with high output current requires its transistor sizes to be inevitably large in order to reduce the conduction power losses. This will in turn result in a longer turn on/off time, which is incompatible with high frequency applications [46], [47] and [48].

Full-wave rectifiers are widely used for wireless power transfer applications, because of their high voltage conversion ratio. In general, full-wave rectifiers can be divided into three categories: 1) passive rectifiers with four diodes, 2) half-active rectifiers [94], [95], and 3) full-active rectifiers [46], [47], [48] and [41]. For 1) and 2), the forward voltage of the diode limits the power efficiency and voltage conversion ratio of the rectifier. For 3), four power transistors are used as four respective switches with small on-state resistance, hence having a much lower power loss than that of the diodes. However, the techniques used for milliamps applications in [46]- [48] are not suitable for high power applications

because of the large capacitances of the transistors. Delay locked loop (DLL) based on a voltage controlled delay circuit can be used to solve this problem [41]. To compensate for the delay of the gate driver, a replica delay unit has to be used in the DLL structure. Since the design is intended for high power application, the driver transistor itself can be very large; hence such replica unit would cause significant area overheads. Meanwhile, once there is a load change from high output current to low current, the active duration of the transistor would be shorter, which requires a fast response of the time delay response. However, it would take DLL tens of cycles to response, which would course the latent reverse current loss. In this chapter, we describe a ATD controller based on successive approximation register (SAR) algorithm to solve the above-mentioned problems. This solution is capable of extending the active duration of the rectifier while eliminating the reverse leakage.

The remaining of this chapter is organized as follows: 4.2 describes the architecture and operating principles of the proposed full-active rectifier; detailed circuit blocks such as the rising/falling edge detector, proximity detector, and the time-to-digital conversion circuit are described in 4.3; experimental results are shown in 4.4; whereas 4.5 draws the conclusions.

4.2 Proposed Rectifier with Adaptive Time Delay (ATD) Control

4.2.1 Structure and Operating principles of the Proposed Rectifier

The topology and operating principle of a basic cross-connected full active rectifier with the ATD controller is shown in Figure 4.2. The magnetic flux arising from the PA produces an AC voltage, V_{AC} (refer to Figure 4.1). C_{DC} denotes the DC smoothing capacitor and R_L is the load resistor. Two GaN transistors, N_3 and N_4 , with low voltage drop, connected in a cross-coupled configuration act as power switches turned on/off by V_{AC} . In addition to the switch pair, in order to

Figure 4.2: Topology and operation of the proposed active rectifier.

avoid the reverse leakage current from the load to the source, another two GaN transistors, N_1 and N_2 , are controlled by an ATD controller, which turns on N_1 and N_2 only when V_{AC} is larger than V_{REC} . The ATD controller monitors and controls the time difference between the gate drive signals (V_{G1} and V_{G2}) and the input voltages (V_H and V_L). A SAR algorithm is employed to maximize the conduction time of N_1 and N_2 to enhance the efficiency.

The operating principle of the rectifier during steady-state for the positive half cycle is illustrated in Figure 4.2. At the beginning of the positive half cycle (Figure 4.2 (a)), the input AC voltage, V_{AC} , has not yet reached the threshold voltage level of the GaN transistor, V_{th} . In other words, when $0 < V_{AC} < |V_{th}|$,

N_4 is “off” and N_1 is also turned off by the controller, there is no current going through N_1 and N_4 . When V_{AC} exceeds V_{th} , i.e. $V_{AC} > |V_{th}|$, N_4 will be turned on, causing V_L to be equal to VSS, whereas N_1 is still turned off, meaning that the current flows through N_4 and diode of N_1 . If V_{AC} continues to increase, since V_L is now connected to VSS by N_4 , the rising V_{AC} will be further pulled up to V_H . Once V_{AC} exceeds V_{REC} , the controller will begin to turn on N_1 . Consequently, due to the conduction of N_1 and N_4 , current begins to flow from V_{AC} to the load the smoothing capacitor C_{DC} . Since N_1 and N_4 are two GaN devices working in the deep triode region, their on-state resistance is small, which can help to minimize the conduction power loss, thus increasing the efficiency of the rectifier. V_{AC} begins to drop when it reaches its peak value. The controller sets a time delay to turn off N_2 exactly when $V_{AC} < V_{REC}$ as well as then there is no current going through N_1 . After once V_{AC} decreases to V_{th} , both N_1 and N_4 are turned off and the positive half cycle will not be terminated until V_{AC} drops to zero.

The conduction time of N_1 is controlled by the ATD controller, as shown in the Figure 4.2. If the conduction time is insufficient, N_1 is turned on at first such that the parallel diode can then conduct the current before V_{AC} drops to V_{REF} . The power losses in the diode deteriorate the performance of the rectifier. On the other hand, if the conduction time is too long, N_1 may not be fully turned off when $V_{AC} < V_{REC}$. As a result, a reverse leakage current will flow from V_{REC} to V_{AC} , thus causing a high power loss.

The operating principle of the negative half cycle is similar to that of the positive half-cycle except that in the negative half cycle the current flows through N_2 and N_3 .

4.2.2 Adaptive Time Delay Control Algorithm

Rectifiers reported in [46] and [47] use active switches controlled by high-speed comparators to eliminate the reverse leakage current. The basic idea is that the high-speed comparator can shut off the transistor once V_{AC} decreases to

V_{REC} . Since the propagation time of the comparator is very short, it can prevent the leakage current from flowing through the transistor completely. However, this method is suitable only for those applications with small output current, e.g. in the range tens to hundreds of milliamperes. For wireless power transmission in cell phones or tablets, the large output current needs be handled by large-size power transistors, the input capacitance of which is expected to be several times of those of low power application. As a result, when V_{AC} decreases to just below V_{REC} , the comparator cannot turn off the upper side transistors (N_1 and N_2) in a short given time. Taking the positive half cycle as an example, both N_1 and N_4 are still turned on when V_{AC} drops just below V_{REC} , in this case, a reverse current hence flows from C_{DC} back to the rectifier. Consequently, the power loss due to such a leakage current may be detrimental. In [48], an unbalanced-biased comparator design is utilized to compensate the turn off time of the transistor. However, the input AC voltage V_{AC} in a high power wireless power transmission application is more of a square wave than a sinusoidal wave. This implies that the unbalanced-biased comparators may cause instability issues.

Figure 4.3 shows the topology of the proposed rectifier with the ATD controller, and Figure 4.4 shows the timing diagram of the controller. In the proposed rectifier, the duty cycle of the gate drive signal $t_{d(n)}$ of N_1 and N_2 is adjusted to compensate the propagation time of the gate driver such that the transistor will be turned off exactly when V_{AC} decreases to V_{REC} , so that to remove the reverse current. A controller is used to detect the time difference between the transistor shut off time and the moment when V_{AC} decreases to V_{REC} , which is represented by Δt , as shown in Figure 4.4. A successive approximation feedback algorithm is utilized to control the duty cycle of the transistor. Now consider the positive half cycle ($V_{AC}=V_H-V_L > 0$) of the rectifier: As V_H increases above V_{REC} , the ATD controller commands S_1 to generate a setting signal to trigger the SR latch to turn on N_1 . In the n -th cycle, an adjustable time delay is set to generate a resetting signal, R_1 , to trigger the SR latch to turn off N_1 after an elapse of $t_{d(n)}$, which

Figure 4.3: Details of the adaptive time delay control.

equals t_{fix} , if $n=0$. After gate propagation, t_p , N_1 is then fully turned off. The ATD control module records the $t_{d(n)}$. In the next cycle, the control module sets the duty cycle of N_1 as

$$t_{d(n-1)} = t_{d(n)} + \frac{1}{2} \times \Delta t_{(n)} \quad (4.2)$$

This iteration will continue until Δt closely approaches zero after several cycles. When the iteration completes, the controller locks the time delay, and the rectifier then works in a steady state. It is worth noting that the additional propagation delay for the turn-on of N_1 is acceptable, because it does not generate any reverse leakage current.

The full active rectifier with ATD control has the following advantages. First, the SAR algorithm can gradually align the edge of the gate drive signal to the edge of the AC voltage, which maximizes the conduction time of the GaN transistors and in turn increases the efficiency. Second, the time delay lock function circuits guarantee that the total delay of the circuit does not push the edge of the gate

Figure 4.4: Timing of the adaptive time delay control.

drive signals after the AC voltage, which eliminates the leakage current of the rectifier. In addition, in this design, the controller automatically detects both the Δt and the total time delay thus further protecting the circuit from the reverse leakage.

4.2.3 Start-up Mechanism

At the very beginning, V_{REC} has not been powered up by wireless power from the transmitter; both the input AC voltage and the output DC voltage are close to 0 V, hence causing the controller to work in a unstable state, which is undesirable. It is therefore important to ensure that the proposed rectifier is able to startup in such a low voltage condition [96], [97], [98] and [99].

The timing diagram during start-up of the proposed active rectifier is shown in Figure 4.5. When the transmitter begins to transfer wireless power, the receiving coil generates an input AC voltage for the rectifier. However, the initial operating points of the V_{REC} are too low to activate both the control circuit and the GaN transistor. At the same time, the diodes form a passive full bridge diode rectifier

Figure 4.5: Timing diagram of the rectifier at start-up.

and start to charge up the load.

After the input voltage reaches the threshold voltage of the GaN devices, transistors N_3 and N_4 can then be turned on by V_{AC} . However, the V_{REC} at this stage is still too low to activate the controller. As a result, the rectifier can only work in the half-active mode. In such a mode, it is worth noting that the gate voltages of N_1 and N_2 should be set to “0”, since the controller would give out unpredictable output with a low power supply. In view of this, the start-up control circuit sets S_1 , S_2 , to “0” and R_1 , R_2 to “1”, making sure that the rectifier to work in the half-active mode, as shown in Figure4.5. The rectifier will continue to operate in the half-active mode until V_{REC} is charged up to a level, at which all remaining control circuits can work predictably.

As long as V_{REC} is at a proper level as discussed above, the rectifier will work

well in the full-active mode. In this mode, the adaptive controller starts to control the duty cycle of N_1 and N_2 as an active switch.

4.3 Circuit Structure and Key Blocks

4.3.1 Topology of the Proposed Rectifier

Figure 4.6: Topology of the proposed rectifier with detailed blocks.

The proposed active rectifiers along with all detailed blocks are illustrated in Figure 4.6. To meet the requirements of the high power devices, the rectifier is designed to support a 15 W output at 5 V so as to provide rapid charging for devices such as cell phones and tablets. The resonance frequency of the wireless power transmission system is targeted at 6.78 MHz for mid-distance applications.

The ATD controller marked in red dashed lines as shown in Figure 4.6 were fabricated based on a 0.18 μm high-voltage AMS process. The GaN power transistors (N_1 - N_4) are not incorporated in this chip design due to the integrability issues. As described earlier, the lower side GaN FETs (N_3 and N_4) are self-turned

on/off by the AC input voltage, whereas the upper side GaN FETs (N_1 and N_2) are controlled by the ATD controller. The ATD controller consists of an SAR feedback circuit, a fixed time delay unit and an adjustable time delay unit. For N_1 and N_2 , level shifters are used to boost the voltage level of the control signal. The gate drive circuit is used to provide a large gate charging/discharging current for GaN devices with associated large parasitic capacitance.

The SAR feedback circuit is composed of a rising edge detector, gate proximity detector, falling edge detector, time-to-digital converter (TDC), right shift, adder, registers, and digital lock circuit.

Since N_1 and N_2 should be controlled separately, typically two sets of controller and delay units need to be used. However, since only one upper side transistor (either N_1 or N_2) conducts current in one single half period, the power consumption of the controller are therefore limited.

As shown in Figure 4.6. The proximity detector generates a step signal once the gate voltage falls to the AC voltage (either V_H or V_L), while the falling edge detector generates a step signal once $V_{AC} < V_{REC}$. The TDC measures the time difference between these two step signals. The digital circuits including a right shift, an adder, and two registers realize the function described in Equation (4.2). The synchronous clock generator unit is used to switch the TDC and other digital circuits in different cycles. The digital result of delay time is stored in Register2, as shown in Figure 4.6. Once the Δt reach the set threshold value, the input of time delay line will be locked and the TDC keep monitoring the Δt in order to protect the circuit from potential reverse leakage current.

4.3.2 Rising Edge Detector and Falling Edge Detector

The rising and falling edge detector [100], [101], and [102] are used to detect the moment when V_{AC} crosses V_{REC} . For the rising edge detector, it is desired to have a fast response when V_{AC} surpasses V_{REC} to maximize the conduction time of N_1 or N_2 . For the falling edge detector, the response should be fast as well, to

ensure more accurate Δt detection.

Figure 4.7: Topology of the rising and falling edge detector.

The widely used common-gate comparator is chosen to construct the rising and falling edge detector. The schematic of the detector is shown in Figure 4.7. Compared to the common-source comparator, the common-gate structure will suffer less from the input offset voltage. The detector has three inputs: V_H or V_L , V_{REC} and VSS. For the rising edge detector, the output is set to 1 if the input AC voltage (V_H or V_L) signal is larger than V_{REC} -VSS. As for the falling edge detector, it provides the output signal only when the input AC voltage is smaller than V_{REC} . Consider now the positive half cycle: When the AC voltage (V_H or V_L) increases above the level of V_{REC} , the gate voltages of M_3 and M_1 will correspondingly increase to a certain level, at which M_4 is turned off and M_2 is turned on. As a result, M_7 is turned on and M_8 is turned off. This generates a high V_O .

When V_H or V_L decreases below V_{REC} , M_4 is turned on and M_2 is turned off. This generates a low V_O . An inverter is thereby added for the falling edge detector. Additional drivers are added to drive the input capacitance associated with the post-processing circuits. Transistors M_5 and M_6 specifically serve as bias-limiting transistors for this high voltage application. They guarantee a small dead-time between the states of on and off so as to eliminate the shoot-through current during the transient period.

In order to speed up the transient response of the detector, the size of M_2 is designed to be larger when compared to M_1 to improve the pull-down response

for the rising edge detector, and the same goes for transistor M_7 ; For the falling edge detector, on the contrary, larger sized transistor M_4 and M_8 are used to improve the response when V_{AC} decreases below V_{REC} . The power consumption of the rising/falling detector is about 3mW.

Figure 4.8: Simulation results of the rising edge detector.

The simulation results are shown in Figure 4.8 and Figure 4.9. It can be seen both detectors can have a fast response when the rising or falling edge come. In the simulation, the amplitude of both the input is determined by the consideration of the current and the on-resistance of the GaN devices. The frequency of the input is 6.78 MHz and the rising and falling time for the input pulse is around 12 ns due to the simulation results of the rectifier. The response time of the rising edge detector is around 0.3 ns, while the response time of the falling edge detector is 0.8 ns.

Figure 4.9: Simulation results of the falling edge detector.

4.3.3 Proximity Detector

The proximity detector is used to detect the very moment the gate driving signal falls to V_{REC} . At this time instance, the transistor is fully shut off. The proximity detector also provides the input signal for TDC. Figure 4.10 depicts the structure of the proposed detector, which consists of two unbalanced bias current (I_{S1} and I_{S2}), a pair of current amplifier with two current mirrors (M_9 , M_{10} and M_{13} , M_{14}), and a buffer used to drive the post-processing circuit. In the proposed detector, the branch consisting of M_{13} and M_{15} acts as a push-pull stage to accelerate the charging and discharging of the node capacitance.

Figure 4.10: Scheme of the gate proximity detector.

The proposed unbalanced-biased current generator is used to create a voltage bias. If the sizes of transistors M_{14} and M_{16} are the same as those of M_{10} and M_{12} , the currents in M_{15} and M_{13} will be balanced. In that case, when V_G drops to V_{REC} , node A will not be pulled up to V_{REC} . The actual circuit function is realized by setting $I_{S1} > I_{S2}$ so that the detector is able to pull the node A high as V_G approaches V_{REC} . The difference between I_{S1} and I_{S2} is controlled by setting the size of M_{21} to be larger than that of M_{22} . As a result, the pull-up capability of M_{21} is stronger than that of M_{22} , thus causing the pulling up current of the M_{15} larger the pulling down current of M_{13} when the V_G comes down below V_{REC} .

If V_G is much larger than V_{REC} , it is apparent that I_{S1} is then much smaller than I_{S2} . I_{S2} is mirrored by the current mirror which consists of M_{13} and M_{14} .

The difference between the two currents has a push-pull effect, which can pull node A down to ground. As V_G decreases to the vicinity of V_{REC} , due to the size difference of M_{12} and M_{16} , I_{S1} will be larger than I_{S2} . The current of $I_{S1}-I_{S2}$ thereby begins to charge the capacitance at node A, which pulls A to a higher voltage level to turn on the buffer and set the voltage of the output of the detector to be high. It should be pointed out that like the circuit in Figure 4.7, the limiting transistors are used in this circuit as well. However, since signals for GaN devices are over 5 V, so high voltage transistors (M_9-M_{16}) have to be used in this design. Usually, a high voltage transistor has a larger threshold voltage compared to normal transistors. As a result, the resistance of the limiting transistors, M_{21} and M_{22} , should be designed smaller so that they will not slow down the response of the detector. The power consumption of the rising/falling detector is about 3.5 mW.

The simulation results are shown in Figure 4.11. It can be seen that the proximity detector can have a fast response when the falling edge of the gate driving signal came. In the simulation, the amplitude of the gate drive signal was 10 V - 0 V at a frequency of 6.78 MHz. The response time of the rising edge detector is around 5 ns, while the response time of the proximity detector is 0.6 ns. This time delay is the reason that why the cross-connected rectifier can not be used in this design.

Figure 4.11: Simulation results of the proximity detector.

4.3.4 Digital Blocks

The digital blocks serve three main purposes: data conversion, calculation and recording. The data conversion circuit is carried out by TDC, whose structure is shown in Figure 4.12. In this design, a basic digital delay-line based TDC is constructed([103], [104], [105]and [106]). The “start” signal is delayed by a delay-line consisting of 64 delay elements. Those delayed signals are sampled in parallel, once the “stop” signal arrives. The output of this TDC is a sequence of thermometer codes. To help the post-processing circuit to deal with such data, a thermometer to binary decoder is designed to transfer the output to a 6 bit binary code.

Figure 4.12: Scheme of the TDC and the operation principle.

Following the time-to-digital conversion, the output is immediately sent to the right shift which performs the divided by 2 function. The output of the shift and the data in Register2, which records the digital outcome of time delay in the current cycle, are added together to form a new time delay for another coming cycle. Since the input of the digital circuit such as adders and other blocks should

be stable during the calculation phase-stage, two registers (Register 1 and Register 2) are thereby used for this design. A dual phase non-overlapping clock is constituted from V_H and V_L and it is synchronized by a clock generator. CLK1 and CLK2 ensure that the adder, the right shift and the updating of the time delay information to operate in a correct pace. It should be noted that the time interval between CLK2 and the new cycle should be sufficient for data processing and transfer. The timing diagram of the digital control is shown in Figure 4.13.

Figure 4.13: Timing diagram of the digital control.

A digital lock circuit is used to lock the time delay when the falling edge of the control signals V_{G1} (V_{G2}) gets very close to the V_H (V_L). This function is carried out by a multiplexer and two logic comparators. The lock operation occurs under two conditions. The first is that the adder reaches the upper limit of the setting value. At that time, the lock circuit pulls CLK1 and CLK2 to ground, to avoid data refresh. Second, the outcome of TDC reaches the bottom limit of the setting value, which to prevent the reverse leakage happen.

The output of the 6-bit digital delay signals control the delay circuit to adjust the time interval between S_1 (S_2) and R_1 (R_2). The delay circuit consists of a fixed delay line which is used to shorten the control time of the circuit and an

adjustable delay which converts the digital signal to the time delay. The structure of the delay circuit is shown in the Figure 4.14.

Figure 4.14: Schematic of the delay circuits.

4.4 Experimental Results

The proposed 15 W WPT active rectifier, not including the power stage, is fabricated with a 0.18 μm high voltage CMOS process. The photo of the proposed die is depicted in Figure 4.15. The entire receiver board with the proposed rectifier and communication module, a full bridge Class-D power amplifier operating at ZVS (zero-voltage-switching) and a planar PCB coil pair are also shown in Figure 4.15. The specifications of the circuit elements are given in Table 4.1. The wireless power transfer system works with a transmission distance of 5 cm.

Figure 4.15: Die photo and whole rectifier with GaN FETs on PCB.

The operating frequency of the system is 6.78 MHz and the output of the

rectifier is adjusted to 5 V by the communication module. Figure 4.16 shows the measurement results of the proposed rectifier with ATD control. Figure 4.16 (a) shows that the AC input voltage (V_H in Figure 4.6) and the corresponding gate signal (V_{G1} in Figure 4.6) from fixed delay to locked steady-state, for an output at 5V/3A. The zoom-in wave form in Figure 4.16 (b) shows that the gate signal successfully shut down the power transistor exactly when V_H fell below V_{REC} . Similar waveform has been shown in Figure 4.16 (c) and (d) for an output of 5V/0.15 A. It can be seen for different output current, the duration time of the conduction of the GaN devices are different. But for both case, they have been shut down in time.

Table 4.1: Testing components of the proposed 6.78-MHz Rectifier

Circuit Components	Value
Power Transistors	EPC2014
Transmitting coil	4.1 μ H/0.43 ohms @ 6.78 MHz
Receiving coil	4.1 μ H/0.43 ohms @ 6.78 MHz
Compensation Capacitor	100 pF/500V // 30 pF/500V // 5 pF/500V
Smoothing capacitor	22 μ F/50V

To define the performance of rectifiers, power conversion efficiency (PCE) is often the most important index. It is defined as the ratio of P_{out} to P_{in} , where the former is the DC power consumed at R_L while the latter is the AC power at the input to the rectifier. The AC power is defined as the average value of the product of V_{AC} and I_{AC} in Figure 4.6, given by

$$P_{in} = \int_0^t V_{AC} I_{AC} dt \quad (4.3)$$

while the output power is given by

$$P_{out} = V_{out} \times I_{out} \quad (4.4)$$

As a result, the efficiency of the rectifier is expressed as

Figure 4.16: (a) Waveform of the active rectifier at 5V/3A. (b) Zoom-in view of (a). (c) Waveform of the active rectifier at 5V/0.15 A. (d) Zoom-in view of (c).

$$\eta = \frac{P_{out}}{P_{in}} \quad (4.5)$$

The test is conducted with the output current changing from 3A to 0.15 A at a charging distance of 5cm. The measurement results shown in Figure 4.17 (a)

Figure 4.17: Measured efficiency of the rectifier.

indicates a 91.8% PCE at about 2 A. The efficiency of the rectifier with the fixed delay line and half-active rectifier are shown and compared in Figure 4.17 (a). At light load, the efficiency drops below 80%. A pie chart in Figure 4.17 (b) and (c) shows the simulation result of power loss including sub-circuit loss, switching loss, and conduction loss. The turn on delay loss is the power loss due to the delay of the rising detector. Since the power consumption of the ATD controller is a static loss source and switching loss is due to the capacitor and gate voltage, both of which are almost unchanged at different power level. Such power loss contributes close to 93% of the total power loss in light load mode and 25% of the total power loss in heavy load mode. This explains the reason of the efficiency degradation of the rectifier in the light load condition. A solution for this problem is to change the rectifier to half-active mode at the light load.

Table 4.2 summarizes the performance comparison with state-of-the-art rectifiers for WPT megahertz applications [49], [48] and [41]. This single-stage regulator provides the highest output power and comparable receiver efficiency is better than most of other works.

Table 4.2: Performance comparison with the proposed rectifier

	[107]	[108]	[49]	[48]	[41]	This Work
Structure	Half Active	Active diode	Active diode with RCC	Un-bal. comp.	DLL control	ATD control
V_{out} (V)	5.83	3	3.5	1	8.6	5
P_{out} Max (mW)	11	5	6.8	5	6000	15000
Resonant f. (MHz)	4	5	13.56	1.5	6.78	6.78
Peak eff.	N.A.	80 %	87 %	87.5 %	91.5 %	91.8 %
Process	CMOS 0.35 μ m	CMOS 0.35 μ m	CMOS 0.35 μ m	CMOS 0.35 μ m	BCD 0.35 μ m	CMOS 0.18 μ m (off-chip GaN)

4.5 Conclusions

In this chapter, a full-active rectifier with adaptive delay control (ATD) for 6.78 MHz wireless power transfer application is presented. This rectifier has a high power conversion efficiency because the proposed control algorithm not only eliminate the leakage current but also maximize the conduction of the active switch. This chip was implemented using 0.18 μ m AMS technology with four GaN transistors being used for the power stage of the rectifier. The maximum efficiency of the proposed rectifier is highly competitive at 90.6 % under output DC voltage condition of 5 V while providing 15 W output power, which is the highest among relevant works reported to-date.

Chapter 5

Analysis and Hybrid Control Method for 6.78 MHZ Wireless Power Transfer System with Varying Transmission Distance

5.1 Introduction

Although magnetic resonant WPT can improve the performance of the system for centimetres application, one problem is that as the transmission distance increases to a certain value, the power transfer capability decreases dramatically, known as frequency splitting [109], [51], [52] and [110]. As a result, the system is not able to provide enough output power to the load. To overcome the frequency splitting problem, a combined control method of frequency tuning and phase shift tuning is proposed in this chapter. This control method tunes the operating frequency of the magnetic resonant WPT system for short power transfer distance while tunes the phase shift for large power transfer distance. Meanwhile, the control is able to keep the overall efficiency at a high level in a wide range of power transfer distance and load condition. The rest of this chapter is organized as follows. Section 5.2 describes detailed analysis of the frequency splitting phe-

nomena based on a basic topology of a magnetic resonant WPT system and the equivalent circuit model. Equations of power transfer efficiency, output power, and frequency splitting have been derived. Based on the theoretical analysis, the control flow of the system and the parameter optimization of the system are described in Section 5.3. Experimental results and conclusions are presented in Sections 5.4 and 5.5, respectively.

5.2 Analysis of Magnetic Resonant Wireless Power Transfer System

5.2.1 System Overview

Figure 5.1: Wireless power transfer (WPT) system.

The WPT system is redrawn in Figure 5.1. The power is transferred wirelessly from the transmitting coil to the receiving coil. Both coils have been compensated to enhance the system efficiency and power transfer capability required for different applications. In the transmitter, the power amplifier (PA) transforms the input DC voltage to a high frequency AC voltage that drives the transmitting coil to generate AC magnetic fluxes. In the receiver, the coupled AC voltage is

rectified into a DC voltage (V_{REC}) by a rectifier. An additional DC-DC block and LDO is added to regulate the output voltage at a constant value. Wireless communication has always been utilized between the transmitter and the receiver to control the power flow of the system and to optimize the overall efficiency.

Figure 5.2: Simplified WPT circuit model: (a) Topology, (b) RLT model, (c) S-S circuit topology.

The WPT system can be simplified to an equivalent circuit model as shown in Figure 5.2.(a). L_1 and L_2 are the self-inductance of the transmitting and receiving coils, respectively. The interaction between the two coils induces an effect of mutual inductance, M . Each coil has an equivalent series resistance identified as R_1 and R_2 . The transmitter including the DC input, DC-AC converter and the compensation network is replaced by an AC voltage source V_{in} and an equivalent source series impedance Z_s , which is equal to $R_s + jX_s$. Since the source impedance is usually very negligibly small, it is omitted in the following analysis. As for the receiver, the compensation network, rectifier and DC-DC converter are replaced by an equivalent load impedance Z_o , which is equal to $R_L + jX_o$. The loop impedance of the receiver Z_2 is given by:

$$Z_2 = j(\omega L_2 + X_O) + R_2 + R_L = jX_2 + R_2 + R_L \quad (5.1)$$

Based on the simplified model and Kirchhoffs voltage law (KVL), the equations for the voltage -current relationships of the WPT system are derived as

$$\begin{cases} V_{in} = I_1(j\omega L_1 + Z_S + R_1) + I_2 j\omega M \\ 0 = I_2 Z_2 + I_1 j\omega M \end{cases} \quad (5.2)$$

where M is the mutual inductance

$$M = k\sqrt{L_1 L_2} \quad (5.3)$$

in Equation (5.3), k is the coupling coefficient between the two coils. According to Equation (5.1), (5.2) and the reflected load theory (RLT) ([14], [21], and [111]), the entire receiver can be replaced by a reflected impedance Z_r , and be integrated into the transmitter, as shown in Figure 5.2(b). Z_r is expressed as

$$Z_r = \frac{\omega^2 M^2}{Z_2} \quad (5.4)$$

Considering only the active power of the system, the efficiency is the ratio of the active power dissipated by Z_L to the active power developed by V_{in} . The overall efficiency can be expressed as the product of the efficiency of the transmitter and the receiver, which is given by,

$$\begin{cases} \eta_1 = \frac{\Re(Z_r)}{\Re(Z_r) + R_1 + R_S} \\ \eta_2 = \frac{R_L}{R_L + R_2} \end{cases} \quad (5.5)$$

According to Equation (5.5), the system efficiency is derived as

$$\eta = \eta_1 \eta_2 = \frac{\omega^2 M^2 R_L}{\omega^2 M^2 (R_L + R_2) + (R_1 + R_S) [(R_L + R_2)^2 + X_2^2]} \quad (5.6)$$

From Equation (5.6), it can be seen that the efficiency will be improved if the receiver is compensated to be resistive, thereby nullifying X_2 . For this reason,

series compensation technique is used for the receiver since a series resonant LC network can achieve a zero X_2 . The transmitter is also compensated in series to give a unity power factor (UPF), thus enhancing the power transfer capability. An equivalent circuit representation of the series-series (S-S) model is shown in the Figure 5.2 (c). Meanwhile, the R_s has been removed for simplicity, since it is substantially small when compared to R_1 . The analyses presented in the subsequent sections of this thesis are based on this model thereafter.

5.2.2 Maximum Efficiency of a Series-Series (S-S) Compensation

A lot of research has analysed the maximum power transfer efficiency ([112], [113], [114], [115], [116], [117], [118], [119], [120], [91], and [121]). In an S-S compensation WPT system as shown in Figure 5.2 (c), assuming that the power losses of inverter and ferrite shielding is omitted, the KVL state equations can be written as

$$\begin{cases} V_{in} = I_1(R_1 + j\omega L_1 + \frac{1}{j\omega C_1}) + I_2 j\omega M \\ 0 = I_2(R_2 + R_L + j\omega L_2 + \frac{1}{j\omega C_2}) + I_1 j\omega M \end{cases} \quad (5.7)$$

The Z_2 , Z_r can be derived as

$$Z_2 = R_2 + R_L + j\left(\omega L_2 - \frac{1}{\omega C_2}\right) \quad (5.8)$$

$$Z_r = \frac{\omega^2 M^2}{R_2 + R_L + j\left(\omega L_2 - \frac{1}{\omega C_2}\right)} \quad (5.9)$$

Then the efficiency of the system is computed as

$$\eta = \frac{\omega^2 M^2 R_L}{\omega^2 M^2 (R_L + R_2) + R_1 (R_L + R_2)^2} \quad (5.10)$$

if the system works in the resonant condition which means

$$\omega = \omega_0 = \frac{1}{\sqrt{L_1 C_1}} = \frac{1}{\sqrt{L_2 C_2}} \quad (5.11)$$

where ω is operating frequency and ω_0 is the resonant frequency. The equation can be further derived that

$$\begin{aligned} \eta &= \frac{\omega^2 M^2}{\omega^2 M^2 + 2R_1 R_2 + R_1 R_L + \frac{\omega^2 M^2 R_2 + R_1 R_2^2}{R_L}} \\ &\geq \frac{\omega^2 M^2}{\omega^2 M^2 + 2R_1 R_2 + 2\sqrt{R_1 R_L \times \frac{\omega^2 M^2 R_2 + R_1 R_2^2}{R_L}}} \\ &= \frac{\omega^2 M^2}{\omega^2 M^2 + 2R_1 R_2 + 2\sqrt{R_1 (\omega^2 M^2 R_2 + R_1 R_2^2)}} \end{aligned} \quad (5.12)$$

when

$$R_L = R_{OPT} = \sqrt{\frac{\omega^2 M^2 R_2 + R_1 R_2^2}{R_1}} = \sqrt{\frac{\omega^2 k^2 L_1 L_2 R_2 + R_1 R_2^2}{R_1}} \quad (5.13)$$

Equations (5.12) and (5.13) imply that the optimized load value is related to the coupling coefficient which is proportional to the distance between the transmitter and receiver. Figure 5.3 shows the system efficiency over a wide range of load characteristics. It can be concluded that, for strong coupling ($k=0.5$), the efficiency is not much affected by the load conditions, while for loose coupling ($k=0.07$), the efficiency drops drastically as the load deviates greatly from the optimum value. Thus, efforts should be elicited in optimizing the load for the case of loose coupling.

5.2.3 Frequency Splitting Analysis

Another important design specification of the magnetic resonant WPT system is the output power. The output power is defined as the level of real power that can be transferred to the load in the WPT system. In extreme strongly coupled conditions ($k>0.9$), the output power is a function of the ratio of the number of turns of the transmitting coil to the number of turns of the receiving coil. In a

Figure 5.3: Efficiency with load variation.

loosely coupled S-S WPT system, the output power is affected by the coupling coefficient, the value of the capacitor and the operating frequency([109], [51], [52], [110], [122] and [123]). For this reason, a detailed analysis on the maximum output power conditions and frequency splitting is carried out for a varying range of coupling coefficients.

From Figure 5.2, the output power of the system can be calculated as

$$P_{out} = I_2^2 R_L = \left| \frac{j\omega M I_1}{Z_2} \right|^2 R_L \quad (5.14)$$

where I_1 is the transmitting coil current, which can be calculated by the RLT based on the Figure 5.2 (b) as

$$I_1 = \frac{V_{in}}{j\left(\omega L_1 - \frac{1}{\omega C_1}\right) + R_1 + \frac{\omega^2 M^2}{j\left(\omega_2 L_2 - \frac{1}{\omega_2 C_2}\right) + R_2 + R_L}} \quad (5.15)$$

For simplification, the efficiency of the overall system can be expanded as

$$P_{out} = \frac{V_{in}^2 C_1^2 C_2^2 M^2 R_L}{F} \quad (5.16)$$

where

$$F = A\omega^2 + B + C\frac{1}{\omega^2} + D\frac{1}{\omega^4} + \frac{1}{\omega^6} \quad (5.17)$$

which is expressed in terms of parameters A, B, C, and D:

$$A = M^2 C_1^2 C_2^2 - 2M^2 C_1^2 C_2^2 L_1 L_2 + C_1^2 C_2^2 L_1^2 L_2^2 \quad (5.18)$$

$$\begin{aligned} B = & 2M^2 C_1^2 C_2^2 R_1 R_2 + 2M^2 C_1^2 C_2^2 R_1 R_L + C_1^2 C_2^2 L_1^2 R_2^2 + 2C_1^2 C_2^2 L_1^2 R_2 R_L + C_1^2 C_2^2 L_2^2 R_L^2 \\ & + C_1^2 C_2^2 L_2^2 R_1^2 + 2M^2 C_1^2 C_2 L_1 + 2M^2 C_1 C_2^2 L_2 - 2C_1^2 C_2 L_1^2 L_2 - 2C_1 C_2^2 L_1 L_2^2 \\ & \doteq 2M^2 C_1^2 C_2 L_1 + 2M^2 C_1 C_2^2 L_2 - 2C_1^2 C_2 L_1^2 L_2 - 2C_1 C_2^2 L_1 L_2^2 \end{aligned} \quad (5.19)$$

$$\begin{aligned} C = & C_1^2 C_2^2 R_1^2 R_2^2 + 2C_1^2 C_2^2 R_1^2 R_2 R_L + C_1^2 C_2^2 R_1^2 R_L^2 - 2C_1^2 C_2 L_2 R_1^2 - 2C_1 C_2^2 L_1 R_2^2 \\ & - 4C_1 C_2^2 L_1 R_2 R_L - 2C_1 C_2^2 L_2 R_L^2 - 2M^2 C_1 C_2 + C_1^2 L_1^2 + 4C_1 C_2 L_1 L_2 + C_2^2 L_2^2 \\ & \doteq -2M^2 C_1 C_2 + C_1^2 L_1^2 + 4C_1 C_2 L_1 L_2 + C_2^2 L_2^2 \end{aligned} \quad (5.20)$$

$$D = C_1^2 R_1^2 + C_2^2 R_2^2 + 2C_2^2 R_2 R_L + C_2^2 R_L^2 - 2C_1 L_1 - 2C_2 L_2 \doteq -2C_1 L_2 - 2C_2 L_2 \quad (5.21)$$

Replacing the L_1 , L_2 , C_1 and C_2 with Equations (5.3) and (5.11), it can be shown that

$$F = \left(\frac{k^2}{\omega_o^8} - \frac{2k^2}{\omega_o^8} + \frac{1}{\omega_o^8} \right) \omega^2 + \left(\frac{6}{\omega_o^2} - \frac{2k^2}{\omega_o^2} \right) \frac{1}{\omega^2} - \frac{4}{\omega_o^2} \frac{1}{\omega^4} + \frac{1}{\omega^6} \quad (5.22)$$

By differentiating Equation 5.22 with respect to ω^2 and calculating the roots of the first-order equation, given by

$$\frac{\partial F}{\partial \omega^2} = 0 \quad (5.23)$$

it can be found that the valid roots of Equation (5.23) are:

$$\omega_1^2 = \frac{\omega_0^2}{1+k} \quad (5.24)$$

$$\omega_2^2 = \frac{\omega_0^2}{1-k} \quad (5.25)$$

$$\omega_3^2 = \frac{\left(\sqrt{4-3k^2}-1\right)\omega_0^2}{1-k^2} \quad (5.26)$$

Substituting Equations (5.24), (5.25) and (5.26) into the second derivative of Equation (5.22), it can be determined whether ω_1 , ω_2 , and ω_3 are local maximum or minimum:

$$\frac{\partial^2 F}{\partial(\omega^2)^2} \Big|_{(\omega^2=\omega_1^2)} = \frac{8(1+k)^2 k^2}{\omega_0^{10}} > 0 \quad (5.27)$$

$$\frac{\partial^2 F}{\partial(\omega^2)^2} \Big|_{(\omega^2=\omega_2^2)} = \frac{8(1-k)^2 k^2}{\omega_0^2} > 0 \quad (5.28)$$

$$\frac{\partial^2 F}{\partial(\omega^2)^2} \Big|_{(\omega^2=\omega_3^2)} = \frac{8(1-k)^3(1+k)^3 K(K-2)}{3(K-1)^3 \omega_0^{10}} < 0 \quad (5.29)$$

where

$$K = \sqrt{4-3k^2} \quad (5.30)$$

It is deduced that F has local minimum value at ω_1 and ω_2 , and local maximum value at ω_3 . From Equations (5.24) to (5.26) it is clear that when the coupling is strong, ω_1 , ω_2 and ω_3 are far away from each other. This is known as frequency splitting phenomena. The splitting coefficient is defined as

$$d_\omega = \left(\frac{1}{\sqrt{1-k}} - \frac{1}{\sqrt{1+k}} \right) \omega_0 \quad (5.31)$$

Equation (5.31) determines the splitting frequency of the two peaks of the output power. Figure 5.4 depicts a 2-D plot of calculated data of output power versus

operating frequency, which demonstrates the frequency splitting occurrence. The picture includes five curves for different coupling coefficients. It can be seen that when the coupling coefficient is below 0.07, ω_1 , ω_2 and ω_3 converge. To verify the analysis, the comparison of calculations results and experimental results ($k=0.5$, $k=0.07$) are shown in Figure 5.5. The parameters used for Figure 5.4 are shown in Table 5.1.

Figure 5.4: Frequency splitting effect (Calculation results).

Table 5.1: Parameters in Figure 5.4

Symbol	Parameters	Value
f_0	Resonance frequency	6.78 MHz
L_1 and L_2	Coil inductance	$4 \mu\text{H}$
C_1 and C_2	Compensation Capacitor	138 pF
R_1 and R_2	ESR of coils @ f_0	0.43Ω
R_L	Load resistance	10Ω

Another interesting phenomenon is that, at frequencies ω_1 , ω_2 and ω_3 , the phase angles of the input impedance Z_{in} (as seen from the AC source), which is given by

$$Z_{in} = Z_r + R_1 + j \left(\omega L_1 - \frac{1}{\omega C_1} \right) \quad (5.32)$$

Figure 5.5: Frequency splitting effect (Calculation and experimental results).

are zero. To approve that, Equation (5.32) can be expanded as

$$Z_{in} = \frac{\omega^2 M^2 (R_2 + R_L)}{(R_2 + R_L)^2 + \left(\omega L_2 - \frac{1}{\omega C_2}\right)^2} + R_1 + j \left(\omega L_1 - \frac{1}{\omega C_1} - \frac{\omega^2 M^2 \left(\omega L_1 - \frac{1}{\omega C_1}\right)}{(R_2 + R_L)^2 + \left(\omega L_1 - \frac{1}{\omega C_1}\right)^2} \right) \quad (5.33)$$

Using Equation (5.11), and assuming that the two coils are identical, the imaginary part of Z_{in} is normalized and simplified as

$$\Im(Z_{in}) = \frac{(\omega_0 + \omega)(\omega_0 - \omega)(k\omega^2 - \omega^2 + \omega_0^2)(k\omega^2 + \omega^2 - \omega_0^2)}{\omega_0^2 C_1 \omega E} \quad (5.34)$$

where

$$E = \omega^2 C_1^2 R_2^2 \omega_0^4 + 2\omega^2 C_1^2 R_2^2 R_L^2 \omega_0^4 + \omega^2 C_1^2 R_L^2 \omega_0^4 + \omega^4 + \omega_0^4 - 2\omega^2 \omega_0^2 \quad (5.35)$$

From Equation (5.34), it is clear that the phase angle is zero at ω_1 , ω_2 and ω_0 .

5.3 Voltage Control Method and Efficiency Optimization

5.3.1 Voltage Control Method Description

Effective control of output voltage is one of the critical criteria for achieving optimal wireless power transfer for portable consumer electronics. The regulation approach can be placed in either the transmitter or the receiver, depending on the requirements of different applications, like the area of the device, output voltage tolerance and response time ([124], [125], [126], [127] and [128],).

According to the previous analysis, the system always has peak efficiency at resonant frequency. So it is important to optimize the system efficiency as the output voltage requirement is met ([129], [130], [131], [132], [133], [121] and [134]). For loose coupling condition, the output power is enough to provide the standard voltage. For strong coupling condition, unfortunately, the output voltage is not enough, as shown in Figure 5.4 ($k=0.3-0.5$).

According to the circuit analysis presented in section 5.2, the system will always operate at its peak efficiency at the resonant frequency. For loose coupling conditions, the output power is adequate to provide the required standard voltage. For strong coupling conditions, unfortunately, the output power is far from being practical, as shown in Figure 5.4 for k varies from 0.3 to 0.5.

This implies that it is possible to use phase shift tuning to control the power flow for weakly coupled cases whereas the operating frequency must be tuned to meet the output power for strongly coupled cases. Since the equivalent series resistance (ESR) of the coil is proportional to the frequency, the operating frequency will be tuned towards ω_1 instead of ω_2 .

The detailed voltage control flow chart of the proposed control methodology is shown in Figure 5.6. Its objective is to guarantee that the power transferred to the load is adequate to provide the standard voltage for general-purpose applications.

To start with, an initial AC input voltage V_{in} is applied to the transmitter. The

Figure 5.6: System control flow.

frequency of input voltage equals the resonant frequency (6.78 MHz) and the phase shift of the amplifier is fixed at 10%.

After initiating the power transfer process, the receiver begins to transfer the voltage error packets back to the transmitter and the micro-controller begins the phase shift tuning process according to the error packets. If the output voltage is able to meet the minimum value, the control is considered successful. Otherwise, if the output voltage is still lower than the minimum required output voltage when the phase shift is tuned to θ_{max} , the phase shift control is considered not meeting the requirement.

If the phase shift test fails in this way, frequency tuning begins. The controller tunes the frequency towards ω_1 . We can in fact construe from Figure 5.4, that it is also possible to tune the frequency towards ω_2 . However, it is imperative to take note that the ESR of the coil increases with increased frequency, hence a lower frequency is preferred([135], [136], [137], [138] and [139]). Once the output voltage is larger than the minimum value, the frequency tuning process ends. If the load or the transfer distances change during power transfer, the process continues according to one of following scenarios:

Scenario 1: This is pursued provided the operating frequency is different from the resonant frequency. If the output voltage is lower than the required voltage (V_{min}), the frequency will be set lower to provide sufficient voltage. If the V_{min}

requirement is still not met despite the frequency adjusted lower than ω_1 , the control is considered as failed. Conversely, if V_{out} is higher than V_{max} , the frequency will be tuned higher. If the frequency is tuned to resonant frequency but V_{out} is still higher than V_{max} , the phase is tuned lower to match the requirement.

Scenario 2: This happens when the operating frequency equals the resonant frequency. If the output voltage is larger than V_{max} , the phase is tuned lower to meet the requirement. If the output voltage is lower than V_{min} , the phase is tuned higher. If, when the phase is tuned to θ_{max} , it still cannot provide V_{min} , the frequency control starts as described in Scenario 1

It is worth noting that in real-life application, the signal of output voltage would be kept being sent back to the transmitter with a communication ZigBee module. Unlike the single-stage voltage regulator mentioned in the Chapter 3, the voltage control proposed in this chapter is used to extend the power transmission distance over a wide range. Such applications target to high end products so the requirement of the cost is not as stringent.

5.3.2 Parameter Design for Power Transfer Efficiency

From Figure 5.4, it can be seen that the output power would be very low in short distance ($k=0.5$). So in this case, the frequency is supposed to be tuned lower to meet the power requirement. However, due to the analysis mentioned before, the efficiency would decrease when the system works under non-resonant condition. The efficiency deterioration factor, $Dete$, is defined as:

$$Dete = \frac{1}{\eta_{(\omega)}} \quad (5.36)$$

The deterioration factor at 1 and 2 can be calculated as

$$\begin{cases} Dete_{(\omega1)} = \frac{1}{\eta_{(\omega1)}} - \frac{1}{\eta_{(\omega0)}} = \frac{R}{R_L} \\ Dete_{(\omega2)} = \frac{1}{\eta_{(\omega2)}} - \frac{1}{\eta_{(\omega0)}} = \frac{R}{R_L} \end{cases} \quad (5.37)$$

It can be seen that $Dete$ is related to the ratio of ESR of the coil to load resistance. The worst-case scenario happens when the system works under full load condition, at which the efficiency would drop around 4%. In practical applications, this drop could be less because: 1) frequency tuning towards ω_1 means a smaller ESR due to lower frequency; 2) the operating frequency does not have to be equal to ω_1 , but low enough to provide adequate required voltage. Besides, frequency splitting happens when k is large, which further increases the efficiency due to Equation 5.6. This approach is adopted in the proposed control method presented in the subsequent section. The theoretical and experimental data of efficiency vs. f is given in Figure 5.5. This analysis validates that the system being able to work with high efficiency and output power in both strong and loose coupling.

Figure 5.7: Frequency splitting effect (Efficiency drop at ω_1 and ω_2 : calculation and testing results.).

The efficiency of a wireless power transfer system, which is not operating in at the resonant frequency, is given by

$$\eta = \eta_1 \eta_2 = \frac{R_L}{R_1 \frac{(R_2 + R_L)^2 + (\omega_0 L_2 \Delta)^2}{\omega^2 M^2} + R_2 + R_L} \quad (5.38)$$

where

$$\Delta = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \quad (5.39)$$

From Equations (5.38) and (5.39), it can be seen that the efficiency decreases as the operating frequency (ω) deviates from the resonant frequency (ω_0). To replace the ω_0 of Equation (5.38) with Equation (5.11), the efficiency at ω_1 is given by

$$\eta_{\omega_1} = \frac{R_L(\omega_0^2 L_2^2 k^2)}{(1+k)R_1 \left[(R_2 + R_L)^2 + (\omega_0 L_2)^2 \left(\frac{1}{\sqrt{1+k}} - \sqrt{1+k} \right)^2 \right] (\omega_0^2 L_2^2 k^2) + (R_2 + R_L)(\omega_0^2 L_2^2 k^2)} \quad (5.40)$$

It can be seen from the derived equation that the efficiency at ω_1 is related to the coupling coefficient and the coil inductance. Figure 5.8 shows the change in efficiency with varying coupling coefficients for different coil inductances.

Figure 5.8: Efficiency at ω_1 for different coupling value.

Figure 5.8. shows that when the coupling is strong, the efficiency at ω_1 is high regardless of the value of the coil inductance. However, when the coupling is loose, the efficiency of the system with low inductance value is not high enough. This therefore suggests that the value of the inductance of 4 μH is used in this thesis.

For a 6.78 MHz wireless power transfer system, a planar coil is always utilized because of its compact structure. The equation for the inductance of a coil is given in [140], [141], [142], [143] and [144] as

$$L = \frac{1}{2}N^2\mu d_{avg}C_1 \left(\ln \frac{C_2}{F} + C_3F + C_4F^2 \right) \quad (5.41)$$

where N is the number of turns, μ is the permeability of the material, d_{avg} is the average diameter and C_1 - C_4 are the parameters related to the shape of the coil. According to [140], C_1 - C_4 of a coil that is of a square shape is larger than those of a circular coil. In view of that, a square shaped coil is chosen in this design.

Another important parameter of the coil is its resistance, especially when the coil works at high frequency. The AC resistance of the coil is mainly determined by two factors: the skin effect and the proximity effect. To mitigate skin effect losses, the thickness of the coil should be designed to be larger than the skin depth. To alleviate the proximity effect, the space between each coil turn should be as least at least one diameter apart.

5.3.3 Design of a Zero Voltage Switching (ZVS) CLASS-D Amplifier

Another challenging issue with the 6.78 MHz wireless power transfer system is the significant power loss of the power amplifier at the transmitter end, due to the high operating frequency of the system. To reduce the switching losses, a full-bridge zero voltage switching (ZVS) class D amplifier is adopted in this design ([145], [146], [147], [148], [149], [150], [151] and [152]).

Figure 5.9 shows a basic topology of a typical full-bridge ZVS class D amplifier and its corresponding operating waveforms [147]. The GaN transistors (A_1 , A_2 , B_1 and B_2) form the power stage of the amplifier; the inductance L_{zvs} is used to store the energy for ZVS operation; L_1 and C_1 is the resonator including the coil and the compensation capacitor; Z_r , as described before, is the equivalent reflected impedance of the entire receiver including the receiving coil, compensation capacitor, rectifier, DC-DC converter and the load.

Figure 5.9: ZVS Class-D Amplifier.

The operating mechanism of the amplifier is depicted in Figure 5.10. The overall operation of the ZVS class-D amplifier period can be segmented into four phases, as presented below.

In Phase I, the two diagonally positioned GaN transistors, A_1 and B_2 , are conducting. The current of the coil L_1 flows through the two transistors to transfer power to the load. The V_{PN} is positive hence causing the current in L_{zvs} to increase. The current stored in the L_{zvs} will ultimately be used to realize zero voltage switching.

During Phase II, transistor B_2 shuts down. At this point, the current to load should be close to zero if it is well compensated. The current in L_{zvs} begins to charge the capacitance of B_2 while discharging B_1 . This phase should end once the voltage across B_1 is discharged to zero. Otherwise, the current will flow through the diode of B_1 causing unnecessary power loss.

In Phase III, transistor B_1 is turned on and A_1 is turned off after that. Since the voltage across B_1 is during Phase II was at zero level, the switching operation is considered as ZVS. I_1 is still close to zero and I_{zvs} begins to discharge the capacitance of A_2 until the voltage across A_2 is zero.

In Phase IV, A_2 turns on. Since the voltage across it is zero in Phase III, it is also considered that ZVS is in action. The current of the coil L_1 goes through A_2

Figure 5.10: Operation of the ZVS.

and B1 to transfer power to the load. The V_{PN} is negative triggering the current in L_{zvs} to decrease. The current stored in L_{zvs} will be used to realize the ZVS action of A_1 and B_2 in the next switching action.

The waveform of the ZVS class-D amplifier is shown in Figure 5.11. It can be seen that in Phase I, the current of the L_{zvs} is built by VDD, given by

$$I_{ZVS} = \frac{1}{2L_{ZVS}} \int_{t_{ZVS}}^{\frac{1}{2}T - t_{ZVS}} VDD \cdot dt = \frac{VDD(T - 4t_{ZVS})}{4L_{ZVS}} \quad (5.42)$$

According to the analysis above, the I_{zvs} should discharge the C_{OSSQ} , which is the equivalent output capacitance of the GaN devices, in the time duration t_{zvs} . Since t_{zvs} is short so that I_{zvs} can be considered as constant during t_{zvs} . It can be derived that:

$$L_{zvs}t_{zvs} = C_{OSSQ}VDD \quad (5.43)$$

Figure 5.11: Voltage-Current waveform of the ZVS class-D amplifier.

According to Equation 5.43, L_{ZVS} is thus expressed as

$$L_{ZVS} = \frac{t_{ZVS}(T - 4t_{ZVS})}{4C_{OSSQ}} \quad (5.44)$$

5.4 Experimental Results

The experimental test bench (Figure 5.12) was built based on the parameters tabulated in Table 5.2. Experimentations were carried out to verify and evaluate the proposed voltage regulation and efficiency optimization methodologies of the magnetic resonant wireless power transfer system. A full-bridge class-D ZVS power amplifier is incorporated at the transmitting end to generate a 6.78 MHz AC voltage to drive the transmitting coil. The DC input voltage for the amplifier is fixed at 12 V, which is the required minimum voltage for long-distance WPT. As described above, both the operating frequency and phase shift are tuned by a MCU (MICROCHIP DSPIC33FJ16GS402-I) with a high speed PWM function.

Testing Case I: Figure 5.13 shows the output waveform of the class-D amplifier in the steady state when the system is working at a 5 cm distance (loose coupling) and 10 ohm load. It can be seen that the output pulse width is close to 100% at resonant frequency and that the output voltage across the load is around

Figure 5.12: Test bench for the proposed 6.78-MHz WPT system.

Table 5.2: Testing Components of the proposed 6.78-MHz WPT system

Circuit Components	Value
Power Transistors	EPC2014C
Transmitting coil	$4.1 \mu\text{H}/0.43 \text{ Ohms @ } 6.78 \text{ MHz}$
Receiving coil	$4.1 \mu\text{H}/0.43 \text{ Ohms @ } 6.78 \text{ MHz}$
Compensation Capacitor	$100 \text{ pF}/500\text{V} // 30 \text{ pF}/500\text{V} // 6.9 \text{ pF}/500\text{V}$
Smoothing capacitor	$22 \mu\text{F}/50\text{V}$
Micro-controller	MICROCHIPDSPIC33FJ16GS402-I
Communication Module.	XBee by Digi

10 V. It can be seen that the frequency is close to 6.78 MHz (period = 148 ns), which implies that the system works under phase shift tuning mode at the resonant frequency.

Testing Case II: If the distance and the DC input voltage are kept the constant at 5 cm and 12 V, respectively, and the load is 50 ohms, the controller will change the duty cycle of the output voltage of the amplifier according to the feedback signals as what has been described as the control flow in Figure 5.6. At the beginning, the duty cycle is about 70 ns as can be seen from Figure 5.13. The controller then progressively controls the duty cycle to approximately 20 ns as shown in Figure 5.14. Meanwhile, it is corroborated that the output DC voltage is successfully regulated to maintain at around 10 V. The voltage regulation time from Testing Case I to Testing Case II is around 300 ms.

Testing Case III: If the load is kept at 10 Ohms but the power transfer distance shortened to 1 cm. The system will then work in strong coupling mode and

Figure 5.13: Oscilloscope waveform of the output voltage cross the load ($R_L = 10 \Omega$) and the power amplifier output voltage at 5 cm.

Figure 5.14: Oscilloscope waveform of the output voltage cross the load ($R_L = 50 \Omega$) and the power amplifier output voltage at 5 cm.

frequency splitting phenomenon will adversely impact the power transfer capability. In line with the control flow chart depicted in Figure 5.6, the controller tunes the operating frequency lower once it detects that it is impossible for the system working under resonant frequency. As shown in Figure 5.15, the operating frequency is tuned to 5.65 MHz (period= 177.5 ns). Meanwhile, the output DC voltage is maintained at around 10 V. The voltage regulation time from Testing Case I to Testing Case III is around 400 ms.

Testing Case IV: If the load is increased from 10 Ohms to 50 Ohms under similar strong coupling condition, once the controller detects an overflow output DC voltage, it tunes the operating frequency closer to the resonant frequency in an effort to regulate the output voltage at a desired level 10 V. The waveforms are

Figure 5.15: Oscilloscope waveform of the output voltage cross the load ($R_L = 10 \Omega$) and the power amplifier output voltage at 1 cm.

shown in Figure 5.16. In the steady state, the output DC voltage was maintained at around 10 V and the frequency was tuned to 5.92 MHz (period= 168.8 ns). The voltage regulation time from Testing Case III to Testing Case IV is around 350 ms.

Figure 5.16: Oscilloscope waveform of the output voltage cross the load ($R_L = 50 \Omega$) and the power amplifier output voltage at 1 cm.

The overall system efficiency is tested based on varying coupling and load conditions. The results of both the overall efficiency and break-down efficiency are shown in Table 5.3 and the comparison with other works is shown in Table 5.4. It can be seen the proposed WPT system has the voltage regulation function and can be applied in a wide dynamic transferring distance. It also can be seen that in case II, the efficiency of the Class-D amplifier drops to 55.1% due to the free-wheeling conduction of the phase-shift modulation [77], while in case II and

IV the efficiency of class D drops to around 50 % since the switching loss do not decrease as the output power decrease.

Table 5.3: Testing results of the proposed WPT system

Testing Cases	DC-AC η (Class D η)	AC-AC η (Coil-Coil η)	AC-DC η (Rectifier η)	Overall η .
I:10 Ohm @ 5cm	85.5%	91.2%	92.2%	71.8%
II:50 Ohm @ 5cm	55.1%	84.6%	92.0%	42.7%
III:10 Ohm @ 1cm	86.3%	93.1%	91.1%	73.1%
IV: 50 Ohm @ 1cm	50.2%	96.6%	92.1%	44.6%

Table 5.4: Performance comparison with the proposed WPT system

Ref.	TBCAS 2008 [153]	AWPL 2014 [110]	TIA 2014 [51]	JSSC 2015 [44]	TPEL 2016 [138]	This Work
Varying Dist.?	Yes	Yes	Yes	Yes	No	Yes
Dist.	$\leq 1\text{cm}$	1-1.4cm	15-35cm	0.3-1cm	N.A.	1-5cm
P_{MAX}	12 W	9 W	N.A.	102 mW	20 W	10 W
Freq.	60-70 kHz	270- 330 kHz	3.8- 6 MHz	13.56 MHz	6.78 MHz	6.78 MHz
Control?	Yes	No	No	Yes	No	Yes
V_{out}	3 V	N.A.	N.A.	3.6 V	22 V	10 V
Effic.	80% @0.5cm DC-DC	90% @1.5cm Coil-Coil	65% @35cm Coil-Coil	50% @0.3cm DC-DC	84% @k 0.13 DC-DC	71% @5cm DC-DC

5.5 Conclusions

In this chapter, we present a detailed empirical analysis on power transfer efficiency and output power of a magnetic resonant wireless power transfer system for both strong and weak coupling modes. It has been derived that the frequencies of output power peak points when frequency splitting happen are only related to the coupling coefficient. In addition, an effective voltage regulation control flow has been developed and experimentally demonstrated to be capable of regulating the output voltage at a constant level over a wide range of transmission distance

(1 cm- 5 cm). Meanwhile, the system efficiency also optimized. The proposed wireless power transfer system is able to transfer 10 W over a 5 cm with an overall efficiency of 71.6 %, and over a 1 cm with an overall efficiency of 73.1%. Compared to the state-of-art technology, the proposed WPT system has a high efficiency within a wide range of coupling conditions.

Chapter 6

Conclusions and Future Work

In this chapter, the main contributions of this research project are summarized and the conclusions of the thesis are presented. Recommendations for the future work of this research project is also proposed.

6.1 Conclusions

In recent years, wireless power transfer (WPT) technology has become more and more popular as it has been touted as a viable solution for non-contact charging of electronic appliances. However, the performance of the technology is still far from satisfying the requirements of the mass consumers market.

The main target of this thesis is to improve the performance of wireless power transfer (WPT) system for consumer electronic devices, particularly in terms of high performance circuits design and the smart control method design. The background of the three main categories of wireless power transfer technology: Electromagnetic (EM) wave WPT, induction WPT and magnetic WPT, and the main building blocks of a WPT system, for instance DC/AC converter, compensation network, rectifier, DC/DC converter and their corresponding control algorithm have been comprehensively reviewed. It has been found it is challenging to design high performance circuits and smart control solutions, and there is still a serious lack in low cost high performance AC-DC converter, reverse leakage control and smart control which can overcome the frequency splitting mentioned in

Chapter 5.

In this dissertation, a single stage AC/DC voltage regulator for the induction WPT has been proposed, which is to replace the traditional rectifier plus DC-DC converter or voltage regulation with wireless communication approach. The cross-coupled full active rectifier has been adopted and a fast response comparator has been used to eliminate the reverse leakage current. The power loss, voltage gain and transfer function is analyzed. A voltage PWM control is used to provide the regulator with a constant and stable 5 V output. The operating frequency of the WPT is 125 kHz and the switching frequency of the regulator is 7 kHz. This novel regulator design with the proposed control mechanism was fabricated based on AMS 0.35 μm high voltage CMOS technology. For measurement and testing of the regulator, the load resistance was adjusted between 1.6 Ohms and 25 Ohms and the output voltage ripple was maintained at about 40 mV. The achievable peak efficiency of the regulator is at 92 % and this value can be further improved if a BCD process or GaN devices is used. When compared to the existing works, this regulator is a true single stage design without extra DC-DC converter and other peripheral communication circuits.

Subsequently, a full active rectifier for the magnetic wireless power transfer application has also been designed and physically implemented. The proposed full digital adaptive delay control is able to work for a WPT system at 6.78 MHz operating frequency. A SAR algorithm is used to align the gate driving signal to the AC input voltage. As a result, the reverse leakage current is eliminated, which in turn successfully optimized the efficiency of the rectifier. To further verify the proposed full active rectifier, this design was fabricated in AMS 0.18 μm CMOS technology. Measurements were carried out based on a wide output power range from 15 W to 1 W. The fabricated rectifier has a peak efficiency of 90 % at the output of 5 V and 2 A. It has been proven that this method effectively removed the reverse leakage of the rectifier.

Finally, a magnetic WPT system coupled with adaptive voltage regulation

and efficiency optimization capability for 15 W applications has been developed. For a high frequency WPT system in a strong coupling mode, the power transfer capability of the system at the resonant frequency is pulled down due to the frequency splitting phenomena. Conversely, while in the loosely coupling mode, the system has to work at resonant frequency to maintain the efficiency at a high level. A new voltage regulation technique aimed at circumventing the frequency splitting issue and enhancing efficiency optimization has been developed to solve this problem. Besides, a zero voltage switching class D amplifier is designed to decrease the power loss of the DC/AC stage. This design has been verified using GaN devices, micro-controller, and ZigBee communication models and it has been corroborated that the system is able to provide a constant 10 V from 15 W to 1 W and capable of achieving peak efficiency of 73 % . This control method is able to provide a stable output voltage and keep the performance (power transfer efficiency and capability) of the system at a high level. To the best of our knowledge, this is the first time a WPT system with such functions has ever been developed and physically implemented.

6.2 Future Work

For the single stage regulator, the efficiency for light load condition still has room for improvement. One possible solution is to use envelope detection method to detect the duty-cycle of charging-mode and zero-mode so as to adjust the output voltage of the AC/DC converter. In light load condition, the power loss in zero-mode, which is mentioned in Chapter 3, deteriorates the efficiency. In that case, the AC/DC converter is supposed to be able to reduce the output voltage to reduce the zero-mode duty-cycle.

For the 6.78 MHz wireless power transfer system, the ZVS class D amplifier could be further improved. Since in the phase shift tuning operation, there is a freewheel duration for light load condition, the power loss in this duration will pull down the efficiency. This could be solved by changing the input voltage in-

stead of changing the phase shift. Meanwhile, although the power consumption of ZigBee communication is lower than Bluetooth communication, it is slower in the meantime. In that case, it is possible to consider using Bluetooth communication for future designs.

APPENDIX A

Phasor Transformation

Phasor transformation is used to transfer a rotary AC circuit into an equivalent DC circuit model. The basic principle of the Phasor transformation is to find a DC stationary sub-circuit model for each AC rotary circuit components, which makes it much easier to find the value of envelopes of AC rotary variables in an AC rotary circuit. Some traditional transformation like D-Q transformation [DQ] can only solve the switching and transformer transformation, but not able to deal with the inductor, capacitor and so on. So in this thesis, the phasor transformation is adopted. To build a valid model for WPT system, the equivalent component model has to be able to reflect not only the magnitude and the phase of a sinusoidal signal in the steady state but also the response of the transient state. The phasor which can represent any sinusoidal signal is:

$$X(t) \equiv \Re[\sqrt{2}x(t)e^{(j\omega t)}] \quad (\text{A.1})$$

where the $x(t)$ indicates the complex time-varying variable. It is worth to note that the $x(t)$ and ω are not necessary to be sinusoidal or constant but also can be non-sinusoidal function only when the $x(t)$ is sinusoidal with the frequency of ω . In the steady state, $x(t)$ becomes just the conventional phasor. The three basic circuit elements of the WPT system: inductor, capacitor and resistor, are phasor-transformed respectively, as follows.

Inductor Phasor Transformation

The procedure of the phasor transformation for the inductor is shown in Figure A.1. The voltage-current circuit equation is

Figure A.1: Inductor phasor transformation. (a) AC-Rotational circuit. (b) DC-Stationary circuit.

To transform the current and voltage of the inductor to phasor, it would have:

$$\begin{cases} i_L \equiv \Re[\sqrt{2}i_L e^{(j\omega t)}] \\ v_L \equiv \Re[\sqrt{2}v_L e^{(j\omega t)}] \end{cases} \quad (\text{A.2})$$

Applying Equation (A.2) to Equation (A.1) results in

$$L \frac{d}{dt} [\Re[\sqrt{2}i_L e^{(j\omega t)}]] = \Re[\sqrt{2}v_L e^{(j\omega t)}], \quad (\text{A.3a})$$

$$\text{or} \quad L \Re\left[\frac{di_L}{dt} e^{(j\omega t)} + j\omega i_L e^{(j\omega t)}\right] = \Re\left[\left(\frac{di_L}{dt} + j\omega i_L\right) e^{(j\omega t)}\right] = \Re[v_L e^{(j\omega t)}]. \quad (\text{A.3b})$$

Equation (A.3) is just equivalent to

$$L \frac{di_L}{dt} + j\omega L i_L = v_L. \quad (\text{A.4})$$

The equivalence of Equation (A.4) with Equation (A.3) can be proved as follows.

Theorem

For any x , y and $\omega \neq 0$, $\Re[xe^{(j\omega t)}] = \Re[ye^{(j\omega t)}]$, if and only if $x = y$.

Proof

$$\begin{aligned}
\Re[xe^{(j\omega t)}] &= \Re[|x|e^{(jarg(x))}e^{(j\omega t)}] \\
&= |x|\Re[e^{(jarg(x)+j\omega t)}] \\
&= |y|\Re[e^{(jarg(y)+j\omega t)}] \\
&\Leftrightarrow |x| = |y| \quad \text{and } arg(x) = arg(y) + 2\pi n, n = 0, 1, 2, \dots \\
&\Leftrightarrow x = y
\end{aligned}$$

It is gain from Equation (A.4) the form of circuit in Figure. A.1(a) could be transferred to Figure. A.1(b), for establish DC analysis for the WPT system. The imaginary resistor, $j\omega L$, the meaning of which is just conventional inductor impedance in the steady state. However it should be noticed that this circuit model in Figure. A.1(b) is still valid in transient analysis for the DC stationary circuit model. The imaginary resistor does not vanish even though it is in the transient state.

Capacitor Phase Transformation By a similar procedure with the inductor case, the rotationary circuit equation of Figure A.2(a),

Figure A.2: Capacitor phasor transformation. (a) AC-Rotationary circuit. (b) DC-Stationary circuit.

is changed by the phasor transformation to

$$C \frac{dv_C}{dt} + j\omega C v_C = i_C. \quad (\text{A.5})$$

The circuit model of the DC stationary model of Equation A.2 is shown in Figure A.2(b).

Resistor Phasor Transformation

The phasor transformation of a resistor, as shown in Figure A.3 is straightforward since

$$v_R = \Re[\sqrt{2}v_R e^{(j\omega t)}] = i_R R = \Re[\sqrt{2}i_R e^{(j\omega t)R}], \quad (\text{A.6a})$$

$$\text{or} \quad v_R = i_R R. \quad (\text{A.6b})$$

Figure A.3: Resistor phasor transformation. (a) AC-Rotational circuit. (b) DC-Stationary circuit.

APPENDIX B

Author's Publications

B.1 Journal Papers

- **Xiaoyin Bai**, Zhi Hui Kong and Liter Siek, “A Single-Stage AC-DC Voltage Regulator for 15W Wireless Charging,” submitted to *IET Electronics letter*.
- **Xiaoyin Bai**, Zhi Hui Kong and Liter Siek, “A High-Efficiency 6.78 MHz Full Active Rectifier with Adaptive Delay Control for Wireless Power Transmission,” Accepted by *IEEE Transactions on Very Large Scale Integration (VLSI) Systems*.
- **Xiaoyin Bai**, Zhi Hui Kong and Liter Siek, “Analysis and Hybrid Control Method for Wireless Power Transfer System with Varying Transmission Distance,” submitted to *IEEE Antennas and Wireless Propagation Letters*.

B.2 Patent Proposals

- **Xiaoyin Bai**, Zhi Hui Kong, Liter Siek and Tan Yen Kheng, “A 92% Efficiency 15 W Single-Stage Wireless Charging Receiver With 5 V DC Voltage Regulation”,SG Provisional Patent Application No. 10201404736R had been re-filed in 2016 Jan .

- **Xiaoyin Bai**, Zhi Hui Kong, Liter Siek and Tan Yen Kheng, “A Single-Stage AC-DC Voltage Regulator for Wireless Charging”,SG Provisional Patent Application No. 10201404736R. Filed in 2014 September.

B.3 Conference Papers

- **Xiaoyin Bai**, Zhi-Hui Kong and Liter Siek, “A novel control method for magnetic wireless charging system”, *IEEE International Conference on Electron Devices and Solid-State Circuits* (EDSSC). June. 2015
- **Xiaoyin Bai**, Yen-Kheng Tan, Zhi-Hui Kong and Liter Siek, “Novel Active Tuning Approach for Resonant-mode Wireless Charging System”, *IEEE International Conference on Industrial Technology* (ICIT). Feb.-March. 2014
- **Xiaoyin Bai**, Yen-Kheng Tan, Tran Duong, Zhi-Hui Kong and Siek Liter, “A Wireless Charging Resonant Frequency Checking Technique for LVDC”, *Asia Smart Grid 2013*. Oct. 2013

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