Wireless power transfer systems

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WIRELESS POWER TRANSFER SYSTEMS

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2018

WIRELESS POWER TRANSFER SYSTEMS

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Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research and has not been submitted for a higher degree to any other University or Institution.

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Abstract

Wireless power transfer (WPT) is a potential technology that enables transmission of electrical power without the necessity of electrical contact. WPT can be employed in a wide variety of applications and this can be achieved due to WPT's inherent advantages over its wired counterpart. These advantages enable WPT to eliminate numerous drawbacks faced by its wired counterpart. However, even with much research interest in this technology, WPT still need to overcome numerous technical challenges before it can truly be ubiquitous in the modern connected world. This thesis has proposed several approaches to improve WPT's performance in terms of efficiency, transferred power, etcetera. In addition, a novel control scheme for dynamic WPT (D-WPT) systems without needing external sensors, and this helps solve some complexity and implementation issues of a D-WPT system.

WPT systems use inductive coils and its electromagnetic coupling to transfer power wirelessly without electrical contact. These loosely coupled inductive coils are powered by high frequency power sources and cause large leakage inductances and input impedances. By employing compensating capacitors and electromagnetic resonance operation, the input impedance and VA (volt-ampere) rating of the WPT system can be significantly reduced, leading to an increase in the performance in terms of efficiency and power transfer, as well as transfer gap distance. However, there are many different types of compensation circuits available for WPT systems and these compensation circuits perform differently under varied conditions. The thesis dwelled into the circuit analyses on the various compensation circuits and derives the efficiency optimal operating frequencies these WPT systems. Using the optimized operating frequency, the efficiency analysis of these compensation circuits are detailed. Resonant frequency responses of the compensation circuits are used to further differentiate. With these analyses, the efficiency orientated design flow chart for selecting the

optimized compensation circuit is derived with consideration for efficiency, load resistances, transferred power, transconductance and voltage gain. Experimental results also verified the usefulness of the flow chart and its respective analyses, which showed efficiency discrepancy of 28% (highest at 500 Ω) equivalent output resistance when comparing between compensation circuits. The LCC compensation circuit is also analyzed and its frequency bifurcation effect is investigated to be significantly mitigated. The LCC compensation circuit also shows significantly higher transferred power and lower harmonic distortion

The concept of Dynamic WPT (D-WPT) is to allow the receiver (Rx) to move freely while receiving power from the D-WPT system wirelessly. The thesis proposes a novel control scheme for the purpose of individual magnetic coupler control using the sensed transient response of the magnetic coupler's input current. This would improve the D-WPT system's efficiency, while circumventing the necessity for external positional sensors or auxiliary circuitry. The proposed control algorithm yielded a low standby power to operating power percentage of 7.2% as compared to 44.2% found in existing literatures. Thus, the proposed control algorithm for D-WPT system could be a promising approach to reduce D-WPT system complexity, improve power consumption and EMI/EMC performances during standby mode, as well as power transfer efficiency levels.

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Abbreviations

WPT	Wireless Power Transfer
Rx	Receiver
Tx	Transmitter
M	Mutual inductance
k	Coupling coefficient
EV	Electric vehicle
AV	Autonomous vehicle
RF	Radio frequency
S	Series
P	Parallel
PA	Power Amplifier
SOC	State of charge
VA	Volt-ampere
LCC	Compensation circuit (series inductor, parallel and series capacitors)
LCL	Compensation circuit (series inductor and parallel capacitor)
PTE	Power transfer efficiency
TP	Transferred power
FHA	Fundamental harmonic analysis
ZVS	Zero voltage switching
THD	Total harmonic distortion
EMI	Electromagnetic interference

Chapter 1

Introduction

This chapter provides an introduction to WPT systems, applications and achievements. The motivation, objectives, scope and contributions of this thesis are highlighted as well.

1.1. Background

1.1.1. Introduction to Wireless Power Transfer system

The first wireless power transfer (WPT) via electrostatic induction means was demonstrated by Nikola Tesla over a century ago [1]. Due to the technical limit of semiconductor technology at that point in time, developments were limited and restricted. In recent decades, WPT has been gaining attention with the advent of enabling modern semiconductor and winding technologies.

A WPT system can be divided into two main parts: transmitting (Tx) and receiving (Rx) as shown in Fig. 1.1. On the Tx side, the DC or AC power source is connected to the power electronic converter which generates the high frequency AC current. This high frequency AC current is used to magnetize the Tx coil which magnetically couples with the Rx coil. The magnetic field created by the Tx coil induces voltage in the Rx coil, transferring electrical energy over a predefined air gap. The wirelessly transferred energy can be converted into suitable form to drive a load. The setup of WPT is similar to transformers where a magnetic core provide a strong coupling between the respective coils. However, in a WPT system, the Tx and Rx coils are loosely coupled as these coils are placed significantly apart. As a result of the low coupling, and the Tx and Rx coils' inductive nature, there is significant leakage inductance. Thus, the overall impedance of the system increases. To mitigate such effect, compensation circuits can be connected to the Tx and Rx coils to reduce or eliminate unnecessary reactance to maximize the magnetizing current and magnetic field.

Fig. 1.1 A typical WPT system

1.1.2. WPT Applications

One major application of WPT is electric vehicles (EV) is illustrated in Fig. 1.2. Due to climate change, it is widely accepted that emission of greenhouse gases must be reduced and energy production to be shifted to renewable energies. Thus, with the global trend of renewable energies, EV is poised to reduce our dependency on petroleum and enable the use of electricity which is often domestic and renewable. Many governments have been granting subsidies to EV, which have spurred the sales of EVs in recent years. It has been projected that there will be 20 million EV on the road worldwide by 2020 [2]. However, when compared to conventional gasoline vehicles, EV still lags behind in terms of driving range and charging time. The limited driving range has pushed manufacturers to increase the battery size. Even though Tesla's EV Model S P100D (which is the world's longest range production EV) is able to exceed 300 miles of range on a single charge, its 100 kWh battery pack significantly adds to the EV price tag of \$134,500 [3]. Also, the world's fastest charger, Tesla supercharger, is able to provide 170 miles in a 30 minutes charge [4], which pales in comparison with the few minutes of the conventional gasoline vehicle refueling. By implementing dynamic WPT (D-WPT), the above issues can be mitigated by wirelessly charging the EVs while in motion which can improve its

overall driving range, reduce battery capacity as well as the need for recharging. Static WPT can also improve the reliability, usability and safety as the contactless technology does not require users to plug and unplug connectors and charging points can be better weather proofed.

Fig. 1.2 WPT to wirelessly charge EV [5]

Another promising application of WPT is in implantable biomedical devices. Due to the operation requirements of such devices, they are usually located within the body as shown in Fig. 1.3. Traditionally, these implantable devices are powered by implantable batteries or percutaneous wires. These forms of power delivery severely limit the performance of the implantable devices and also expose the users to infections through the wires or the invasive surgical procedures to replace batteries. Instead, transcutaneous power transfer can be achieved by WPT technology. WPT offers an unlimited power source for these biomedical implantable devices and alters the operating norms of such devices. For example, depending on the patient, the battery in the conventional pacemaker has to be replaced within 5 to 15 years via an invasive surgical procedure which can be easily avoided by implementing WPT technology [6]. Infections arising from the wires which deliver power to the implantable device or battery can also be reduced.

Fig. 1.3 Implantable biomedical devices [7], [8]

Another challenging application is mobile consumer devices such as mobile phones. Mobile phones have become so ubiquitous in our lives that there are more mobile phones than human beings now [9]. However, adoption of WPT in mobile phones have been received lacklustre response due to the technical challenges of employing WPT in such challenging environment (high misalignment, large gap distance and small coil sizes, metallic mobile phone housing, high powered charging, etcetera). Instead, the wired counterpart has seen improvements in convenience, particularly the invention of reversible connectors USB type C and Apple's lightning connector and Qualcomm's QC3.0. In terms of charging speed, for example, WPC Qi wireless fast charging supports up to 15 W which is comparable to Qualcomm's QC2.0 wired standard [10]. However, wired standards Qualcomm's QC3.0 and USB type C are able to support up to 18 W [11] and 100 W [12] respectively. As such, adoption of the WPT in mobile phones has significantly slowed during recent years. To gain traction, WPT has to improve its misalignment, gap distance and charging power.

Fig. 1.4 WPT to charge multiple mobile devices (Rezence) [13]

1.1.3. WPT Achievements

In the recent few decades, the progress of WPT has been spurred due to the development of modern power electronics and winding technologies. This development has enabled various research groups in WPT to attain the following achievements:

- In 2007, researchers from MIT were able to transfer 60 watts with about 40% efficiency over distances in excess of 2 meters [14].
- In 2011, researchers from The Hong Kong Polytechnic University have demonstrated the flexibility of WPT in various domino forms. A 14 W compact fluorescent lamp has been powered by a circular domino repeater arrangement as shown in Fig. 1.5 [15].
- In 2012, researchers from Tsinghua University demonstrated WPT over a distance of 1 meter from the floor to the battery-less capsule endoscopy via the power relay in jacket as seen in Fig. 1.6 [16]. Even though the efficiency is considerably lower than MIT's work (3%), the small Rx coil

used (within the capsule at 0.11 cm in radius), makes this a remarkable achievement.

- In 2012, researchers from Utah State University accomplished a WPT of 5 kW at 90% efficiency over a gap of 20 cm [17].
- In 2014, researchers from University of Michigan exhibited WPT by transferring 7.7 kW power over a gap of 20 cm with an efficiency of 96% [18].

Fig. 1.5 14 W compact fluorescent lamp powered by circular domino repeater arrangement [15]

Fig. 1.6 Wirelessly powered capsule endoscopy via jacket power relay [16]

On the commercial side, fully functional WPT systems have been employed in the real world.

- In 2012, Utah-based company Wireless Advanced Vehicle Electrification (WAVE) deployed electric buses to University of Utah's shuttle bus system. Charging pads are installed at bus stops to charge the electric buses in order to minimize onboard energy storage [19].
- Korea Advanced Institute of Science and Technology (KAIST) developed Online Electric Vehicle (OLEV) which is an electric vehicle that can be charged while stationary or driving. In 2013, OLEV was introduced into the public transit network and run an inner city route for a total of 24 km roundtrip. The OLEV charging system allows the OLEV to reduce its battery capacity down to one-third of the battery equipped with a regular electric car [20].

Fig. 1.7 WAVE (left) and OLEV (right) electric bus [20], [21]

Also, in order to standardize operating standards for mobile devices, several consortiums have been implemented.

- In 2008, the Wireless Power Consortium was founded and its members consists of Samsung, Qualcomm, Verizon, and etcetera [22] .
- In 2015, two WPT alliances namely, The Alliance for Wireless Power (A4WP) and Power Matters Alliance (PMA), merged to become AirFuel alliance. Its members consists of Intel, WiTricity, AT&T, and etcetera [23].

1.2. Motivation

The application of WPT will continue to diversify along with the development of modern technologies. WPT has the potential of overcoming the physical limitations of conventional wired power in the following areas:

- Operation flexibility
Compared to the rigidness of wired power, WPT is able to operate in environments which are previously impractical. Powering a mobile robot to operate in inaccessible places and charging EVs while in-motion are possible with WPT technology. WPT would enable driving-autonomous EVs to be charged automatically without any human intervention, creating a truly autonomous transportation tool.

- Convenience

The ease of operation without wired power would enable seamless charging of EVs, implantable biomedical devices, mobile devices, etcetera, which would significantly change our way of living.

- Reliability and low maintenance

WPT will eliminate the need for mechanical and electrical contacts, preventing wear and tear and lowering maintenance costs. Also, WPT allows the system to be fully waterproof and dustproof which improves reliability and product lifespan.

- Safety

WPT allows the system to transfer power without the use of external contacts. This eliminates the risk of electrical shock to the user. For implantable biomedical devices, WPT allows the user to be safe from infections arising from charging cables or invasive battery change surgeries.

- Isolation

Apart from the physical isolation (fully insulating the system from the environment), WPT also offers additional electrical safety: electrical isolation of Tx and Rx sides.

- Wide power application

WPT power delivery varies from milli-watts for implantable biomedical devices to kilo-watts for EV battery charging.

- Freedom in Design

The wireless charging technology allows for innovative designs which would otherwise be impossible with wired conductors.

The above mentioned potentials have propelled WPT to become one of the most popular research topic. Although wireless charging has been implemented in several applications ranging from low, medium to high powered applications, WPT systems are still being researched with the aim of optimizing its performances such as power transfer efficiency, coils misalignment, coupling

enhancement techniques, and etcetera. In addition, D-WPT system allows the WPT to take place while the Rx is in motion. This greatly expand the operation usability and flexibility of the system. For example, EV no longer needs to be stationary in order to receive power wirelessly. Instead, EV can still be wirelessly charged from Tx's placed along the EV's path of motion via D-WPT.

1.3. Objective and Scope

The objective of this thesis is to improve the overall performance of a WPT system by implementing a design methodology for compensation circuits under various load conditions. The improvements include power transfer efficiency, gap distance and misalignment tolerance. In addition, a novel control strategy is proposed for the control of D-WPT which allows the control to be implemented at the Tx side, simplifying communication and complexity of the WPT system.

1.4. Contributions of the Thesis

The thesis addresses:

Chapter 1 provides a rationale for this research and outlines the goals and scope of this thesis.

Chapter 2 reviews the literature concerning WPT systems. The WPT systems are categorized into static and dynamic WPT systems. The different types of compensation circuits used in WPT systems are also covered.

Chapter 3 discusses the design principles of WPT systems and its compensation circuits, according to the application. The various compensation circuits are analyzed, characterized and analytically formulated to create a design flowchart where the main optimization (of the compensation circuit) goal is power transfer

efficiency. Other performance metrics such as transferred power, transconductance, and voltage gain are also covered in this chapter.

Chapter 4 discusses the characteristic and performance of the LCC compensated WPT system. The circuit analysis, equivalent circuits, impedance frequency response, transferred power and efficiency as well as harmonic performance are covered.

Chapter 5 discusses the novel control algorithm for D-WPT applications. The derivation of the proposed control algorithm is discussed in depth as well as its operation. The chapter also evaluates the performance of this algorithm and benchmarks the power and energy losses with other D-WPT systems.

Chapter 6 concludes the thesis and recommends possible research areas that can be further explored in the future.

Chapter 2

Literature Review

This chapter provides a detailed literature review on the various types of WPT namely, static WPT and dynamic WPT systems. Applications propelling WPT forward include electric vehicles, biomedical implants and consumer electronics. WPT compensation circuits are also briefly covered. Lastly, the performance indices for the WPT system are discussed.

2.1. Overview of WPT systems

The wireless power transfer (WPT) technology enables the transmission of electrical energy without requiring physical connections. WPT can be achieved via different technologies such as radio frequency (RF) waves, microwaves, laser, inductive coupling, capacitive coupling, magnetic resonance. While most of these technologies leverage on electromagnetic wave principles (except capacitive coupling), these technologies have different operation principles and mechanisms. Generally speaking, these technologies can be broadly categorized according to the transmission range, namely, near and far field.

RF waves, microwaves and laser can be considered as far field WPT approaches as the transfer distances are usually significantly longer than the wavelength of its electromagnetic wave. These technologies can transfer power wirelessly up to tens of thousands of kilometres (i.e. from solar satellites to ground rectenna) and their wavelengths are short (operating frequencies from MHz to GHz). However, the far field WPT approaches suffer from poor efficiencies and are directional, which requires precise alignment for power transfer to take place.

On the other hand, inductive, capacitive and magnetic resonance coupling, are considered near field WPT approaches as the transfer distances are shorter than the operating wavelengths. In comparison, near field WPT approaches do not need to be precisely aligned and are able to transmit high power at high efficiency levels. However, efficiency of the near field WPT approaches rapidly deteriorates with increasing transfer distance.

Inductive coupling based WPT systems require high frequency electric currents to excite the Tx (primary coil). The operating frequency ranges from 10 kHz to MHz [24]. The high frequency electric current generates an electromagnetic field from the Tx, which inductively couples with the Rx (secondary coil), thereby establishing power transmission. This approach is regarded as uncompensated as inductive impedances of the primary and secondary coils are significant (due to the high operating frequency) at both the input and output of the WPT system [25].

In order to reduce such impedances, compensating capacitors are added to both the primary and secondary coils. The addition of capacitance (in the form of externally lumped capacitors) leads to the magnetic resonance approach where the inductive impedances of both coils are compensated with the capacitive impedances [26]. By operating the WPT system at its resonance frequency, higher power transfer, efficiency and transfer gap can be achieved as compared to the uncompensated inductive coupling approach [14]. Also, the parasitic coil capacitance can be used to compensate the coil inductance. However, the required operating frequency by relying on the coil parasitic capacitance has to be significantly higher in order to reach the coil self-resonance frequency [27], [28].

2.1.1. Static WPT system

In static WPT systems, the electrical loads are stationary during the transmission of power. These loads include EVs [29]–[31], biomedical implants [32]–[35], consumer electronics [36]–[43], etcetera. By electromagnetic coupling means between two or more electromagnetic couplers (or coils), the transmission of power can be established wirelessly through a separation gap. The working principle of inductive coupled WPT is illustrated in Fig. 2.1, where the high frequency electric current in the transmitter coil generates the magnetic field (B). The magnetic field (B) generated by the transmitter coil couples with the receiver (which has a separation distance of z) [44]. The electromagnetic field produced by the transmitter coil induces an electric current at the receiver coil. By this electromagnetic coupling method, the WPT system is capable of transferring power of microwatts to kilowatts at efficiencies of more than 90% [45]–[47]. This electromagnetic working principle allows the inductive coupled WPT system to be immune to dust, dirt, water, wear and tear, unlike wires and connectors which are susceptible to its operating environment.

A promising proponent of static WPT is EV charging. By employing static WPT systems, the charging cables and connectors can be replaced which significantly

reduces running costs due to maintenance, theft, etcetera. Static WPT system can also be implemented in Autonomous Vehicle (AV) charging routine, effectively removing the need for user intervention for charging purposes. With the said benefits, static WPT does not effectively solve the limited driving range of the EV and has slower charging speed. Dynamic WPT (D-WPT) systems can be supplementary used to solve these issues.

Fig. 2.1 Inductively coupled WPT system [44]

2.1.2. Dynamic WPT system

In D-WPT systems, the electrical loads are moving and mobile during the transmission of power. For example, this would allow EVs, ingestible biomedical electronic sensors and consumer electronic devices to be continuously powered while in motion. For the application of D-WPT in EVs, the implementation of D-WPT infrastructure would enable ‘opportunistic charging’ for the EVs to be charged while in motion, significantly alleviating the ‘distance anxiety’ issue faced by most EVs owners and extending the EVs’ driving ranges far beyond its battery capacities. The ‘opportunistic charging’ also reduces the amount of time needed for the EV to be charged via static WPT system, if necessary.

There are generally two main categories of D-WPT systems: track and segmented coils based. Track based D-WPT system mainly consists of a single long transmitter track which couples with the significantly shorter receiver. Some of

the notable implementations of track based D-WPT system are UC Berkeley [25] and Korea Advanced Institute of Science and Technology (KAIST) [48]–[50]. KAIST has demonstrated track based D-WPT system using the online electric vehicle (OLEV) and has achieved 20 kW power transfer over 20 cm at 83% efficiency, using Shaped Magnetic Field in Resonance (SMFIR) transmitter rails. These rails are 5 meters to 60 meters long, while the pickup module is 80 cm long [48], translating to significantly lower coupling levels and a very small relative transfer gap (in between transmitter and pickup modules) distance. The large transmitter rails also contribute significantly to the EMI/EMC issues as well as the unwanted human exposure to magnetic fields.

On the other hand, segmented coil based D-WPT system uses an array of transmitter coils which is connected to high frequency power sources as shown in Fig. 2.2. Oakridge National Laboratory (ORNL) demonstrated segmented coil based D-WPT system [51]–[53] to transfer 2 kW over 10 cm using 33 cm diameter coils to power EV with a maximum efficiency of 85% [54].

Fig. 2.2 Typical D-WPT system

Although a segmented coil based D-WPT system reduces the coupling and EMI/EMC issues faced by the track based D-WPT system, there are additional challenges to implement a segmented coil based D-WPT system. Segmented coil based D-WPT systems are more susceptible to the coils' relative displacement. This would impact on the coupling of the electromagnetic fields and ultimately affect efficiency and transferred power levels. In order to smooth out the power flow, ORNL has demonstrated the use of electrochemical capacitor to reduce power variations [51]. In addition, there is a need to control each segmented coil to optimize the operating efficiency and to make sure that there is no unnecessary magnetic field formed without the presence of the mobile load. In [55], an auxiliary LCC compensation is used to enable automatic power distribution of the segmented coils which are connected to a single high frequency power source. The auxiliary circuit is used to regulate the coils' current without the need of any external controller or sensor network.

2.2. WPT Compensation Circuits

In general, WPT compensation circuits can be used to reduce VA rating of the WPT system, increase the power transmission level and improve system efficiency. Basic WPT compensation circuits are capacitive and can be connected according to different configurations and requirements. Fig. 2.4 shows the four different basic configurations, namely, series-series (S-S), series-parallel (S-P), parallel-series (P-S) and parallel-parallel (P-P). Using Kirchhoff's law, the respective system input impedance equations can be derived in (2.1) - (2.4) [56].

For S-S compensation circuit:

$$Z_{in} = Z_{C1} + Z_{L1} + \frac{(\omega M)^2}{Z_{C2} + Z_{L2} + Z_L} \quad (2.1)$$

$$Z_{in} \approx \frac{(\omega_0 M)^2}{Z_L}$$

For P-S compensation circuit:

$$Z_{in} = Z_{C1} - \frac{Z_{C2}^2 (Z_S + Z_L)}{Z_P (Z_S + Z_L) + (\omega M)^2} \quad (2.2)$$

$$Z_{in} \approx Z_{C1} - \frac{Z_{C2}^2 Z_L}{(\omega_0 M)^2}$$

For S-P compensation circuit:

$$Z_{in} = Z_P + \frac{(Z_{C2} + Z_L)(\omega M)^2}{Z_S (Z_{C2} + Z_L) - Z_{C2}^2} \quad (2.3)$$

$$Z_{in} \approx k^2 (Z_{C2} + Z_L)$$

For P-P compensation circuit:

$$Z_{in} = Z_{C1} - \frac{Z_{C1}^2 [Z_S (Z_{C2} + Z_L) - Z_{C2}^2]}{Z_P [Z_S (Z_{C2} + Z_L) - Z_{C2}^2] + (\omega M)^2 (Z_{C2} + Z_L)} \quad (2.4)$$

$$Z_{in} \approx Z_{C1} + \frac{Z_{C1}^2 Z_{C2}^2}{(\omega_0 M)^2 (Z_{C2} + Z_L)}$$

These basic compensation circuits are usually chosen due to their simplicity and can be compatible with different sources and loads. For example, for WPT systems with series transmitting compensation circuits (S-S and S-P), voltage sources can be used (low input impedances as shown in (2.1) and (2.3)), while parallel transmitting compensation circuits can be used with current sources (high input impedances as shown in (2.2) and (2.4)). On the secondary side, series receiving compensation circuits are suitable for higher powered loads (lower load resistances) while parallel receiving compensation circuits are suitable for lower powered loads (higher load resistances). It should be noted that only the S-S compensation circuit maintains its resonance frequency regardless of the coupling and load conditions. However, the voltage across the series compensating capacitors are significantly large, and high powered applications

(such as EV and transportation) tend to choose the parallel compensation circuits for their current limiting capabilities [56], [57].

Fig. 2.3 WPT system with basic compensation circuits: (a) S-S, (b) P-S, (c) S-P and (d) P-P

2.2.1. Other Compensation Circuits

Besides the basic compensation circuits, other compensation circuits are proposed over the basic compensation circuit in order to improve the WPT system. The LCL compensation circuit consists of a series inductor and a parallel capacitor, as shown in Fig. 2.4 (a). The LCL is reported to reduce VA ratings, increase power and efficiency levels and notably, to behave as a current source when operated at its resonance frequency [58]–[60]. This behaviour will allow the WPT system to control the power flow by regulating the transmitter current via the current source. In order to improve on the LCL compensation circuit, the CLCL compensation circuit, as shown in Fig. 2.4 (b), reduces the harmonic distortion of the input current [61] which can increase switching losses and reduce the efficiency of the LCL compensated WPT system [62]. However, the control for the high order CLCL compensated WPT system is complicated and expensive. The load transformation compensation circuit shown in Fig. 2.4 (c)

overcomes the loading effect of the S-S compensation circuit. Although the load transformation compensation circuit improves the efficiency at higher loading as compared to S-S compensation circuit [35], adding an additional inductor would increase the required footprint and its parasitic resistance will affect efficiency. Thus, the implementation of the load transformation compensation circuit has to be largely determined by the application and its operating conditions. Last but not least, the LCC compensation circuit improves on the harmonic performance of the LCL compensation circuit by introducing another capacitor to achieve unity power factor as shown in Fig. 2.4 (d) [29]. The LCC compensation circuit maintains the current source capability of the LCL compensation circuit, which refers that the current in the transmitter coil would be controlled by the high frequency power source regardless of the coupling and load conditions. This simplifies the power control on the transmitter side and this benefit is favorable from the perspective of D-WPT systems, as such systems tend to have dynamic coupling and load conditions [18], [55], [63]. However, during the stand-by phase where the receiver is not in the vicinity of the transmitter, efficiency and EMI issues arise due to the constant current in the transmitter coil.

Fig. 2.4 WPT system with other compensation circuits: (a) LCL, (b) CLCL, (c) Load transformation and (d) LCC

2.3. Circuit Analysis

In order to evaluate the respective compensated WPT systems, circuit analysis via a simple and accurate modelling method can be performed. The circuit analysis would also quantify system performance indices to facilitate component selection as well as system optimization.

The basic S-S compensated WPT system powered by a high frequency AC source is shown in Fig. 2.5. The high frequency AC source can either be a Power Amplifier (PA) or a power electronic converter (inverter). The transmitter and receiver coils are coupled together with mutual inductance of M which facilitates the wireless transmission of power. At the other end, the receiver coil and its compensation circuit can be connected to a battery charging circuit. The battery charging circuitry load can be approximated to resistive AC load, R_L which will be varied according to the battery's state-of-charge (SOC).

Fig. 2.5 Series-series compensated WPT system

For loosely coupled WPT systems, the leakages and mutual inductances can be modelled into a two port network equivalent circuit as shown in Fig. 2.6. The circuit equations of the S-S compensated WPT system is as shown:

$$\begin{bmatrix} V_{in} \\ 0 \end{bmatrix} = \begin{bmatrix} Z_{C1} + Z_{L1} + R_1 & j\omega M \\ j\omega M & Z_{C2} + Z_{L2} + R_2 + R_L \end{bmatrix} \begin{bmatrix} i_T \\ i_R \end{bmatrix} \quad (2.5)$$

2.3.1. Performance Indices

With the two port network equivalent circuit, the performance indices of the WPT system can be derived. The key performance indices of the WPT system are power transfer efficiency (PTE) and transferred power (TP). The PTE of the S-S compensated WPT system as shown in Fig. 2.6 is as follows:

$$PTE = \frac{|i_R|^2 R_L}{|i_T|^2 R_1 + |i_R|^2 (R_2 + R_L)} \quad (2.6)$$

Next, TP is the normalization of the WPT system output power with its input voltage, v_{in} . The TP of the S-S compensated WPT system is as defined:

$$TP = \frac{P_{out}}{v_{in}^2} \quad (2.7)$$

The performance indices allow the optimization of the WPT system in terms of the lumped components, coupling coefficient and loading resistance.

Fig. 2.6 Equivalent circuit of series-series compensated WPT system with leakage inductances

Chapter 3

Compensation Circuits for WPT

WPT systems employ a wide variety of compensation circuits. These compensation circuits perform differently under varied conditions. In this chapter, the efficiency-optimum operating frequency for the compensation circuits is determined by using induced and reflected load equivalent circuits. Next, by using the optimized frequency, a detailed efficiency analysis on the load variation of the basic compensation (series-series, series-parallel) and series-LCL circuits is presented. The efficiency analysis is supplemented with the introduction of crossover resistance which is the point in load resistance when the respective circuits' efficiencies intersect. Additionally, resonant frequency response analysis is used to further differentiate these circuits. The efficiency-orientated design flow chart is constructed with the theoretical analysis presented in this chapter. This is followed by transferred power, transconductance and voltage gain analyses of selected circuits. Lastly, experimental verification of the series-series and series-LCL results validate the presented analysis, which shows constant transconductance and voltage gain, and also reflects an efficiency dependency on the equivalent load resistance. The maximum efficiency difference between these two compensation circuits reached as high as 28% at 500 Ω and the disparity could increase at higher load resistances.

3.1. Introduction

The wireless power transfer (WPT) has been gaining popularity owing to the numerous advantages it has over its conventional wired counterpart. WPT offers convenience, reliability, safety, power isolation, and operation flexibility in hostile conditions. These benefits lead to the increased demand of WPT in areas such as medical implants [35], transportation [50], material handling systems [64], electric vehicles [65], [29], mobile devices [43], etcetera. In addition, figure-of-merits of WPT include power levels [66], efficiencies [67], gap distance variations [68] or misalignment tolerances [69] of the applications' system requirements. These challenges has fuelled the developments of the WPT systems in power electronics [70], [71], controls [30], [31], [72] and coil optimization [73], [74] areas.

The key segments of the WPT system are the transmitter (Tx) and receiver (Rx) coils. The mutual inductance (M) between these two coils allow electrical power to be wirelessly transferred. In order to increase the power transfer and to optimize its efficiency, compensating reactive components are added to both sides. A typical WPT system is shown in Fig. 3.1. The Tx side consists of a high frequency AC source, compensation circuit and the Tx coil. The Rx side contains the Rx coil, compensation circuit, rectifier and load. Most WPT designs use simple circuits such as series or parallel compensations, though more complex circuits have been used for specific purposes, such as regulating Tx current [65] or maximizing power transfer [75]. The coils and compensation circuits' equivalent output resistance, R_O is a function of the load resistance, R_L , which may varies. Practical examples of load variation are battery charging, motor drives, dynamic WPT systems with multiple loads, etcetera. According to the compensation circuit used, the overall circuit performance and efficiency will be affected by this variation.

Fig. 3.1 Typical WPT system

In [56], a detailed analysis of the four basic compensation circuit topologies (which are the combination of series and parallel connections of the compensation capacitors) is presented. The maximum power transfer capability is attained by reducing the reactive power flow which in turn can be achieved by variable frequency operation. However, frequency bifurcation incurs control complexity and may induce instability in the system. In addition, the efficiency analysis and comparison of the four compensation circuits were not discussed in [25]. According to [65], coupling variation affects the compensation capacitance required in the parallel compensation on the receiving side. Similar to [25], both efficiency analysis and comparison were not provided in [6]. In [76], secondary series and parallel compensation are analyzed for both their efficiency with load quality factor variation and voltage gain. However, the analyzed frequencies required are coupling dependent. Also, transconductance gain analysis was not included.

In [77], a dual topology is proposed for the receiving side. It comprises a series or a parallel compensated circuit that is to be matched to the corresponding

charging states of a battery. The efficiency of the system when using either the series or parallel compensation on the receiving side were however, not analyzed and compared in [26] and [77]. In [78], two fixed operating frequencies were proposed to realize the respective charging states using a series-series compensation, and an efficiency dip has been reported in the constant voltage charging state. Simulation results shown in [35] has illustrated the efficiency differences between the series, parallel and the proposed compensation method (on the receiving side), all with varying load resistances. In [79], comparison study has been done for series-series and double-sided LCC compensation topologies. Likewise, efficiency analysis and comparison on the load variation have not been done. Although both the series and parallel compensated Rx's have been widely deployed, these circuits have not been compared extensively, especially on their respective characteristics and efficiencies. The LCL compensation proposed in [80] reduced the VAR power flow and has load-independent output voltage and current at their respective frequencies [58]. However, the LCL compensation has been reported to suffer from current distortion [59], [62]. In order to reduce the distortion, this chapter proposes to replace the LCL compensation on the transmitting side with a series compensation (and receiving compensation is LCL) and will be named as S-LCL. The input currents of the S-LCL and LCL compensation circuits can be found in the Appendix. Also, as parallel compensated Tx requires current source (which requires additional energy storage components to transfer energy from a voltage source), this chapter focuses on series compensations which are suitable with voltage sources.

Therefore, this chapter analyses the series-series (S-S), series-parallel (S-P), and series-LCL (S-LCL) compensation circuits, to determine the respective efficiency optimal operating frequency. Using this efficiency-optimized frequency, the optimum efficiency range with respect to the equivalent output resistance will be analyzed. To further differentiate, the response of system resonance frequencies with respect to coupling are analyzed. The resonance

frequency response is essential for systems that employ soft switching for the front-end power electronic converter. Following which, the design flow chart for efficiency based on the above analyses will be detailed. Other notable analyses such as transferred power (TP), output current and voltage gains will also be covered. Lastly, the efficiency versus output resistance characteristic is demonstrated in a 2-coil WPT system, and the experimental results verified that uniform transconductance and voltage gains are achievable by the respective circuits.

3.2. Optimum Operating Frequency

In a typical WPT system, the Tx and Rx can be of different resonant frequencies. In this section, the focus is on the efficiency-optimized frequency where both reflected and induced source models [17,19] will be used for the general analysis of the compensation circuits.

3.2.1. Series Receiving Compensation (S-S)

The series receiving compensation includes S-S compensation circuit. For this compensation circuit, the reflected load and induced source equivalent circuits can be modelled as shown in Fig. 3.2. Firstly, the reflected impedance equations are shown as:

$$V_{in} = (Z_{1s} + Z_{refS}) i_{1ss} \quad (3.1)$$

$$\text{where } Z_{1s} = Z_{C1s} + Z_{L1} + R_1 \text{ and } Z_{refS} = \frac{(\omega M)^2}{(Z_{C2s} + Z_{L2} + R_2 + R_o)}$$

Fig. 3.2 S-S compensation circuit: (a) Reflected impedance and (b) induced source models for S-S compensation circuit

When evaluating efficiency we can consider R_1 and the reflected impedances of both circuits to be in series since the reactive components, C_{1p} and L_1 , do not consume real power. Therefore, only the real part of the reflected impedance in (3.1) is considered and overall system efficiency from the equivalent models of both compensation circuits can be shown as:

$$\eta = \frac{\Re(Z_{refS})}{\Re(Z_{refS}) + R_1} \times \frac{R_O}{R_O + R_2} \quad (3.2)$$

As we are interested in the compensation circuit, the focus is on the reflected impedance component, Z_{refS} . Z_{refS} has to be maximized while R_1 has to be minimized, in order to optimize efficiency and differential technique is applied to determine its maximum point with respect to frequency. The differential equation is solved to obtain the optimal frequency. Also, the terms with higher orders of C_{2s} ($C_{2s} \ll L_2 \ll R_2$) are neglected to form its approximate below.

$$\omega_{SS-eff} = \frac{i\sqrt{2}}{\sqrt{C_{2s}^2 R_2^2 + 2C_{2s}^2 R_2 R_O + C_{2s}^2 R_O^2 - 2L_2 C_{2s}}} \approx \frac{1}{\sqrt{L_2 C_2}} \quad (3.3)$$

It is clear from (3.3) that efficiency-optimized frequency is the resonance frequency of the Rx coil and the S-S compensation circuit.

3.2.2. Parallel Receiving Compensation (S-P and S-LCL)

Parallel receiving compensations include S-P and S-LCL. Similarly, only the real reflected impedance is considered and the overall system efficiency from the equivalent models of both compensation circuits can be shown as:

$$\eta = \frac{\Re(Z_{refP})}{\Re(Z_{refP}) + R_1} \times \frac{\Re(Z_{refP,1})}{\Re(Z_{refP,1}) + R_2} \quad (3.4)$$

$$\text{assuming } R_O \gg R_2 \text{ and } \Re(Z_{refP,1}) = \frac{R_O}{\omega^2 C_2^2 R_O^2 + 1} \gg R_2$$

Using the same circuit analysis technique in S-S compensation circuit, the efficiency-optimized frequency for S-P and S-LCL compensation circuits are derived. Also, when $R_O \gg R_2$ and the terms with higher orders of C_2 are neglected, its approximate can be formed in (3.5).

$$\omega_{SP/SLCL-eff} = \sqrt{\frac{L_2 \sigma + 2C_2 R_2^2 R_O^3 + C_2 R_2^3 R_O^2 + C_2 R_2 R_O^4 + C_2 R_2 R_O \sigma}{-C_2^3 R_2^3 R_O^4 + 2L_2 C_2^2 R_2 R_O^4 + L_2^2 C_2 R_O^3}} \approx \frac{1}{\sqrt{L_2 C_2}} \quad (3.5)$$

$$\text{where } \sigma = \sqrt{R_O^3 (R_2 + R_O)^3}$$

It is clear from (3.5) that the efficiency-optimized frequency is the resonance frequency of Rx coil for S-P and S-LCL compensation circuits. From (3.3) and (3.5), it is evident that the optimum operating frequency in terms of efficiency is closely tied to the resonance frequency of the Rx and its compensation circuit. Thus, the Rx resonance frequency will be used as the main operating frequency throughout this chapter and circuits will be tuned accordingly to this frequency.

3.3. Circuit Analysis and Design Flowchart

In this section, the compensation circuits are analyzed and their load-dependency efficiencies, crossover resistances, resonance frequency and design flowchart are derived accordingly.

3.3.1. Series Transmitting Compensations (S-S, S-P and S-LCL)

The S-S compensation circuit is commonly used due to its simplicity. Capacitors (C_{1s} and C_{2s}) are connected in series with the respective coils at both the transmitting and receiving sides. Assuming the coils' skin and proximity effects are included in the respective coil resistances (R_1 and R_2) and the coils are mutually coupled, the circuit equations for this system become:

$$V_{in} = i_{1ss} (Z_{C1s} + R_1 + Z_{L1}) - i_{2ss} (Z_M) \quad (3.6)$$

$$i_{1ss} Z_M = i_{2ss} (Z_{L2} + R_2 + Z_{C2s} + R_o) \quad (3.7)$$

where $Z_{C1s, C2s} = 1/(j\omega C_{1s, 2s})$, $Z_{L1, L2} = j\omega L_{1, 2}$ and $Z_M = j\omega M$.

Throughout the analysis, fundamental harmonic approximation (FHA) will be used to simplify the analysis. From (3.6) and (3.7), the simplified current equations at resonance frequency are derived as:

$$i_{1ss} = \frac{R_2 + R_o}{R_1 R_2 + R_1 R_o + \lambda_1 \lambda_2 k^2} V_{in} \quad (3.8)$$

$$i_{2ss} = \frac{j\omega_o M}{R_1 R_2 + R_1 R_o + \lambda_1 \lambda_2 k^2} V_{in} \quad (3.9)$$

where coupling coefficient $k = M/\sqrt{L_1 L_2}$, $\lambda_{1,2} = \sqrt{L_{1,2}/C_{1s, 2s}}$,

$$\omega_o = 1/\sqrt{L_{1,2} C_{1s, 2s}}$$

From (3.8) and (3.9), the efficiency of the S-S compensation circuit is derived as:

$$\eta_{s-s} = \frac{\lambda_1 \lambda_2 R_o k^2}{(R_2 + R_o)(\lambda_1 \lambda_2 k^2 + R_1 R_2 + R_1 R_o)} \quad (3.10)$$

which agrees with [81] and shows a non-linear relationship with R_o . By differentiating (3.10) with respect to R_o and using high quality coils, the maximum efficiency occurs approximately at:

$$R_{O(SS),\eta=\max} = \frac{\sqrt{R_1 R_2 (\lambda_1 \lambda_2 k^2 + R_1 R_2)}}{R_1} \approx \sqrt{\lambda_1 \lambda_2 k^2} \quad (3.11)$$

which shows that WPT systems with S-S compensation exhibit an efficiency that peaks around low load resistance, R_o for loosely coupled WPT systems (where $k < 0.2$). Besides the coupling coefficient, the peak efficiency load resistance ($R_{O(SS),\eta=\max}$) is also dependent on the ratio between coil (Tx or Rx) to compensation capacitance, which can be used as a design parameter for the WPT system to tune its $R_{O(SS),\eta=\max}$.

Beyond the maximum point where $R_o \gg R_{l,2}$, the efficiency becomes:

$$\eta_{s-s} \approx \frac{\lambda_1 \lambda_2 k^2}{\lambda_1 \lambda_2 k^2 + R_1 R_2 + R_1 R_o} \quad (3.12)$$

Table 3.1 shows the important equations and characteristics for the compensation circuits. The derivation for S-P and S-LCL can be found in the Appendix which follows the same approach.

Fig. 3.3 shows the various compensation circuits' efficiency against load resistance which shows that the efficiency-to-load-resistance characteristic is dictated by the type of compensation circuit on the Rx side (e.g. comparing between S-S and S-P compensation circuits). As the double-sided LCC compensation is a popular compensation circuit, it is included for comparison. However, due to the inherent higher order equations, analytical solutions are not practical. Instead, numerical solutions of its efficiency have been obtained using parameters from Table 3.3. Despite its similarities with the S-S compensation circuit in terms of efficiency-to-load-resistance characteristic, the efficiency of the LCC compensation circuit can be seen to fall behind S-S compensation circuit in Fig. 3.3. This can be attributed to the additional parasitic resistances from the higher component count. Another factor which would degrade efficiency further (not included in this analysis) is the self-couplings between the Tx (L_1 and L_2)

and Rx (L_3 and L_4) coil/inductor pairs. For these reasons, we will focus on S-S, S-P and S-LCL compensation circuits.

Using the respective compensation circuits' $R_{O,\eta=\max}$ and their relationships between the efficiency and R_O , the equivalent ohm load suitability can be determined. However, the suitability of each compensation circuit, for a chosen application can still be unclear. As a design criteria, the next sub-section will introduce the crossover resistance.

Table 3.1 Circuits efficiency analysis summary

	$R_{O,\eta=\max}$	η_{approx}
S-S	$\frac{\sqrt{R_1 R_2 (\lambda_1 \lambda_2 k^2 + R_1 R_2)}}{R_1}$	$\frac{\lambda_1 \lambda_2 k^2}{\lambda_1 \lambda_2 k^2 + R_1 R_2 + R_1 R_O}$
S-P	$\frac{\lambda_2 \sqrt{R_2 (\lambda_1 \lambda_2 k^2 + R_1 R_2) (\lambda_2^2 R_1 + \lambda_1 \lambda_2 R_2 k^2 + R_1 R_2^2)}}{\lambda_1 \lambda_2 R_2 k^2 + R_1 R_2^2}$	$\frac{\lambda_1 \lambda_2^3 k^2}{\lambda_1 \lambda_2^3 k^2 + \frac{\lambda_2^4 R_1 + \lambda_1 \lambda_2^3 R_2 k^2}{R_O}}$
S-LCL	$\frac{\lambda_2^2 \sqrt{R_1 R_2 (\lambda_1 \lambda_2 k^2 + R_1 R_2)}}{\lambda_1 \lambda_2 R_2 k + R_1 R_2^2}$	$\frac{\lambda_1 \lambda_2 k^2}{\lambda_1 \lambda_2 k^2 + 2 R_1 R_2 + \frac{\lambda_2^2 R_1}{R_O}}$

Fig. 3.3 Efficiency against load resistance for S-S, S-P, S-LCL and LCC compensated WPT system

3.3.2. Crossover Resistance

Having known the maximum efficiency load resistance for the respective compensation circuit, it can be simple to decide which circuit to use. However, if the equivalent load resistance falls in between two or more maximum points, it is unclear which circuit is able to yield better efficiency. As both S-P and S-LCL compensation circuits produce similar efficiencies, we define $R_{SS-SP/SLCL}$ as the crossover resistance from S-S to both S-P and S-LCL compensation circuits, at which the respective circuits' efficiencies intersect. By assuming $\lambda_{1,2} \gg R_{1,2}$, the crossover resistance between series (S-S) and parallel (S-P and S-LCL) receiving compensation circuits is derived from (3.12), (A3.5) and (A3.10):

$$R_{SS-SP/SLCL} = \frac{\sqrt{R_1 (4R_1 \lambda_2^2 + 4\lambda_1 \lambda_2 R_2 k^2 + R_1 R_2^2)} - R_1 R_2}{2R_1} \approx \lambda_2 \quad (3.13)$$

Essentially, the crossover resistance is significantly dependent on the receiving coil and compensation capacitor designs. With this, designers can make decisions on which compensation circuits to use, in order to optimize the efficiency of the WPT system. Next, as the efficiency of the S-P and S-LCL are shown to be similar in Fig. 3.3, the next sub-section will compare the resonance frequency of these two circuits.

3.3.3. Resonant Frequency Response (S-P and S-LCL)

It is well known that the frequency response of S-S and LCC are independent of the coupling coefficient [79]. On the other hand, while the S-P and S-LCL compensation circuits may yield similar efficiency, their resonant frequency response are shown in Fig. 3.4. For the case of S-P configuration, the overall resonant frequency response varies with the coupling level [53]. It is disadvantageous for the resonance frequency to vary especially when the power converter employs soft switching techniques such as Zero-Voltage Switching

(ZVS), where an inductive impedance is required. Therefore, frequency tracking is required in S-P compensation circuit in order to maintain ZVS operation. Thus, even if the efficiencies of the S-P and S-LCL are similar, the highly dynamic resonance frequency of the S-P compensation circuit poses a significant disadvantage for WPT systems utilizing single frequency operation.

On the other hand, the resonant frequency of S-LCL compensation circuit does not change, when $L_3 = L_2$. While it is practically impossible to match the inductances, when L_3 is 110% of L_2 , the S-LCL resonant frequency decreases slightly as compared to the S-P circuit when coupling increases. This provides the designer with a control parameter to maintain ZVS. To conclude the efficiency analysis and comparison of the compensation circuits, the design flowchart will be presented next.

Fig. 3.4 Normalized resonant frequency of S-P and S-LCL (of various L_3 values) compensation circuits

3.3.4. Compensation Circuit Design Flowchart

The compensation circuit design flowchart shown in Fig. 3.5, is constructed from the analyses presented in this section. Firstly, the designer has to define the circuit parameters and required operating conditions. Secondly, the flowchart prioritizes efficiency as it determines which compensation circuits are best suited for the

application by the crossover resistance. Additionally, if regulated resonance frequency is required, S-LCL is selected as its additional inductor L_3 , mitigates the effects from coupling variation.

Next, other notable analysis such as the transferred power and the voltage and current gains will be presented.

Fig. 3.5 Compensation circuit design flowchart

3.4. Transferred Power and Voltage and Current Gains

Other notable analysis are also included in this section. Here, TP is defined as normalizing output power to the input voltage, V_{in} .

3.4.1. Transferred Power

For S-S compensation circuit, TP can be obtained from (3.9):

$$TP_{ss} = \left| \frac{k\sqrt{\lambda_1\lambda_2}}{R_1R_2 + R_1R_o + \lambda_1\lambda_2k^2} \right|^2 R_o \quad (3.14)$$

which shows a non-linear relationship with R_o . By differentiating (3.14) with respect to R_o , the maximum TP point occurs at:

$$R_{O(SS),TP=\max} \approx \frac{\lambda_1\lambda_2k^2}{R_1} \quad (15)$$

Another point of interest is the amount of TP available when the load resistance is efficiency-optimized as according to (3.11). This can be derived by substituting (3.11) into (3.14) and assuming $\lambda_{1,2} \gg R_{1,2}$:

$$TP_{SS(\eta=\max)} \approx \frac{R_2}{2R_1R_2 + k\sqrt{\lambda_1\lambda_2R_1R_2}} \quad (16)$$

Extending the above circuit analysis, the TP maximum point and amount of TP at optimized efficiency points are derived for other compensation circuits discussed in Table 3.2. The TP equations provide information on the amount of power each circuit is able to provide per unit volt.

Table 3.2 Circuits transferred power analysis summary

	$R_{O,TP=\max}$	$TP_{\eta=\max}$
S-S	$\frac{\lambda_1\lambda_2k^2}{R_1}$	$\frac{R_2}{2R_1R_2 + k\sqrt{\lambda_1\lambda_2R_1R_2}}$
S-P	$\frac{\sqrt{\lambda_1^2\lambda_2^2k^4 + 2\lambda_1\lambda_2R_1R_2k^2 + \lambda_2^2R_1^2}}{\lambda_1k^2}$	$\frac{2R_1R_2 + k\sqrt{\lambda_1\lambda_2R_1R_2}}{4R_1k\sqrt{\lambda_1\lambda_2R_1R_2} + \lambda_1\lambda_2R_1k^2 + 2\lambda_1^2R_2k^4}$
S-LCL	$\frac{\lambda_2R_1}{\lambda_1k^2}$	$\frac{k\sqrt{\lambda_1\lambda_2R_1R_2}}{2R_1k\sqrt{\lambda_1\lambda_2R_1R_2} + \lambda_1\lambda_2R_1k^2}$

3.4.2. Transconductance and Voltage Gains

Due to the need for frequency tracking in the S-P compensation circuit, the gains of S-S, and S-LCL compensation circuits are presented. Apart from the efficiency performance disparity, output gains differ between these two compensation circuits as well.

In certain applications, it is essential to have a constant source of voltage or current even as the load varies. Most applications are dependent on a constant voltage source, while some applications such as motor drives and battery charging require a constant current source.

From (3.6) and (3.7), the transconductance and voltage gain of the S-S compensation circuit is derived:

$$\left| \frac{i_2}{V_{in}} \right| = \left| \frac{Z_M}{(Z_S + R_L)Z_P - Z_M^2} \right| \quad (3.17)$$

$$\left| \frac{V_O}{V_{in}} \right| = \left| \frac{Z_M R_L}{(Z_S Z_P - Z_M^2) + Z_P R_L} \right| \quad (3.18)$$

where $Z_M = j\omega M$, $Z_{P,S} = Z_{C1s,2s} + Z_{L1,2} + R_{1,2}$ and $V_O = i_2 R_O$.

In order to preserve efficiency, selected load resistances within the optimal efficiency region of the S-S compensation circuit (5, 50 and 100 Ω) are used to illustrate transconductance and voltage gains against frequency (normalized by its resonance frequency) in Fig. 3.6 (a) and Fig. 3.6 (b) respectively. It can be clearly seen that there is near uniform transconductance and voltage gain at the resonance (ω_o) and bifurcation frequencies respectively. It is relatively simple to obtain uniform transconductance as resonance frequency can be easily determined by circuit parameters [79]. However for uniform voltage gain, bifurcation frequencies have to be tracked as they are dependent on the coupling coefficient [56].

Similarly, the transconductance and voltage gain of the S-LCL compensation circuit are derived as:

$$\left| \frac{i_4}{V_{in}} \right| = \left| \frac{Z_M Z_{C2}}{(Z_3 + R_L)(Z_P Z_S - Z_M^2) - Z_P Z_{C2}^2} \right| \quad (3.19)$$

$$\left| \frac{V_O}{V_{in}} \right| = \left| \frac{Z_M Z_{C2} R_L}{(Z_3 + R_L)(Z_P Z_S - Z_M^2) - Z_P Z_{C2}^2} \right| \quad (3.20)$$

where $Z_3 = Z_{C2} + Z_{L3} + R_3$.

Likewise, to preserve efficiency, selected load resistances within the optimal efficiency region of the S-LCL compensation circuit (200, 300 and 400 Ω) are used to illustrate transconductance and voltage gains against frequency (normalized by its resonance frequency) in Fig. 3.7 (a) and Fig. 3.7 (b) respectively, whereas 4000 Ω acts as reference points to show the bifurcation in Fig. 3.7 (a). It can be clearly seen that there is near uniform transconductance and voltage gain at the bifurcation and resonance frequencies respectively. Contrary to the S-S compensation circuit, it is not possible to employ resonance frequency operation to obtain uniform transconductance. However, it is possible to obtain a near uniform voltage gain through simple resonance frequency operation at ω_o . After concluding the analysis on the compensation circuits with the design flowchart and other notable analyses, the next section will be on the experimental verification. According to the flowchart for regulated resonance frequency response, S-S and S-LCL compensation circuits are selected for the discussion in the next section.

Fig. 3.6 (a) Transconductance and (b) voltage gain of S-S compensation circuit

Fig. 3.7 (a) transconductance and (b) voltage gain of S-LCL compensation circuit

3.5. Experimental Verification

The compensation circuits and coils are constructed to verify the analytical results. The experiment is carried out using Keysight E5061B ENA Series Network Analyzer and Advanced Design System (ADS) [82], and the experimental setup is shown in Fig. 3.8 and the experiment parameters are listed in Table 3.3.

Selected load resistances within the optimal efficiency region are used to obtain the experimental efficiency results for S-S and S-LCL. The S-S and S-LCL efficiency results are provided in Fig. 3.10 and fit well with the analytical results. According to (3.13), the crossover resistance is approximately equals to 172.7Ω

which agrees well with the efficiency results shown in Fig. 3.10. Next, the experimental transconductance of S-S compensation circuit and the voltage gain of S-LCL compensation circuit are selected based on their uniformity response against varying load. The S-S transconductance and S-LCL voltage gain results are given in Fig. 3.9 (a) and (b) respectively, which show similar gains throughout but has more discrepancies compared to the analytical results presented in Section IV. This can be attributed to the increased parasitic resistance from the cables, connectors and coil designs, which agrees with (3.17) to (3.20) on the effect of parasitic resistances.

Fig. 3.8 Experimental hardware setup

Table 3.3 Circuits parameters

Parameter	Simulation	Experiment
Tx and Rx coil inductance, L_1, L_2 (μH)	31.6	31.6
Coil Resistances R_1 and R_2 at 870 kHz (Ω)	0.3	0.3
Coils' radius (cm)	-	8
Rx series inductor, L_3 (μH) (S-LCL)	31.6	32
Tx and Rx side capacitors, C_1, C_2, C_3 and C_4 (LCC) (nF)	0.405	0.405
Tx and Rx series capacitor, C_{1s}, C_{2s} (nF)	1.059	1.06
Tx parallel capacitor, C_{2p} (nF)	1.059	1.06
Gap distance (cm)	-	8
Operating frequency, f_o (kHz)	870	
Coupling coefficient, k	0.1	
Load Resistances, R_o (Ω)	5 to 500

(a)

(b)

Fig. 3.9 (a) Experimental transconductance of S-S compensation circuit and (b) experimental voltage gain of S-LCL compensation circuit

Fig. 3.10 Experimental efficiency of S-S and S-LCL compensated WPT system

3.6. Conclusion

A detailed efficiency analysis of the key compensation circuit types, i.e. S-S, S-P and S-LCL, mainly with respect to the load variation, has been presented. Firstly, efficiency-optimized frequencies were derived for these various compensation circuits. With this frequency assumption and from the efficiency analysis, it can be seen that the receiving compensation type affects the efficiency characteristics. In addition, crossover resistance was introduced to assist the designers to further optimize their efficiency by choosing the appropriate compensation circuit according to the load requirements. While S-P and S-LCL yielded similar efficiency, their resonant frequency responses against varying coupling levels differ, and for S-LCL the additional inductor allows the designer to plan accordingly. To conclude the efficiency analyses, a compensation circuit design flowchart is constructed based on the efficiency optimizations.

Secondly, the TP of the compensation circuits was derived. Specifically, the load resistance at which maximum TP occurs and the TP at the efficiency-optimized load resistances were investigated. Regarding the output gains, only the S-S and S-LCL (S-P compensation circuit is dependent on load and coupling levels) compensation circuits are analyzed for their voltage and transconductance gains. Simple resonance frequency operation can be employed to obtain uniform

transconductance and voltage gains from S-S's and S-LCL compensation circuit respectively. However, complex frequency tracking is required to obtain uniform voltage gain and transconductance from S-S and S-LCL compensation circuits respectively.

Lastly, analytical efficiency characteristic and crossover resistance have been verified by the experimental setup. Also, experimental examination of the transconductance and voltage gain for the S-S and S-LCL compensation circuits, respectively, had been validated to be constant at its resonance frequency. In practice these two constant gains were affected by parasitic resistances in the system, which should be minimized by appropriate engineering efforts in the coil design.

By analyzing the optimal load conditions, WPT systems can be better designed to suit their applications (i.e. low or high powered applications) in terms of efficiency and TP. In addition to simple resonance frequency operation, applications requiring constant transconductance or voltage gain can employ the appropriate compensation circuit.

3.7. Appendix

Using MATLAB/Simulink and square-wave unit voltage input source, input currents of S-LCL and LCL compensation circuits are shown in Fig. A1. The comparison clearly shows the highly distortion of the LCL compensation circuit.

Fig. A1 Input currents of S-LCL (top) and LCL (bottom) compensation circuit
The efficiency of the S-P and S-LCL compensation circuits are derived in this appendix. The equations of the S-P compensation circuit are:

$$\begin{aligned}
 V_{in} &= i_{1sp}(Z_{C1s} + R_1 + Z_{L1}) - i_{2sp}(Z_M) \\
 i_{1sp}Z_M &= i_{2sp}(Z_{L2} + R_2 + Z_{C2s}) - i_{3sp}Z_{C2s} \\
 i_{1sp}Z_M &= i_{2sp}(Z_{L2} + R_2) + i_{3sp}R_O
 \end{aligned} \tag{A3.1}$$

The input (i_{1SP}) and output currents (i_{3SP}) equations at resonant frequency are:

$$\begin{aligned}
 i_{1SP} &= \frac{(\lambda_2^2 + R_2R_O - j\lambda_2R_2)V_{in}}{\lambda_1\lambda_2R_Ok^2 + \lambda_2^2R_1 + R_1R_2R_O - j(\lambda_1\lambda_2^2k^2 + \lambda_2R_1R_2)} \\
 i_{3SP} &= \frac{-\omega_o M \lambda_2 V_{in}}{\lambda_1\lambda_2R_Ok^2 + \lambda_2^2R_1 + R_1R_2R_O - j(\lambda_1\lambda_2^2k^2 + \lambda_2R_1R_2)}
 \end{aligned} \tag{A3.2}$$

Using (A3.1) and (A3.2), the efficiency can be derived as:

$$\eta_{S-P} = \frac{\lambda_1 \lambda_2^3 R_O k^2}{\lambda_2^4 R_1 + \lambda_1 \lambda_2^3 k^2 (R_O + R_2) + 2\lambda_2^2 R_1 R_2 R_O + \lambda_1 \lambda_2 R_2 R_O^2 k^2 + R_1 R_2^2 (R_O^2 + \lambda_2^2)} \quad (\text{A3.3})$$

(A3.3) shows a non-linear relationship with R_O and using differentiation technique, the maximum efficiency can be shown to approximately occur at:

$$R_{O(SP), \eta=\max} = \frac{\lambda_2 \sqrt{R_2 (\lambda_1 \lambda_2 k^2 + R_1 R_2) (\lambda_2^2 R_1 + \lambda_1 \lambda_2 R_2 k^2 + R_1 R_2^2)}}{\lambda_1 \lambda_2 R_2 k^2 + R_1 R_2^2} \quad (\text{A3.4})$$

Before the maximum point, the efficiency is approximated by removing lower orders of $\lambda_{1,2}$:

$$\eta_{S-P} \approx \frac{\lambda_1 \lambda_2^3 k^2}{\lambda_1 \lambda_2^3 k^2 + \frac{\lambda_2^4 R_1 + \lambda_1 \lambda_2^3 R_2 k^2}{R_O}} \quad (\text{A3.5})$$

Next, equations of the S-LCL compensation circuit are:

$$\begin{aligned} V_{in} &= i_{1slcl} (Z_{C1s} + R_1 + Z_{L1}) - i_{2slcl} (Z_M) \\ i_{1slcl} Z_M &= i_{2slcl} (Z_{L2} + R_2 + Z_{C2s}) - i_{3sp} Z_{C2s} \\ i_{1slcl} Z_M &= i_{2slcl} (Z_{L2} + R_2) + i_{3slcl} (Z_{L3} + R_3 + R_O) \end{aligned} \quad (\text{A3.6})$$

The input (i_{1SLCL}) and output currents (i_{4SLCL}) equations at resonant frequency are:

$$\begin{aligned} i_{1SLCL} &= \frac{(\lambda_2^2 + R_2 R_O - j\lambda_2 R_2 \sigma_1) V_{in}}{\lambda_1 \lambda_2 R_O k^2 + \lambda_2^2 R_1 + R_1 R_2 R_O + j\omega_o \lambda_1 \lambda_2 k^2 \sigma_2 - j\lambda_2 R_1 R_2 \sigma_1} \\ i_{4SLCL} &= \frac{\omega_o M \lambda_2 V_{in}}{\lambda_1 \lambda_2 R_O k^2 + \lambda_2^2 R_1 + R_1 R_2 R_O + j\omega_o \lambda_1 \lambda_2 k^2 \sigma_2 - j\lambda_2 R_1 R_2 \sigma_1} \end{aligned} \quad (\text{A3.7})$$

where $\sigma_1 = 1 - \omega_o^2 C_2 L_3$ and $\sigma_2 = L_3 - L_2$.

From currents' equation (A3.7), we can see the difference between the S-P and S-LCL compensation circuits. By tuning the additional inductor, L_3 to be of the equal value with Rx coil, L_2 and also to resonate with the Rx compensation capacitor, C_2 , the imaginary components can be eliminated. Using (A3.6) and (A3.7), the efficiency can be derived as:

$$\eta_{S-LCL} = \frac{\lambda_1 \lambda_2^3 R_o k^2}{(\lambda_2^2 + R_2 R_o) (\lambda_2^2 R_1 + \lambda_1 \lambda_2 R_o k^2 + R_1 R_2 R_o)} \quad (\text{A3.8})$$

(A3.8) shows a non-linear relationship with R_o and using differentiation technique, the maximum efficiency occurs at:

$$R_{O(SLCL), \eta=\max} = \frac{\lambda_2^2 \sqrt{R_1 R_2 (\lambda_1 \lambda_2 k^2 + R_1 R_2)}}{\lambda_1 \lambda_2 R_2 k + R_1 R_2^2} \quad (\text{A3.9})$$

Before the maximum point, efficiency is approximated by removing lower orders of $\lambda_{1,2}$:

$$\eta_{S-LCL} \approx \frac{\lambda_1 \lambda_2 k^2}{\lambda_1 \lambda_2 k^2 + 2R_1 R_2 + \frac{\lambda_2^2 R_1}{R_o}} \quad (\text{A3.10})$$

Chapter 4

LCC Compensated WPT System

Typical wireless power transfer systems utilize series compensation circuit which is based on magnetic coupling and resonance principles that was first developed by Tesla. However, changes in coupling caused by gap distance, alignment and orientation variations can lead to reduce power transfer efficiencies and the transferred power levels. This chapter proposes impedance matched circuit to reduce frequency bifurcation effect and improve on the transferred power level, efficiency and total harmonic distortion (THD) performance of the series compensation circuit. A comprehensive mathematical analysis is performed for both series and impedance matched circuits to show the frequency bifurcation effects in terms of input impedance, variations in transferred power levels and efficiencies. Matlab/Simulink results validate the theoretical analysis and shows the circuits' THD performance when circuits are fed with power electronic converters.

4.1. Introduction

Wires have served as the primary medium for the transmission of electricity since its discovery in the 18th century. Since then, the simple wire has seen the evolution of numerous types of wires to suit different applications. The revolution of electricity transmission is wireless power transfer (WPT), which is basically the transmission of electricity without physical conductor. In 2007, MIT has wirelessly transferred 60 W with 40% efficiency over 2 meters [14].

WPT offers numerous advantages over its wired counterparts in terms of convenience, reliability, safety, ruggedness and flexibility. These benefits lead to an increased demand of WPT in areas such as transportation [83], robots [84], electric vehicles [85], mobile devices [43], medical implants [86], etcetera. In addition, the success of implementing WPT in such applications is determined by the power levels [87], efficiencies [88], gap distance variations or misalignment tolerances [89] of the system's requirements. These challenges give rise to advanced developments in WPT systems in power electronics [90], controls [31] and coil optimization [91] areas.

A typical WPT circuit is shown in Fig. 4.1. The circuit typically consists of a high frequency AC source, compensating series capacitor and the transmitting coil on the primary side. Mutual inductance (denoted as M_s) allows the two coils, L_{1s} and L_{2s} to be coupled and power can be wirelessly transferred from the primary to secondary side. The series compensating capacitors allow resonance operation which improves the overall efficiency of the WPT system.

4.2. Proposed Impedance Matched Circuit

Inductive coils are the primary means to transfer power in WPT circuits. When excited individually by a high frequency source, these coils draw inductive currents and increases the volt-ampere (VA) ratings of the source. Therefore, capacitors are usually added to compensate the inductive reactance in the form of series-series, series-parallel and parallel-parallel (the former being on the transmitting end and the latter on the receiving end). However, series-series

(hereafter named as series) compensation is generally accepted as the required compensation capacitor is independent of the coupling level [56]. The impedance response of series resonant circuit can be seen in Fig. 4.3 (a).

However, series compensation circuits face limitations in WPT when the distance, orientation and alignment of the coils are dynamically changing. Reference [92] shows the frequency bifurcation effect on output power when coupling between the coils increases beyond the critical coupling point. One solution is to employ adaptive tuning [85], increasing complexity to the overall WPT system.

In power electronic applications, lower impedances are desired to improve circuit power levels and efficiencies [93]. To lower the impedance and reduce the frequency bifurcation effect of the series compensation circuit, this chapter proposes an impedance matched circuit where a parallel resonant tank is added to the series compensation (hereafter named as LCC) circuit as shown in Fig. 4.2. The impedance of the series compensation circuit is matched with the parallel resonant tank, resulting in one resonant impedance trough and peak within the range of interest as seen in Fig. 4.3 (b).

Fig. 4.1 Typical series compensated WPT system

Fig. 4.2 Parallel resonant tank added to series compensation to form LCC

Fig. 4.3 (a) Impedances of inductive coil, capacitor and series resonant circuits.
 (b) Impedances of series and parallel resonant, and LCC circuits

4.3. Circuit Analysis

4.3.1. Resonance Frequency

In the series compensated circuit, the capacitor compensates the inductive coil, allowing resonance frequency operation and its frequency, $f_{o(series)}$ is shown in (6.1). On the other hand, in the LCC circuit, the series capacitor C_2 does not fully compensate the inductive coil but the overall inductive impedance (of C_2 and L_2) is matched with L_1 and C_1 in order to allow resonance frequency operation. The LCC circuit can be seen in Fig. 4.4. The simplified ($L = L_{1s,2s,1,2,3,4}$ and $C = C_{1s,2s,1,2,3,4}$) resonance frequency $f_{o(LCC)}$ is shown in (6.2).

$$f_{o(series)} = \frac{1}{2\pi\sqrt{LC}} \quad (6.1)$$

$$f_{o(LCCL)} = \frac{1}{2\pi} \sqrt{\frac{3-\sqrt{5}}{2LC}} \quad (6.2)$$

4.3.2. Equivalent Circuits

The equivalent circuit of the series compensated WPT system can be seen in Fig. 4.5. The series compensation's mutual inductance, M_s is different from the other physical passive devices in the circuit. It is dependent on the coupling coefficient, k which depends on coil separation, orientation and alignment.

$$M = k\sqrt{L_x L_y}, \text{ where } x \text{ and } y \text{ are the coils} \quad (6.3)$$

First, the output current equation of the series compensation equivalent circuit is shown in (6.4) and is used to derive the input impedance in (6.5).

$$i_{2s} = \frac{Z_{Ms}}{Z_{C2s} + Z_{L2s} + R_{2s} + R_{Ls}} i_{1s} \quad (6.4)$$

$$Z_{ins} = Z_{C1s} + Z_{L1s} + R_{1s} - \frac{Z_{Ms}^2}{Z_{C2s} + Z_{L2s} + R_{2s} + R_{Ls}} \quad (6.5)$$

Next, the equivalent circuit of the LCC WPT system can be seen in Fig. 4.6. Firstly, the current equations of the equivalent circuit are given in (6.6) – (6.8). Resistances of the inductors L_1 and L_4 which are negligible compared to L_2 and

L_3 as the wire lengths are insignificant when using high frequency ferrite cores. Lastly, the input impedance of the LCC circuit can be derived as (6.9) from (6.6) – (6.8).

$$i_6 = \frac{Z_{C4}}{Z_{C4} + Z_{L4} + R_L} i_4 \quad (6.6)$$

$$i_4 = \frac{Z_M}{Z_{C3} + Z_{C4} + Z_{L3} + R_3 - \frac{Z_{C4}^2}{Z_{C4} + Z_{L4} + R_L}} i_2 \quad (6.7)$$

$$i_2 = \frac{Z_{C1}}{Z_{C1} + Z_{C2} + Z_{L2} + R_2 - \frac{Z_M^2}{Z_{C3} + Z_{C4} + Z_{L3} + R_3 - \frac{Z_{C4}^2}{Z_{C4} + Z_{L4} + R_L}}} i_1 \quad (6.8)$$

Fig. 4.4 Proposed LCC compensation circuit

Fig. 4.5 Series compensation equivalent circuit

Fig. 4.6 LCC equivalent circuit

$$Z_{in} = Z_{C1} + Z_{L1} - \frac{Z_{C1}^2}{Z_{C1} + Z_{C2} + Z_{L2} + R_2 - \frac{Z_M^2}{Z_{C3} + Z_{C4} + Z_{L3} + R_3 - \frac{Z_{C3}^2}{Z_{C4} + Z_{L4} + R_L}}} \quad (6.9)$$

4.3.3. Impedance Frequency Response

The analyses of series compensation and LCC circuit are performed using the simulation parameters from Table 4.1 and equations (6.5) and (6.9) respectively. Fig. 4.7 and Fig. 4.8 show the impedance frequency response analysis of the WPT system with series compensation and LCC respectively.

The series compensated and LCC circuits' input impedance Bode plots are compared and shown in Fig. 4.9. The resonance frequencies of the series compensated circuit as identified from (6.1), apply to the second zero degrees phase crossing and the input impedance of the series compensated circuit can be seen to be significantly higher than that of the LCC circuit. These two observations are due to the frequency bifurcation effect [92] (i.e. the coils become overcoupled). This effect is similar to the trough splitting found in Fig. 4.9 as impedance is considered. In the Bode plots, the location of the zeroes of (6.5) and (6.9) affects the input impedances. For example in Fig. 4.9, the location of the zeroes of the series compensated circuit is further apart than that of the LCC circuit, resulting in higher impedance peak.

For lower coupling coefficients ($k = 0.01$), the LCC circuit exhibits lower input impedance compared to the series circuit as shown in Fig. 4.10. Even though there is no distinct appearance of frequency bifurcation effect (when $k = 0.01$), the series compensation input impedance is increasing exponentially, signifying the onset of the frequency bifurcation effect. Fig. 4.11 shows the frequency of zeroes and the input impedances of the series compensated and LCC circuits with respect to coupling coefficient.

Table 4.1 Simulation parameters

Series		LCC	
Parameters	Value	Parameters	Value
L_{1s}, L_{2s}	$25.33 \mu\text{H}$	L_1, L_2, L_3, L_4	$25.33 \mu\text{H}$
C_{1s}, C_{2s}	2.618 nF	C_1, C_2, C_3, C_4	1 nF
R_{1s}, R_{2s}	0.5Ω	R_2, R_3	0.5Ω
R_{Ls}	4Ω	R_L	4Ω
V_1	$100\sin(\omega t)$	V_1	$100\sin(\omega t)$

Fig. 4.7 Series compensation impedance response at $k = 0.1$ Fig. 4.8 LCC circuit impedance response at $k = 0.1$

Fig. 4.9 Input impedance comparison of series compensated and LCC circuits, with zeroes' locations circled at $k = 0.1$

Fig. 4.10 Input impedance comparison of series compensated and LCC circuits at $k = 0.01$

Fig. 4.11 Normalized zeroes frequencies and input impedances for series compensated and LCC circuits with respect to coupling coefficient (k)

The frequencies of the zeroes are normalized to the respective circuit resonance frequency calculated using (6.1) and (6.2) while the input impedances are the absolute results of (6.5) and (6.9). As the frequency bifurcation effect occurs at a higher coupling for the LCC circuit, lower input impedances across the range of coupling coefficients are observed.

4.3.4. Transferred Power and Efficiency

Transferred power level and efficiency are two important parameters for WPT. Input impedance affects both the efficiency and transferred power. Transferred power can be derived by using (6.6) – (6.9). Fig. 4.12 shows the transferred power and efficiencies with respect to coupling coefficients for both circuits at their respective resonant frequencies. Higher input impedance reduces the input power to the coil which could result in lower transferred power. As coils affect the power transfer and efficiency, this analysis does not consider any optimization of the coil and keeps the both coil elements identical.

The maximum power transfer points for the two circuits can be obtained from Fig. 4.12, where the vertical lines mark the respective critical coupling points. The efficiencies at these points are below 50% as the theoretical maximum efficiency for matched load and source condition is 50%. Due to the lower input impedance

of the LCC circuit at load-source matched conditions, it has a significantly higher level of transferred maximum power. Fig. 4.13 shows the simulation results of both circuits using SimPowerSystems, Simulink®, which agrees with the previous theoretical analysis.

When operating at higher coupling levels, both circuits produce high power transfer efficiencies. However, the LCC circuit is able to produce relatively higher efficiency and transferred power levels compared to series compensation circuit at the given circuit parameters. Simulation results shown in Fig. 4.14 agree with the previous theoretical analysis.

Fig. 4.12 Transferred powers and efficiencies with respect to coupling coefficient (k) from 0.001 to 0.2

Fig. 4.13 Simulation results for input (top) and transferred (bottom) powers for series compensation and LCC circuits at maximum power transfer points

Fig. 4.14 Simulation results for input (top) and transferred (bottom) powers of series compensation and LCC circuits at $k = 0.2$

4.3.5. Total Harmonic Distortion

Power electronic converters function as inputs to these WPT systems. These converters tend to produce square wave voltages (consisting of high frequency components as shown in (6.10), which will be used in the following analysis) instead of sinusoidal wave voltages. Subsequently, the input current THD levels can become significant especially when coupling coefficient is increased as shown in Fig. 4.15. Other simulation parameters follow Table 4.1.

$$V_1 = 100 \frac{4}{\pi} \sum_{n=1,3,\dots}^{\infty} \frac{1}{n} \sin(n\omega_o t) \quad (6.10)$$

For coupling coefficients 0.01 and lower, THD is low and insignificant and are ignored in this part of the analysis. At coupling factor $k = 0.2$, the effect of THD can be seen clearly in the series compensation's distorted input current shown in Fig. 4.16. In contrast, the improved THD levels of the input current in the LCC circuit reduces harmonic losses caused in the compensation circuits, which reduces the input real power fluctuations as shown in Fig. 4.17. The reactive power circulation in series compensation resulted in fluctuating input real and transferred powers as shown in Fig. 4.18, which consequently affects its power efficiencies shown in Fig. 4.19. The efficiency comparison of square and

sinusoidal wave sources, highlight the susceptibility of the series compensation to increased THD levels. The peak voltage of the sinusoidal source is 100V. As the LCC circuit reduces THD levels significantly at higher coupling levels, relatively insignificant reactive power flow from the source is experienced and both its efficiency and transferred power levels can be maintained as shown in Fig. 4.19. Thus, the LCC circuit is able to suppress harmonics at higher coupling coefficients, and does not require additional reactive power compensation.

Fig. 4.15 Total harmonic distortion of series compensated circuit and LCC circuit with respect to coupling coefficient (k)

Fig. 4.16 Input currents of series compensated (top) and LCC (bottom) circuits at $k = 0.2$.

Fig. 4.17 Input apparent (top) and real powers (bottom) of the series compensated (left) and LCC (right) circuits at $k = 0.2$

Fig. 4.18 Input real (top) and transferred (bottom) powers of the series compensated (left) and LCC (right) circuits at $k = 0.2$

Fig. 4.19 Efficiencies of series compensated (top) and LCC (bottom) circuits at $k = 0.2$ for sinusoidal and square waves voltage inputs

4.4. Conclusion

LCC circuit has been proposed for WPT systems. The proposed circuit produces higher maximum transferrable power compared to series compensation circuit. In addition, the proposed circuit produces higher efficiencies with higher transfer power levels at higher coupling coefficients.

The LCC circuit allows WPT system designers to have higher maximum transferred power and efficiency when the overall efficiency of the system is not paramount, e.g. low powered devices. On the other hand, in high power situations where efficiency is critical (e.g. electric vehicle charging and transportation system operations.), the proposed circuit allows relatively higher transferred power levels and efficiencies at higher coupling levels.

When used with power electronic converters, the THD performance becomes a concern. At significant levels of coupling, THD level of the series compensation is increased and efficiency fluctuations occur. However, LCC compensation is able to lower the input current THD level and reactive power flow, effectively maintaining stable levels of efficiency.

Chapter 5

Dynamic Wireless Power Transfer

Wireless power transfer (WPT) is a convenient, flexible and safe alternative to its wired counterpart. Due to these benefits, WPT has seen rapid growth in recent years. Typical WPT systems are usually used for static applications where the load operates within a pre-defined area. On the other hand, dynamic WPT (D-WPT) are usually used to power loads that change its position dynamically and one way to achieve this is by increasing the overall charging area via multiple transmitters (Tx). The multiple Tx requires communication and control which increases the overall complexity of the system. This chapter proposes a control algorithm which aims to maintain the overall efficiency of the D-WPT system by turning on and off the Tx when the mobile receiver (Rx) approaches and departs, respectively. Transient analysis of the D-WPT system is used to derive its control variables. In addition, energy and power loss of this proposed control algorithm is also investigated. The feasibility of the proposed control algorithm is demonstrated with 1-Tx and 3-Tx experimental setups. The experimental results validates the theoretical analysis model and the proposed control algorithm, and the standby (when not charging) input power forms 7.6% of the operational input power.

5.1. Introduction

Wireless power transfer (WPT) has numerous advantages over its conventional wired counterpart which include convenience, reliability, safety, power isolation, and ease of operation in hostile conditions. These benefits lead to the implementation of WPT in areas such as medical implants [1]–[3], transportation [4]–[6], material handling systems [7], electric vehicles [8]–[10], mobile devices [11]–[13], etcetera. However, the WPT systems have two key limitations: charging range and orientation. The dynamic WPT (D-WPT) systems therefore aim to solve these issues by extending the coverage for the mobile receiver to be wirelessly charged within its range of motion. D-WPT can be applied in the above mentioned WPT applications and has particularly gained traction in dynamic EV charging [14]–[16]. There are two main types of Tx magnetic coupler structures. The first structure type is track based where the magnetic track spans the charging area, and the receiver (Rx) is charged when it moves along the long magnetic track [4]. The simple design and minimal component count allows multiple pickups along the track. While the track is able to create a substantial charging space, the relatively larger Tx track meant that the Rx merely covers a small portion of the Tx, leading to poor coupling levels. This, in turn, causes low efficiencies and electromagnetic interference (EMI) issues. The design proposed in [5] mitigated these effects by segmenting the Tx track into segments. Not only are the coupling levels improved, the electromagnetic fields of inactive segments are also reduced significantly. These segments are powered by a single inverter that relies on sensors to turn on or off the segments, and this causes the complexity of the D-WPT system to increase. Another structure type that is segment-based where the individualized magnetic couplers spans the charging area and allows each magnetic coupler to be powered by an independent power source [14], [17], [18]. Fig. 5.1 shows the circuit of a typical segmented D-WPT system where there are n number of Tx and one mobile Rx. By individually controlling each magnetic coupler and its power source, a larger portion of the electromagnetic field created by the Tx can be captured by the Rx. This would

increase the coupling level between a particular Tx and Rx, to boost its efficiency. The disadvantages here are the higher component and power electronics count. To add on, the overall efficiency and received power of the segmental systems are also extremely susceptible to misalignment, as the efficiency of the D-WPT system in [19] is reported to have varied from 85% to 5%. In order to smooth out the power flow during the dynamic charging process, [16] proposed to implement electrochemical capacitors to level out the currents on the transmitting and receiving sides while traversing from one Tx coil to another. Similar to the segmental track systems, sensors are needed to control the individual magnetic couplers and power sources for higher efficiencies, which again increase the system complexity. In order to reduce the required power converters of the segmental approach, [15] employed multiple LCC compensated Tx's that are connected to a single power converter. This is possible as the input power is automatically distributed among the LCC compensated Tx's according to its respective coupling level. Furthermore, coil current regulation is proposed to regulate the otherwise constant coil current. While the EMI issue is mitigated by the dramatically reduced standby coil current, the standby (when not charging) input power forms 44.2% of the operational input power. This implies that the system is highly inefficient during standby. Also, the standby and charging input powers for both coil current regulated and non-regulated environments are increased. This would mean that the auxiliary circuit used in the coil current regulation incurred additional power loss.

Fig. 5.1 Typical segmented D-WPT system

In this chapter, a novel D-WPT Tx control scheme is proposed for a typical D-WPT system as shown in Fig. 5.1 and had used a 3 Tx setup as reference. This control scheme would enable control of the D-WPT system without external position sensors, auxiliary circuits or communications. The transient analysis of the input Tx current would form the basis of the control algorithm where the main control variables are the current limits and pulse frequency. The performance metrics of the control algorithm, which are defined to be its energy and power losses, are also investigated.

This chapter is organized as follows. Section 5.2 introduces the input current transient analysis of the series compensated WPT system using fundamental circuit analysis. Section 5.3 presents the proposed control algorithm and

introduces the control variables: current limits and pulse frequency. Using these variables, the flowchart of the control algorithm is presented. The energy and power losses incurred by this control algorithm are also evaluated. In Section 5.4, experimental verification of the proposed control algorithm is presented using the 1-Tx WPT system. A comparison of the 3-Tx WPT system with the proposed control algorithm is also provided. Last but not least, the conclusion is drawn in Section 5.5.

5.2. Transient Analysis

In this section, the transient analysis of the S-S compensation circuit DWPT system is derived. The analysis will be used as the basis for the proposed control algorithm in the next section.

5.2.1. Circuit Analysis

The series-series (S-S) compensation circuit given in Fig. 5.2 is used in this work where capacitors (C_1 and C_R) are connected in series with the respective coils at both the transmitting and receiving sides. The S-S compensation circuit is used because the compensation capacitance required for both Tx and Rx are independent of the coupling. Assuming the coils' skin and proximity effects are included in the respective coil resistances (R_1 and R_R) and the coils are mutually coupled (M), the system matrix in frequency domain becomes:

$$\begin{bmatrix} sL_1 + \frac{1}{sC_1} + R_1 & -sM \\ -sM & sL_R + \frac{1}{sC_R} + R_R + R_O \end{bmatrix} \begin{bmatrix} i_1 \\ i_R \end{bmatrix} = \begin{bmatrix} \frac{V_{in}\omega^2}{s^2 + \omega^2} \\ 0 \end{bmatrix} \quad (5.1)$$

where $s = j\omega$.

Fig. 5.2 Series-series compensation circuit

According to [20], the coil-to-coil efficiency can be derived from (5.1) to be as:

$$\eta = \frac{1}{1 + \frac{R_R}{R_O} + \frac{R_1}{R_O} \left(\frac{R_R + R_O}{\omega_0 M} \right)^2} \quad (5.2)$$

where ω_0 is the resonance frequency of the WPT system.

By optimizing both the WPT's operating frequency and load resistance, and using measured coil resistances of $0.3 \, \Omega$ given in Table 5.2, Fig. 5.3 shows that the coil-to-coil efficiency decreases as the coupling decreases. This can be attributed to the rise of input power, which is in this case, input current as the setup is using a constant voltage source input. Therefore, the input current can be used as a control parameter for estimating the coupling coefficient.

In order to investigate the input current, the steady state and transient components of the input current, i_1 can be derived from (5.1). As the aim of the analysis is on the transient response and not the coil efficiency, R_1 and R_R are neglected to simplify the analysis. Hence,

$$i_1 = i_{1-SS} + i_{1-T} \quad (5.3)$$

$$i_{1-SS} = \frac{(\omega_0^2 C_1 C_R R_O / k^2) \omega_0}{s^2 + \omega_0^2} V_{in} \quad (5.4)$$

$$i_{1-T} = \frac{(\omega_0^3 C_1 C_R R_O (k^2 - 1)) s^2 + (\omega_0^3 C_1 k^2 - \omega_0^5 C_1 C_R^2 R_O^2) s - \omega_0^5 C_1 C_R R_O}{k^2 [(1 - k^2) s^4 + \omega_0^2 C_1 R_O s^3 + 2\omega_0^2 s^2 + \omega_0^4 C_1 R_O s + \omega_0^4]} V_{in} \quad (5.5)$$

where $k = M / \sqrt{L_1 L_R}$

From (5.3), the time domain of the steady state component can be represented as:

$$i_{1-ss}(t) = \frac{\omega_0^2 C_1 C_R R_o V_{in}}{k^2} \sin \omega_0 t \quad (5.6)$$

However, as (5.5) does not have radical factors, the numerical expression of the transient component in time domain is:

$$i_{1-T}(t) = N_1 e^{-\alpha_1 t} \cos(\omega_1 t + \theta_1) + N_R e^{-\alpha_R t} \cos(\omega_R t + \theta_R) \quad (5.7)$$

where the terms $N_{1,R}$, $\theta_{1,R}$, $\alpha_{1,R}$ and $\omega_{1,R}$ can be solved numerically. Using the circuit parameters in Table 5.1, the input current response can be seen in Fig. 5.4, together with its steady state and transient components. The frequency of the input current follows the steady state component. Due to the active behaviour of the D-WPT system, the coupling coefficient is expected to vary. In addition, depending on the type of application, the load resistance can also vary. With these variations, the maximum input current can be derived from (5.7), and is plotted in Fig. 5.5 against coupling and load resistance using the circuit parameters listed in Table 5.1.

Fig. 5.3 Coil-to-coil efficiency versus coupling coefficient

Fig. 5.4 Components of the input current in transient.

Table 5.1 Simulation Parameters

Parameter	Value
Tx inductance, L_1 (μH)	31.6
Rx inductance, L_R (μH)	31.6
Tx resonance frequency (kHz)	992.5
Rx resonance frequency (kHz)	989
Coil Resistances, R_1 and R_R (Ω)	0.3
Load Resistance R_O (Ω)	20
Cut-off coil-to-coil efficiency (%)	60
Minimum mutual inductance, M_{min} (μH)	0.622

Fig. 5.5 Maximum input current versus coupling coefficient and load resistance

From Fig. 5.5, it is clear that the peak input current increases as coupling coefficient decreases regardless of the load resistance. This noteworthy relation forms the basis of the control scheme which will be explained in the next section.

5.3. Proposed Tx Pulsation Control Scheme

The analysis of the proposed control scheme consists mainly of three parts – (1) input current limit which is derived from the minimum coupling coefficient, k_{min} which in turn is derived from the cut-off coil-to-coil efficiency; (2) pulse frequency that defines the speed and responsiveness of the control scheme; (3) power loss of such control scheme. To start off, we will cover the control algorithm flowchart. Following which, the theoretical analysis of the proposed control scheme and its parameters will be covered with simulation results.

5.3.1. Control Algorithm

Fig. 5.6 Proposed control algorithm flowchart

The proposed control algorithm flowchart is given in Fig. 5.6. At the start, we have to define the cut-off coil-to-coil efficiency for the respective current limits according to (5.9) and (5.10). Basically the operation of the proposed control scheme is as follows: the Tx is being powered on briefly and only when the Tx and Rx are not within proximity (i_{1-on} will be exceed), then Tx will cease to operate and will restart again, depending on the pulse frequency. If both Tx and Rx are within proximity (i_{1-on} will not exceed), then Tx will continue operation. When the Rx remains within the proximity of Tx, the Tx will continue operation but when the Rx moves further away from the Tx (i_{1-off} will be exceeded), the Tx will be turned off. The time differentiation between the two current limits, t_1 is crucial for the proposed control scheme to maintain correctness. The pulse frequency (or its period, T_{pulse}) is an important variable of this control algorithm as it determines the time that the WPT system starts up after being switched off.

5.3.2. Current Limits

In this sub-section, the current limits (i_{1-on} and i_{1-off}) and time, t_1 to differentiate between the two current limits will be derived. In order to determine the current limits, the cut-off coil-to-coil efficiency has to be defined. Using the pre-defined cut-off coil-to-coil efficiency, η_{min} , the minimal mutual inductance, M_{min} can be derived from (5.2) to be:

$$M_{min} = \sqrt{\frac{\frac{\eta_{min} R_l \left(\frac{R_R + R_o}{\omega} \right)^2}{R_o}}{1 - \eta_{min} - \eta_{min} \frac{R_l}{R_o}}} \quad (5.8)$$

The key parameters in this control scheme are the switching on and off input current limits, i_{1-on} and i_{1-off} which are calculated from (5.7) based on the minimal coupling coefficient, k_{min} .

$$i_{1-on} = i_{1-off} + N_1 e^{-\alpha_1 t_1} + N_2 e^{-\alpha_2 t_1} \quad (5.9)$$

$$i_{1-off} = \frac{\omega_0^2 C_1 C_R (R_R + R_O)}{k_{min}^2 + \omega_0^2 C_1 C_R R_1 R_R + \omega_0^2 C_1 C_R R_1 R_O} \sin(\omega_o t) \quad (5.10)$$

where $k_{min} = M_{min} / \sqrt{L_1 L_R}$

The difference between the two current limits is that i_{1-on} may contain overshoot during its transient response as compared to i_{1-off} which only consists of the steady state component. This is shown in Fig. 5.7 where the input current limit, i_{1-on} is normalized to i_{1-off} against coupling coefficient. At lower couplings, the two current limits are similar (negligible overshoot during transient) but when higher cut-off coupling is chosen, the disparity between both current limits increases (more significant overshoot during transient).

Fig. 5.7 i_{1-on} current normalized to i_{1-off} current

Next, to differentiate between the two different current limits, t_1 is defined as the time for the input current, i_1 to reach its transient peak and Fig. 5.8 shows the numerical solution of t_1 against coupling coefficient. The selection of t_1 will affect the integrity of the proposed control scheme. If minimum values of t_1 is used, the two current limits will not be properly differentiated, and the control scheme might be erroneous. Therefore, to differentiate between the two current limits in (5.9) and (5.10), the highest t_1 according to k_{min} must be selected and doubled, so that the rising and falling times during the transient period are

Fig. 5.8 Time (t_1) of the input current's maximum amplitude versus coupling coefficient

accounted, thereby ensuring the algorithm reliability. Next, if the Tx is switched off, the control scheme will restart operation according to the pulse frequency.

5.3.3. Pulse Frequency

The pulse frequency is defined as follows: assuming that the Tx is initially switched off, the time taken for the control scheme to restart its operation as per described in previous sub-section, is determined by its pulse frequency. Therefore, the higher the pulse frequency, the lower the latency between each operation. However, the energy stored in the Tx during the transient period before exceeding i_{1-on} , has to be dissipated prior. If the time period between each pulse is too small for the WPT system to dissipate its energy before the next cycle, the resultant action of the control scheme can be erroneous.

The lowest energy dissipation rate occurs when the WPT system is uncoupled ($k = 0$), since the dissipation of energy will be isolated within the transmitting side only. Using the parameters from Table 5.1 and the uncoupled condition, Fig. 5.9 shows the input current dissipation profile for aforementioned worst case scenario. As there is no finite solution to the dissipation current, the system is deemed have dissipated its input current when the input current falls below 1% (as marked by the dotted line) of its current limit, i_{1-on} . In this case, according to

Fig. 5.9 we take 1 ms as the upper limit of the pulse period (or 1 kHz to be the highest pulse frequency).

After determining the upper limit of the pulse frequency, both power and energy losses of this proposed control scheme can be calculated, as described in the next sub-section.

Fig. 5.9 (a) Input current dissipation profile at uncoupled condition. (b) Zoomed in view of input peak current dissipation profile and dotted line shows the 1% mark

5.3.4. Power and Energy Losses

According to the maximum pulse frequency derived in previous sub-section, the pulse frequency will be selected to be 1 kHz in order to obtain the maximum power loss. The energy loss of the proposed control scheme is defined to occur during the timespan when the Tx is briefly powered on but is subsequently switched off as the Rx is not within the desired proximity of the Tx. The power loss arising from the normal operation of the WPT circuit is not factored in as it is independent of the proposed control scheme. The power loss is derived from (5.3) and the aforementioned timespan, which is dependent on the input current transient response and i_{1-on} from (5.9). The power loss of the proposed control scheme is defined as:

$$P = f_{pulse} \int_0^{t_1} I_1 \cdot V dt \quad (5.11)$$

With the power loss defined in (5.11) and timespan, the energy loss of the proposed control scheme is formed in Fig. 5.10 to illustrate the impact of the coupling coefficient and equivalent load resistance on the proposed control scheme energy loss. Comparing the energy loss of a particular load resistance against coupling coefficient, we can see that for coupling coefficients lower than k_{min} , more energy loss is incurred when the coupling coefficient is nearer to the respective k_{min} (as the time taken for i_1 to exceed the current limit i_{1-on} is significantly more) than when the WPT system is less coupled. It is also important to note when comparing the energy losses between the various load resistances, the same current limit, i_{1-on} , will result in a different k_{min} . In addition, as suggested by [21] that the WPT system's efficiency is dependent on the load resistance, which explains the increasing energy loss that comes along with the increasing load resistance.

In addition, (5.11) clearly shows the proportionality between pulse frequency and power loss. Having such a correlation, the pulse frequency can be chosen based on the nature of the application.

Fig. 5.10 Energy loss of the proposed control algorithm versus coupling coefficient for various load resistances

For low speed applications such as mobile phone charging, a pulse frequency of 1 Hz would translate to a maximum charging latency of 1 second. For high speed applications such as in-motion charging of electric vehicles, choosing a pulse frequency of 1 kHz would incur a displacement of 14 mm during its period (assuming typical urban speeds of 50 km/h), which is relatively small compared to typical vehicular Rx coils. In the following sections, the higher 1 kHz will be used as pulse frequency in the later analyses.

5.4. Experimental Verification

In this section, we will demonstrate the proposed control scheme via experimental verification. The experimental verification is divided into two sub-sections namely, 1-Tx setup and 3-Tx setup. The former is used to verify the transient analysis and control algorithm, the latter is used to demonstrate the coil-to-coil efficiency improvements.

5.4.1. 1-Tx setup

In order to verify the input current transient response and the proposed control algorithm, an experiment system has been implemented based on the 1-Tx WPT setup shown in Fig. 5.11. The two helical coils, Tx and Rx create the displacements, y_{rx} and z_{rx} according to their relative positions. The parameters of the 1-Tx WPT experimental setup are listed in Table 5.2 and the simulation results are obtained from MATLAB/Simulink while the experiment is carried out using Keysight E5061B ENA Series 2-port Network Analyzer (measured S-parameters are used to generate experimental transient response using Agilent ADS tool [22]) which is shown in Fig. 5.12.

Firstly, for the experimental setup, z_{rx} is fixed at 20 cm while y_{rx} varies from 0 to 40 cm with intervals of 5 cm each. The coil-to-coil efficiency against y_{rx} is shown in Fig. 5.13, presenting close comparison of experimental and simulation results. Next, the input current transient response (at $z_{rx} = 10$ cm and $y_{rx} = 0$ cm),

experimental and simulation input current transient responses are verified to match well in Fig. 5.14. The proposed algorithm is implemented in Agilent ADS and its cut-off coil-to-coil efficiency is set at 60%, which corresponds to a minimum mutual inductance of $0.496 \mu\text{H}$ according to (5.8). Using both (5.9) and minimum mutual minimum mutual inductance, the input current switching on limit, i_{1-on} can be calculated to be 0.73A . The proposed algorithm is implemented in Agilent ADS as shown in Fig. 5.15, where the excitation voltage is removed when the transient current exceeds the limit and then current dissipation occurs when $z_{rx} = 20 \text{ cm}$ and $y_{rx} = 15 \text{ cm}$. This is in comparison with the input current transient response when $z_{rx} = 20 \text{ cm}$ and $y_{rx} = 10 \text{ cm}$, where the input current does not exceed the limit and continues to operate normally. Lastly, for dissipation, the same limit i_{1-on} of 0.73A is implemented for the least coupled position ($z_{rx} = 20 \text{ cm}$ and $y_{rx} = 40 \text{ cm}$). The system is dissipated within the period (1 ms) of the pulse frequency as depicted by the transition in Fig. 5.14.

Fig. 5.17 shows multiple pulse excitations where the WPT system ($z_{rx} = 20 \text{ cm}$ and $y_{rx} = 15 \text{ cm}$) has a pulse frequency of 1 kHz . As the input current exceeds the pre-defined i_{1-on} , the input voltage is removed before being applied again at the onset of the next pulse period. When the input current exceeds i_{1-on} , the WPT system's input energy is considered as loss energy whereas when i_{1-on} is not exceeded, the system continues operating and there is no energy loss arising from this control algorithm. Fig. 5.18 shows the comparison of experimental and analytical energy loss against y_{rx} when $z_{rx} = 20 \text{ cm}$. Multiplying the energy loss with the highest pulse frequency, the highest input power during standby mode (when $z_{rx} = 20 \text{ cm}$ and y_{rx} is between 15 cm to 40 cm) is about 14.5 mW . On the other hand, according to Fig. 5.19, when the Rx is aligned to the Tx ($z_{rx} = 20 \text{ cm}$ and $y_{rx} = 0 \text{ cm}$), the WPT system input power is 0.191 W . Therefore, the proposed control algorithm has a lower standby-to-charging-mode input power percentage of 7.6% as compared to the 44.2% reported by [15].

Fig. 5.11 1-Tx WPT system diagram.

Fig. 5.12 Experimental 1-Tx WPT system setup.

Table 5.2 Experimental Parameters

Parameter	Values
Tx inductance, L_1 (μH)	31.6
Rx inductance, L_R (μH)	31.6
Tx resonance frequency (kHz)	992.5
Rx resonance frequency (kHz)	989
Coil Resistances, R_1 and R_R (Ω)	0.3
Coils' radius (cm)	8.9
Load Resistance R_O (Ω)	20
Cut-off coil-to-coil efficiency (%)	60
Minimum mutual inductance, M_{min} (μH)	0.622
Gap distance, z_{rx} (cm)	10 & 20
Range of y_{rx} (cm)	0 to 40 (in intervals of 5)
Distance between Tx, y_{tx} (cm) [for 3-Tx only]	30

Fig. 5.13 Experimental and simulation coil-to-coil efficiency variation with respect to y_{rx} when $z_{rx} = 20$ cm

Fig. 5.14 Experimental and simulation input current transient responses when $z_{rx} = 10$ cm and $y_{rx} = 0$ cm

Fig. 5.15 Experimental implementation of i_{l-on} ($= 0.73$ A) for ($z_{rx} = 20$ cm and $y_{rx} = 10$ cm) and ($z_{rx} = 20$ cm and $y_{rx} = 15$ cm)

Fig. 5.16 (a) Input current dissipation when $i_{1-on} = 0.73$ A, and $z_{rx} = 20$ cm, $y_{rx} = 40$ cm. (b) Zoomed in view of the input current dissipation (dotted line represents the 1% of current limit, i_{1-on})

Fig. 5.17 Pulsing the WPT system at 1 kHz when $z_{rx} = 20$ cm and $y_{rx} = 15$ cm

Fig. 5.18 Experimental and analytical energy losses against y_{rx} when $z_{rx} = 20$ cm

Fig. 5.19 Experimental and analytical normalized input power of the WPT system against y_{rx} when $z_{rx} = 20$ cm

5.4.2. 3-Tx setup

The proposed control scheme is applied to the 3-Tx WPT system and its experimental setup and circuit diagram are shown in Fig. 5.20 and Fig. 5.21 respectively. As compared to the 1-Tx WPT system, two additional Tx's are added to the 3-Tx WPT system and a 4-port Rohde & Schwarz ZVB 8 Vector Network Analyzer is used in this experiment. The Tx's are connected individually to the Vector Network Analyzer's ports. Also, z_{rx} is changed from 20 cm to 10 cm in order to maintain operation continuity within the span of y_{rx} . Without the proposed control scheme, the 3 Tx are kept switched on throughout.

Coil-to-coil efficiency dips significantly when the Rx is aligned to Tx₃, which corresponds to $y_{rx} = 30$ cm. This is due to the low coupling coefficient between both transmitters (Tx₁, Tx₂) and Rx, which increased Tx₁ and Tx₂ input currents, and also overall input power. Employing the proposed control scheme, the 3-Tx WPT coil-to-coil efficiency is compared in Fig. 5.22 where η_{min} is kept at 60%. Firstly when $y_{rx} = 0$ to 10 cm, only Tx₂ is switched on. In the middle ($y_{rx} = 15$ cm), both Tx₂ and Tx₃ are switched on. At the end ($y_{rx} = 20$ to 30 cm), only Tx₃ is switched on. The same switching order can be said for the other side of the 3-Tx setup when y_{rx} is negative. By turning on or off the respective Tx using the proposed control algorithm, the coil-to-coil efficiency of the D-WPT system can be improved. With this control algorithm in place, coil-to-coil efficiency improvements of up to 52% (when $y_{rx} = 30$ cm) can be attained and the coil-to-coil efficiency of the WPT system can be expected to operate above the cut-off coil-to-coil efficiency while enabling dynamic control of the WPT system without any secondary controller or communication.

The proposed control algorithm is employed on a series-series compensation WPT system, but it can be employed on other types of compensation by using the transient analysis found in this chapter and adapting the control algorithm accordingly. Also, the power level can be easily increased by increasing the voltage levels in the proper manner.

Fig. 5.20 3-Tx WPT experimental setup

Fig. 5.21 3-Tx WPT system diagram

Fig. 5.22 Experimental coil-to-coil efficiency results for 3-Tx WPT system ($y_{tx} = 30$ cm, $z_{rx} = 10$ cm and $y_{rx} = 0$ cm to 30 cm) using proposed control algorithm

5.5. Conclusion

This chapter proposed a control algorithm for D-WPT system on the primary side to control the transmitter(s) in order to improve the overall coil-to-coil efficiency. The main control variable, which is the Tx's input current, has been theoretically analyzed and modelled. The other important aspects of the algorithm such as current limits, pulse frequency and power loss have been analyzed. Using these control parameters, a 1-Tx WPT experimental system was used to demonstrate the feasibility of the proposed control algorithm. The control algorithm was also implemented on a 3-Tx experimental setup and it shows significant overall coil-to-coil efficiency improvements of up to 52%.

Chapter 6

Conclusion and Future Works

This chapter concludes the research work covered in this thesis and the research on WPT system will be summarized. Also, potential future research works on WPT system and applications are proposed.

6.1. Conclusion

Wireless power transfer (WPT) is a potential technology that can be employed in a wide variety of application. This can be achieved due to WPT's inherent advantages over its wired counterpart, which enables WPT to eliminate numerous drawbacks faced by its wired counterpart. However, WPT still need to overcome numerous technical challenges before it can truly be ubiquitous in the modern connected world. This thesis has proposed several approaches to improve WPT's performance in terms of efficiency, transferred power, etcetera. In addition, a novel control scheme for dynamic WPT (D-WPT) systems without needing external sensors, and this helps solve some complexity and implementation issues of a D-WPT system.

Early WPT works uses inductive coils and its electromagnetic coupling to transfer power wirelessly without electrical contact. These loosely coupled inductive coils are driven by AC frequency power sources which causes large leakage inductance and significant input impedance. This limits the performance in terms of efficiency and power transfer, which causing the gap transfer distance to be limited. By employing compensating capacitors and operating the WPT system at its electromagnetic resonance frequency, the input impedance of the WPT system can be significantly reduced, and its VA rating reduces. This leads to an increase in the performance in terms of efficiency and power transfer, as well as transfer gap distance. However, there are many different types of compensation circuits available for WPT systems and these compensation circuits perform differently under varied conditions. It is therefore unclear which type of compensation circuit is suitable for any given application or load. The work in this thesis dwelled into the circuit analyses on the series transmitting, and the series and parallel receiving compensation circuits. The efficiency optimal operating frequency is determined using induced and reflected load equivalent circuits of these WPT systems. Using the optimized operating frequency, the efficiency analysis on the load variation of the basic compensation circuits are

detailed. The efficiency analysis is supplemented with the crossover resistance which is defined as the point in load resistance when the respective circuits' efficiencies intersect. Resonant frequency response of the parallel compensation receiving circuits, parallel and LCL are used to further differentiate these circuits where parallel is known to be varying according to coupling coefficient but LCL can be designed according to have better control on the system resonant frequency. The additional inductor in the LCL compensation circuit allows the designer relative control of the resonant frequency and to maintain soft switching if inductive switching is required. Using the aforementioned studies and analyses, the efficiency orientated design flow chart for selecting the optimized compensation circuit is presented. In addition, important output parameters such as transferred power, transconductance and voltage gain are analyzed in depth, which could be useful for certain applications requiring specific output performance. Experimental results also verified the usefulness of the flow chart and its respective analyses, which showed an efficiency discrepancy of 28% at 500 Ω equivalent output resistance when comparing between series and LCL compensation circuits.

Moving on in Chapter 4, the LCC compensation circuit for WPT system is analyzed and compared with the series compensation in terms of efficiency and transferred power impacts caused by coupling variations. The chapter presents the derivation of the WPT system resonant frequency and goes further to analyse the impedance frequency response between the two aforementioned compensation circuits. Bifurcation of the series compensation circuit's impedance can be easily seen whereas for the LCC compensation circuit, the bifurcation effect is mitigated significantly. The mitigation of the impedance frequency response bifurcation shows that LCC compensation circuit is most favourable for WPT systems employing specific operation frequency while having dynamic coupling levels during operation. For applications using variable operating frequency controls, the mitigation of the impedance frequency

response can serve to simplify the control algorithm and improve on the system stability and reliability. Also, LCC compensation circuit allows for significantly higher transferred power as the input impedance of the WPT with LCC compensation circuit is significantly lower than that of the series compensation circuit. In addition, the harmonic distortion of the WPT system with LCC compensation circuit is significantly lower than that of the series compensation, when the WPT system is powered by typical high frequency power electronic converters.

Continuing on the momentum of the WPT technology, dynamic WPT (D-WPT) aims to bring the application of WPT further by allowing the receiver (Rx) to move freely while receiving power from the D-WPT system wirelessly. There are two types of magnetic coupler structures used in D-WPT systems. The first structure type is track based where the magnetic track spans the charging area and the Rx is charged when it moves along the long magnetic track. While the track is able to create a substantial charging space, such magnetic coupler structure faces poor coupling levels, low efficiencies and electromagnetic interference issues. The second structure type is segment-based magnetic coupler where individualized magnetic couplers span the charging area. This magnetic coupler structure enables individual control of each magnetic coupler, allowing a larger portion of the electromagnetic field created by the magnetic coupler to be captured by the Rx and increasing the coupling level and efficiency. However, the control of each individual magnetic coupler contributes to the control complexity, component and power electronics count, and increases the D-WPT system susceptibility to misalignment and the variation in performance. In Chapter 6, the thesis proposes a novel control scheme for the purpose of individual magnetic coupler control by sensing the transient response of the magnetic coupler's input current. This would improve the D-WPT system's efficiency, without requiring the need for external positional sensors or auxiliary circuitry. Chapter 6 begins with the transient analysis of the D-WPT system input

current using circuit analysis. The concept of the proposed Tx pulsation control algorithm is detailed and its main parameters, namely current limits, pulse frequency and its power and energy losses. Moving on, the 1-Tx experimental setup is used to validate the proposed control algorithm concept, and the 3-Tx experimental setup is used to demonstrate the proposed control algorithm where the coil-to-coil efficiency has seen improvements of up to 52%. Comparing the percentage of input power during standby mode to input power during operational mode, the proposed control algorithm produces a low percentage of 7.2% as compared to 44.2% found in existing literatures. Thus, the proposed control algorithm for D-WPT system is a promising approach to improve power consumption and EMI/EMC performances during standby mode, as well as power transfer efficiency levels.

6.2. Future Works

Having covered the thesis work on WPT systems, there is potential areas for more research and development as outlined below:

- The compensation circuit design flow chart proposed in Chapter 3 has evaluated several compensation circuits for a given load. For dynamic load resistances, the selected compensation circuit might not yield optimized power transfer efficiency transfer. In order to improve the power transfer efficiency, an adaptive compensation circuit for the WPT system can be used to adjust the compensation circuit accordingly. However, as the compensation circuits store energy during operation, there would be a need to investigate the energy aspect of the adaptive compensation circuit. This would allow optimized power transfer efficiency for relatively wide load resistance applications.
- The proposed control algorithm presented in Chapter 5 has been used for D-WPT systems with single Rx. For D-WPT (or WPT) systems with relatively larger Tx and smaller Rx, multiple Rxs can be placed within the electromagnetic fields of the Tx. Applications of such D-WPT systems

can be for charging multiple consumer electronic devices on a single Tx, electric vehicles on a tracked Tx system, etcetera. Therefore, it would be interesting to explore using the proposed control algorithm as a baseline for developing a control algorithm for multiple Rx D-WPT systems.

- The investigation of the power grid stability when implementing D-WPT system on a larger scale. This is due to the nature of the D-WPT that the Rxs would be considered as highly intermittent loads (intermittency would be dependent on Rx's rate of displacement) which could create instability and noise into the affected power grid.
- Implementation of WPT in military and para-military environments. Even though WPT technology is able to operate better than its wired counterparts in terms of reliability and maintainability, commercial standards for implementation of WPT technology in such harsh environment often fall short. For example, the typical maximum transfer gap distance is insufficient for military products employing ruggedized casing and protection. WPT for military application has to be further optimized in order to improve on the performance in order to be applicable in such environments. With the digitalization of today's soldier and the battlefield, such research and development topic would be an interesting topic to explore.

6.3. Author Publication List

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