Efficient circuit topologies for inductive power transfer

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EFFICIENT CIRCUIT TOPOLOGIES FOR INDUCTIVE POWER TRANSFER

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**SCHOOL OF ELECTRICAL AND ELECTRONIC
ENGINEERING**

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2015

To my loving family

Abstract

In recent years, inductive power transfer (IPT) systems have gained much attention, interest and development. IPT systems are found to be having numerous advantages over conventional wired power systems in relation to convenience, safety, isolation, operation in hostile conditions and flexibility and have proven its worth in many applications such as electric vehicle charging systems, lighting, material handling, grid-tied PV inverters, etc.

A typical IPT system employs a primary power converter to generate high frequency current into one or more inductive tracks/coils in the primary side and then magnetically coupled to one or more power pick-ups in the secondary side. Development of efficient IPT systems has been receiving increased attention and many attempts have recently been made to improve the overall efficiency. The overall efficiency of an IPT system largely depends on the losses that incur in coupling coils and converters.

This thesis focuses on the development of power converters and modulation strategies for IPT systems to minimize the power losses and improve the power transmission capability. In case of power converters, the power losses consist of conduction losses and switching losses of switching devices. In order to reduce these power losses, the power converters should either have reduced number of power switches and power stages or efficient commutation strategies. In the case of the coupling coil losses, it can be minimized by either with the use of proper magnetic circuit and coil winding design or an appropriate modulation strategy. Towards that end, certain techniques are proposed to develop power converters and control algorithms with optimized overall efficiency for IPT systems.

Firstly, in order to give the reader sufficient knowledge to comprehend the proposed power converters and modulation strategies that will be discussed in this thesis, chapter 2 will present an overview of IPT circuit topologies. Fundamental components of IPT systems will be briefly introduced first. Thereafter, a review of IPT power converters together with switching methods will be presented.

Secondly, a novel direct AC-AC matrix converter topology is proposed to generate a high frequency current with a reduced number of semiconductor switches. The proposed topology transforms 50-Hz 3-phase utility supply to a single-phase high frequency supply which in turn can be used to directly excite the primary side resonant network of an IPT system. With the reduction of number of conversion stages, switching and conduction losses can be reduced. A mathematical model for the proposed system has also been presented with simulation and experimental results to demonstrate the feasibility of the proposed topology. The power loss evaluation of the proposed converter as well as of the conventional IPT converter demonstrated the benefits of the proposed system.

Thirdly, an optimized phase shift modulation strategy has been proposed to minimize coil losses of bidirectional IPT (BIPT) systems. Moreover, a comprehensive study on the impact of system parameters on the overall efficiency has been carried out. The analyses and simulation results provide the conditions for obtaining the maximum overall efficiency. A closed loop controller has been proposed to operate the system with optimized efficiency. Experimental results demonstrate the feasibility of the proposed concept.

Fourthly, multilevel converters including cascaded multilevel converters, diode clamped multilevel converters, capacitor clamped multilevel converters and other advanced multilevel converter topologies with reduced number of power switches are introduced for BIPT systems together with proposed modulation strategies to minimize the power losses of switching devices. The presented power converters can be used in low to high powered applications. Cascaded multilevel converters are suitable in the case of several individual DC sources such as PV cells are connected while diode clamped multilevel converters and capacitor clamped multilevel converters are suitable in the case of high voltage power supply. Multilevel converters with reduced number of power switches are employed to lower switching losses and conduction losses of IPT systems.

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Nomenclature

Acronyms

AC	Alternating current
AGV	Automated guided vehicle
BIPT	Bidirectional inductive power transfer
CC-MLC	Conventional Cascaded Multilevel Converter
CCL	Capacitor–capacitor–inductor connection
CLCL	Capacitor–inductor–capacitor–inductor connection
CPT	Contactless power transfer
DC	Direct current
DSP	Digital Signal Processor
EMI	Electro–magnetic interference
FC	Flying Capacitor
FPGA	Field–programmable gate array
FFT	Fast Fourier Transform
IGBT	Insulated gate bipolar transistor
IPT	Inductive power transfer
LC	Inductor–capacitor connection
LCL	Inductor–capacitor–inductor connection
LCCL	Inductor–capacitor–capacitor–inductor connection
MC-MLC	Modified Cascaded Multilevel Converter
MH-MLC	Modified Hybrid Multilevel Converter
MHB-MLC	Modified H-bridge Multilevel Converter
MOSFET	Metal oxide silicon field effect transistor
ms	Milliseconds
NPC	Neutral Point Clamped
OLEV	Online electric vehicle
PLF	Power loss function
PWM	Pulse width modulation

PV	Photovoltaic
RMS	Root mean square
s	Seconds
Si	Silicon
SiC	Silicon Carbide
SS	Series series compensation
SP	Series parallel compensation
PS	Parallel series compensation
PP	Parallel parallel compensation
WPT	Wireless power transfer
ZCS	Zero current switching
ZVS	Zero voltage switching

Symbols

e_{SW_ON}	Turn-On switching loss (Joules)
e_{SW_OFF}	Turn-Off switching loss (Joules)
V_{DC}	DC voltage (Volts)
V_{AC}	AC voltage (Volts)
V_{in}	Input voltage (Volts)
V_p	Primary side input voltage (Volts)
I_{in}	Input current (Amperes)
V_{ref}	Reference voltage (Volts)
V_f	Threshold voltage (Volts)
v_{ak}	Anode–Cathode voltage (Volts)
C	Capacitor (Farads)
C_p	Primary Capacitor (Farads)
C_s	Secondary Capacitor (Farads)
f_T	Switching frequency (Hz)
i	Instantaneous current (Amperes)
I	Current magnitude (Amperes)
I_0	RMS value of current (Amperes)
j	Complex operator ($j^2 = -1$)

k	Magnetic coupling coefficient
L	Self-inductance (Henrys)
L_a	Splitting inductance (Henrys)
L_b	Splitting inductance (Henrys)
L_p	Primary inductance (Henrys)
L_s	Secondary inductance (Henrys)
M	Mutual inductance (Henrys)
P	Power (Watts)
Q_{RR}	Reverse recovery charge (Coulomb)
R	Resistance (Ω)
r_d	Equivalent on-state resistance of diode (Ω)
R_p	Primary side equivalent resistance (Ω)
R_s	Secondary side equivalent resistance (Ω)
R_{DC}	DC Load (Ω)
R_{Leq}	Equivalent AC load (Ω)
t	Time (seconds)
V	Voltage magnitude (Volts)
v	Instantaneous voltage (Volts)
Z_r	Reflected impedance (Ω)
ϕ	Phase shift angle of converters (radian)
θ	Phase shift angle between primary and secondary converters (radian)
ω_0	Grid side angular frequency (radians/s)
ω_T	Switching angular frequency (radians/s)
η	Efficiency
$\int$	Integral operator
ξ	Converter output voltage ratio

Algebraic Notations

d/dt	Differential operator
$d\eta/d R_{Leq}$	Differential operation to efficiency
Im	Imaginary part of a complex expression

Re	Real part of a complex expression
$\angle$	Represent of complex number phase angle

Subscripts

AC	AC value
AV	Average value
DC	DC value
in	Input
o	Output
m	Magnitude
max	Maximum value
min	Minimum value
MOS	Mosfet
p	Primary track
r	Referred back from the pick-up to the primary track
s	Secondary pick-ups
SW	Switching
cond	Conduction
opt	Optimal
for	Forward
rev	Reverse

Superscripts

⁰	Degree
*	Complex conjugate

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Introduction

1.1 Background

1.1.1 Introduction to Inductive Power Transfer system

In 1890, the contactless power transmission technology was discovered when Nicola Tesla transferred power wirelessly [1]. Although this technique faced a big hurdle due to the limit of the semiconductor devices switching speed at that time, it has been gaining popularity in several recent decades as they offer numerous advantages over conventional wired technology.

Basically, power can be wirelessly transferred by many methods: laser [2 - 4], electromagnetic waves [5 - 7], static electric fields [8 - 10], and static magnetic fields [11 - 13]. All these wireless power transfer technologies offer advantages but come with their inherent constraints.

The term Inductive Power Transfer (IPT) (or Wireless Power Transfer –WPT, or Contactless Power Transfer – CPT) can be used to describe the power transmission between two or more physical objects via electro-magnetic coupling. Unlike the above techniques, IPT employs one or more inductive coils/tracks, called as primary windings, which carry a high frequency current in one side. One or more inductive coils in the other sides, called secondary windings, will pick the current up from the primary side by mutual coupling with no physical contact. Thereby, the power can be transferred in hostile environments, long distance, stationary or movable objects.

A power converter is employed in the primary side to generate high frequency current from DC power supply or AC mains. Depending on the applications, the output pickup side is connected to a power converter to convert from high frequency current to DC or AC current. The primary and secondary coils/tracks are compensated by capacitive

components. The system is operated at the resonant frequency which is tuned with the compensated circuit so that the magnetic field emitted from the coils is maximized.

1.1.2 Achievements in IPT technologies

➤ In research aspects

In recent years, a growing number of research groups involved in the field of IPT, such as Massachusetts Institute of Technology (MIT), Illinois Institute of Technology, University of Michigan, Tokyo University, The University of Auckland, Nanyang Technological University, Tohoku University, City University of Hong Kong, Chongqing University, Korea Advanced Institute of Science and Technology (KAIST), Utah State University (US), etc.

In 2006, “WiTricity”, based on IPT technologies, has been shown in [14] by Professor Marin Soljacic and his colleagues from MIT that 60 W could be transferred over 2 meters with 40% efficiency.

In 2012, Professor Chun-Taek Rim (KAIST) and his colleagues have presented a well performance in IPT technologies when delivering an output power of 1403 W, 471 W and 209 W over 3m, 4m and 5m. Efficiencies have been reported to be 50%, 30% and 16% for distances of 3m, 4m and 5m respectively [15].

At the 2013 IEEE Energy Conversion Congress and Exposition Conference in Denver, USA, there were two special sessions focusing on wireless power for electric vehicles. In this conference, Dr. Hunter Wu from Utah State University has presented their IPT system for electric vehicles (EVs). 50 kW has been transferred wirelessly over ~ 18 cm with more than 90% efficiency [16].

In 2014, Professor Chunting Chris Mi (University of Michigan) and his colleagues have shown their achievement when transferring an amount of 7.7 kW power over an air-gap of 20 cm with efficiency 97% [143].

In 2015, Dr. Jae Hee Kim (the Korea Railroad Research Institute) and his colleagues have presented an 1 MW IPT system for high speed train with the efficiency up to 82.7% [114].

➤ **In commercialization aspects**

Many industrial types of machinery based on IPT systems had been built as a result of the cooperation between Universities or Research Institutes and enterprises such as Conductix-Wampfler in Germany, Daifuku in Japan, and Telementary Research, Halo IPT, Powerbyproxi, Cabco industries, Dream solutions and 3I innovations in New Zealand.

(a) “Ramrum” material handling system (b) AGV powered by IPT system

Figure 1.1: Daifuku IPT applications

Figure 1.2: IPT vehicle in Rotorua, New Zealand

In 1991, the corporation between Daifuku manufacturer (from Japan) and Uniservices (from University of Auckland, New Zealand) has created an Overhead Monorail system, the “Ramrum” material handling system, Cleanway, CleanStocker and AGV systems (Figure 1.1) [17].

In 1998, the first vehicle in the world using IPT for charging has been in operation in Rotorua, New Zealand (Figure 1.2) in partnership between Conductix-Wampfler and Uniservices [18].

Later, their IPT system has been installed in the BMW plant in Germany and then adopted by other car manufactures such as Daimler, Ferrari, GM.

In 2011, Rolls-Royce released “102 EX Phantom Experimental Electric” at the Geneva Motor Show with an inductive mat on its underside, allowing wirelessly charging in-road across an air gap of up to 40 cm.

Figure 1.3: Halo-IPT to wirelessly charge super luxury car [19]

The most significant event is the electric buses which provide public transportation services in South Korea in 2013. The Korea Advanced Institute of Science and Technology (KAIST) developed the Online Electric Vehicle (OLEV) system that can be charged while stationary or driving with 100 kW at 85% efficiency while the air gap between the underbody of the vehicle and the road surface is maintained at 17 cm.

Figure 1.4: OLEV bus runs along an inner city route in Gumi, Republic of Korea, from Aug. 6, 2013 [20].

In the area of low power applications, with the advent of inductive power standard (Qi) developed by Wireless Power Consortium, a new generation of mobile phone, tablet and laptop are to be released in near future with the initiation of Nokia Lumia 920 which is the first smartphone in the world applying IPT technology [21].

Figure 1.5: Energizer-pad for charging small portable devices [22]

Figure 1.6: IPT powered roadway lighting system of 3I Innovations Ltd. [25]

In addition, many other applications in industry were also reported in relation to IPT such as wireless roadway/tunnel lighting systems by 3I-Innovation Ltd. in New Zealand (Figure 1.6), charger adaptors by Energizer [22], Witricity [23], Powercast [24], etc.

1.2 Motivations

With the development of modern technologies, the diversification of IPT systems is rapidly expanding as IPT becomes more feasible due to the difficulties in the conventional wired power transfer as well as the benefits of IPT systems in relation to:

☞ Convenience

A future world without wired power is being expected:

- Putting the phones on the table and it will be automatically charged without any connectors or cables. Adaptor compatibility for cell phones, tablets, laptops, music players will be a thing of the past.
- Purchasing a bulb in the market and replacing the broken one at home without any further wires or other equipment. Installation would be seamless.
- Driving an electric car on a street with automated wireless charging. Range anxiety, the bane of all electric vehicles' owners would be eliminated.

The above examples are some applications that IPT systems are able to offer convenience.

☞ Safety

An IPT system transfers the power completely insulated as all the electrical connectors are eradicated. Therefore, all the risks caused by electrical shock are eliminated.

☞ Operation in hostile or harmful conditions

In some special cases, a robot has to be operated in inaccessible places and IPT offers a solution for charging.

☞ Reliability

The elimination of mechanical and electrical contacts increases the life time of the electric components, thus improving the system reliability as well.

☞ Wide power range

The power levels of the IPT systems range from mili-watts for biomedical applications to few kilo-watts for electrical vehicle battery charging and so on. The applications of IPT systems are limitless.

☞ Environment friendly - Suitable for harsh or sensitive environments

IPT system is unaffected by the environment conditions due to the elimination of electrical contacts. It can work in both harsh and clear environments. In the case where IPT is applied in the production line for some equipment, the elimination of sliding brushes and mechanical parts makes the system noiseless during operations.

➤ **Design freedom**

The flexibility in designing the IPT systems enables innovative solutions.

➤ **Low operating and maintenance costs**

With the elimination of mechanical and electrical contacts, the operating costs and maintenance costs related to wear and tear are greatly reduced.

With numerous benefits listed above, IPT system research has become one of the most popular topics that bring interesting and fruitful results. Although wireless charging is already available for mobile devices which are low power applications, medium to high power IPT systems are being investigated with the aim of optimizing the efficiency and power transfer scalability. Indeed, all the aforementioned advantages of IPT systems will not make sense without the consideration of power efficiency. While researchers from all over the world are putting their efforts to look for the renewable energy resources, the utilization of energy by the most efficient way is also a very big challenge. Therefore, this thesis is conducted with the aims to find out new power topologies (converters and modulation algorithms) which are suitable for IPT systems with the lowest power losses.

1.3 Objective and scope of the thesis

The objective of this thesis is to propose and develop new IPT power converters and modulation strategies to achieve higher overall efficiency and power transmission scalability while lowering cost and physical size. The thesis proposed a novel efficient direct matrix converter topology with a reduced number of power conversion stages and power switches. In addition, multilevel converters with different circuit topologies are developed to excite the primary resonance circuit with improved power transmission capability by taking into account the converter losses. Moreover, an optimized phase shift modulation strategy has been proposed to get maximum overall efficiency of bidirectional IPT (BIPT) systems. The remaining of this thesis is organized as follows.

Chapter 2 presents an overview of IPT systems in relation to circuit topologies and modulation strategies for IPT systems which directly affect the overall efficiency of IPT systems.

Chapter 3 proposes a new SiC based matrix converter topology to excite the primary side resonance circuit. With the reduction of the number of power conversion stages and power switches, the overall efficiency of the IPT system is improved as shown in the analyses and in the simulation results. An efficiency comparison between the proposed topology and conventional topology has been derived to demonstrate the efficacy of the proposed concept.

Chapter 4 proposes a novel modulation strategy with minimized coil losses for BIPT systems with a series-series compensation circuit. Moreover, a comprehensive study on the impact of system parameters has been investigated to obtain the maximum overall efficiency of BIPT systems.

Chapter 5 presents several multilevel converter topologies with proposed advanced modulation strategies to obtain lower converter losses and higher power transmission capability. The proposed power converters are suitable for a wide range of power level which can be employed for almost all industrial applications.

Chapter 6 summarizes the thesis and suggests some directions for future works.

1.4 Contribution of the thesis

This thesis provides some contributions as follows,

- Firstly, a new direct matrix converter topology has been proposed for IPT systems. In the conventional IPT systems, high frequency currents are usually generated from DC power supplies by high frequency power converters. As the DC source is not ubiquitous for converting to high frequency currents to supply IPT systems, an additional power stage AC/DC rectifier is needed. In addition, a DC link is necessary to be added between the rectifier and inverter. The additional power conversion stage and power elements reduce the system efficiency, reliability and increase the cost and physical size. The proposed matrix converter overcomes the drawbacks of the conventional IPT systems. The analyses together with the simulation and experimental results show the benefit of proposed topology compared to the

conventional power converters in relation to the simplicity and efficacy. In addition, the proposed converter can be employed in both the unidirectional and BIPT system.

- Secondly, an optimized phase shift modulation strategy to minimize the coil losses of a series-series compensated BIPT system was proposed. Moreover, the impact of the system parameters such as DC voltage ratio, output AC voltage ratio, output power range on the overall efficiency of the system is also carried out. An algorithm for designing a closed loop controller is proposed to maximize the overall efficiency. Theoretical analyses and simulation results are verified by experimental results to demonstrate the applicability of the proposed system.
- Thirdly, because of the limitation of switching devices power rating, the conventional IPT power converters provide low output power capability which is a drawback in case of high power applications such as high speed EV charging systems, industrial manufacturing chains. A number of multilevel converter topologies have been introduced for IPT systems to improve the power transmission capability. In addition, the proposed phase shift modulation strategies minimized the converter switching and conduction losses which are increased due to the addition switching devices.

Overview of IPT circuit topologies

2.1 Introduction

In IPT systems, a high frequency current is required to supply the primary winding. Therefore, a primary converter is designed to operate at high frequency, typically from 10 kHz to a few MHz [26]. In the secondary side, the picked up high frequency current is converted to a suitable form to feed the load via secondary side power converters. Therefore, the development of power converters is considered to be one of the most important aspects in IPT system design. In the development of IPT power converters, criteria such as the efficacy, system stability, power transmission capability, low cost and compactness need to be considered carefully.

This chapter presents an overview of high frequency power converters in IPT systems and their modulation methods. This work focuses on the power converters for primary windings which are the most importance in IPT systems. A review on the development of pick up converter topologies can be found in [27 – 29].

Besides power converters, in order for the reader to have a better basic knowledge of IPT technology, this chapter will provide a general overview of some other components of an IPT system.

Firstly, some fundamental components of IPT systems including magnetic coupling structures, power conversion types, switching techniques and compensation circuit configurations will be introduced.

Next, typical high frequency DC-AC power converters including voltage fed and current fed power converters and the relevant switching control methods will be presented. The benefits of these topologies are the simplicity in design and modulation. However, there are some disadvantages such as high output voltage distortion, hard switching issues, low voltage and power rating. Moreover, in case

of unavailability of DC power supply, an additional AC to DC power conversion is required at the front end.

In the situations where AC grids are available, direct matrix converters are proposed for exciting IPT systems. IPT matrix converters with different topologies and modulation strategies has been presented in [30 – 38]. These matrix converter topologies used in IPT systems reduce the number of power conversion stages required, thus cutting down the power electronic components, increasing overall efficiency, lowering cost and reducing physical size. However, the complexity in controlling and difficulty in filtering make the system less appealing.

Recently, in the efforts of increasing the power scalability of IPT systems to meet higher power requirement of industrial applications, multilevel converters with different topologies have been investigated [39 – 46]. Beside the power scalability, multilevel converters give more benefits in relation to the reduction of total harmonic distortion in the output voltages and currents, therefore lowering the power losses and increasing the overall efficiency. In most of these IPT systems, phase shift modulation is employed to obtain stair-case output voltage waveform to supply the primary winding.

In an alternative way of increasing power scalability, multiphase IPT systems have been presented in [55 – 62]. Present day three-phase IPT systems employ three-phase H-bridge converters to generate high frequency three-phase currents to excite the primary winding.

2.2 Fundamental components of IPT systems

2.2.1 Magnetic coupling structures

Magnetic coupling is an important part in IPT systems. A good coupling structure gives the benefits in term of high power transmission capability and low power losses by improving the magnetic field linkage between the primary side and secondary side coils/tracks (coupling coefficient) and reducing AC resistance of the windings [64]. Depending on the geometrical structure of the inductive windings, the coupling configurations can be classified into two types: distributed track systems and lumped coupling coil systems as shown in Figure 2.1 [29]. The distributed track systems comprise one primary track with multiple secondary

pickups which are employed in some industrial moving applications such as material handling, industrial manufacturing chains, and on-road electric vehicles charging. The lumped coupling coil systems employ one or more inductive coils in the primary coupled to one or more secondary inductive coils. This kind of coupling structure is more suitable for stationary systems such as biomedical implants, mobile device charging or electric vehicle charging at the stations.

In both of above IPT systems, magnetic cores can be employed to enhance the coupling coefficient between the primary and secondary windings and to reduce the generated Electromagnetic Interference (EMI). Up to date, ferrite core is the best choice for IPT systems due to its good features such as very low eddy current and hysteresis losses at high frequency applications [76]. Meanwhile litz wires are usually used to build inductive coils/tracks to reduce skin and proximity effect at high frequencies [77]. The design and optimization of inductive coils/tracks is one of the most interesting topics. However, it is out of scope of this thesis. A comprehensive research in this area can be found in [63 – 75].

(a) distributed track systems

(b) lumped coupling coil systems

Figure 2.1: The coupling configuration of IPT systems

2.2.2 Primary power conversion types

As mentioned earlier, primary power converters play a very important role in an IPT system. It generates high frequency current to supply the primary side resonant circuit. The primary converters are expected to operate at high frequency with low losses, low harmonic distortion in output voltage and current, compact in size and low cost. Therefore, the design of primary power converters becomes a very

interesting topic. Although many power converter topologies have been proposed for IPT systems with each topology having numerous advantages, there are still drawbacks in each topology that can be further improved. This section gives an overview of primary power conversion types for IPT systems that will be discussed in more details in the subsequent sections.

There are several ways to classify the power converters in IPT systems. In terms of switching conditions, there are linear amplifiers and switch mode power converters [78 – 79]. In terms of power supply conditions, there are voltage source and current source power converters. In terms of power conversion stages, there are two power conversion stages (AC-DC-AC) and single conversion stage direct matrix converters (AC-AC). Depending on the particular application, most of the aforementioned power converters can be potentially employed in IPT systems.

Although linear amplifier is known to be high power losses converter, it is still employed in low power, very high frequency systems where the power efficiency is not a serious concern such as biomedical implants [80].

In contrast to linear amplifiers, switch mode power converters generate high frequency output current by fully turning ON and OFF switches in a methodical manner. Therefore, it can achieve high efficiency with a wide range of output power. Nowadays, switch mode converters are employed in most of industrial applications, including inductive power transmission. The switch mode power converters employed in IPT systems are in the form of class E converters for low power applications [81 – 87], voltage and current fed H-bridge and push-pull converters [101 – 110], direct matrix converters [30 – 38] and multilevel converters [39 – 46] for high power IPT applications. These topologies will be discussed in details in the subsequent sections.

2.2.3 Switching techniques

Primary converters are expected to generate high frequency output voltages with as low total harmonic distortion (THD) as possible so that the power losses through the converters are minimized. Most of applications employed switch mode power converters. Usually, the pulse width modulation (PWM) switching technique is employed in most switch mode power converters to obtain an output voltage with

low THD. The frequency of output voltage varies from several Hz to hundreds of Hz, e.g., inverters for induction motors, while the switching frequency of power switches is equal to the modulation carrier frequency which is much higher than the output voltage frequency. Therefore, conventional PWM switching method becomes unsuitable for IPT power converters which are required to generate high frequency output voltage. In this situation, phase shift modulation becomes a good switching technique for most of IPT power converters; such as, H-bridge voltage-source and current-source converters, direct matrix converters and multilevel converters [88 – 96, 111 – 113]. Typically, a power converter which employs phase shift modulation method generates stair-case output voltage waveform as shown in Figure 2.2.

Figure 2.2: H-bridge converter with phase shift modulation and output stair-case voltage waveform.

The output power can be controlled by adjusting pulse width of the output voltage via phase shift angle. The benefit of phase shift modulation is that the switching frequency of switching devices is equal to the output voltage frequency. An important aspect that needs to be considered when applying phase shift modulation to IPT power converters is the hard switching issue. In some cases, an adjustable DC power supply can be used to control the output power without the need of phase shift modulation. Furthermore, the combination between the primary-side converters to generate square voltage waveform and secondary-side converters to adjust the output power can be an alternative [146]. The benefit of this configuration is that it prevents hard switching issue in the primary converters. However, the additional power conversion stage lowers the efficiency and reliability of the whole system.

In addition, the variable frequency control method is also employed to deal with the hard switching problems [17]. In this situation, the switching frequency is set around the resonant frequency so that the transmission power capability is maximized with as low power losses as possible.

In some multilevel converter topologies, selective harmonic elimination (SHE) is used to eliminate the high order harmonic components of the output voltage [42, 44 – 45].

2.2.4 Compensation circuit configurations

The compensation circuits are used to increase power transmission capability and to reduce VA rating of the IPT systems [17]. Capacitive resonant circuits are proposed in most of IPT systems due to their simplicity and low power losses. The compensation circuits also act as harmonic filters for IPT systems; thereby improving the quality of the output voltage and current. Several resonant topologies have been investigated in literature. Figure 2.3 shows different single phase resonant circuit topologies with series, parallel and the combination between series and parallel connected configurations. Figure 2.3 (a), (b), (c) and (d) show four basic resonant circuit configurations: Series-Series (SS), Series-Parallel (SP), Parallel-Series (PS), and Parallel-Parallel (PP) [129 – 136], whereas Figure 2.3 (e), (f), (g) and (h) show some other combined circuit configurations such as inductor – capacitor – inductor (LCL) [137 – 144], capacitor – inductor – capacitor – inductor (CLCL) [153], capacitor – capacitor – inductor (CCL) [147], and inductor – capacitor – capacitor – inductor (LCCL) [148 – 150]. Each of these configurations has different characteristics which are suitable for diverse situations.

Among four basic compensation circuit configurations, series network is usually used to compensate voltage source converter fed circuit in the primary side of IPT systems due to its desirable features [136]; meanwhile, parallel network is used with current source converters.

LCCL structure is proposed with the efforts to get small current switching in order to minimize switching losses of the converters. The complicated mathematical modeling of LCCL configuration has been investigated in [148 – 150] taking into account the influence of additional inductors of the coupling windings.

LCL compensation circuits are investigated with some advantages which can be applied for distributed track IPT systems [143, 144, 146]. However, the input current quality of LCL network which is supplied by fixed frequency phase shift modulation is very poor. Therefore, the switching stress and power losses can be high.

By adding a capacitor into the LCL topology, CLCL configuration is constructed as shown in Figure 2.3 (f) with reduced input current harmonic distortion [153]. Although input current quality in this work is lightly improved, the additional passive component makes the systems more complicated and more expensive.

CCL structure is proposed in [147] with some good features such as having better performance in transfer distance and efficiency, and having ability in circuit parameter optimization.

Recently, multiphase IPT systems have been proposed in [55 – 62]. Similar to single phase IPT systems, the resonant circuit configurations for three-phase IPT systems can be shown in Figure 2.4 and Figure 2.5, which are classified into two types: wye and delta connections.

Because the magnetic coupling of the series-parallel mixed resonant coupling structures (LCL, CCL, LCCL) for multiphase is very complicated, as of today, the development of multiphase IPT systems is just carried out with simple series and parallel compensated configurations [55, 56].

Figure 2.3: Single phase resonant circuit configurations.

Figure 2.4: Three-phase resonant circuit with wye connection (primary side)

Figure 2.5: Three-phase resonant circuit with delta connection (primary side)

2.3 Typical high frequency DC/AC power converters

A typical IPT system excited by DC/AC power converters in the primary side is as shown in Figure 2.6. The input power can be supplied directly from a DC source such as PV modules or from a rectified AC grid.

A high frequency current will be generated from DC power supply by high frequency DC/AC power converters. Depending on the circuit configuration, the DC power converters for IPT systems can be classified into two groups: voltage fed converters and current fed converters.

Figure 2.6: A block diagram of IPT power modules.

2.3.1 Current fed DC/AC IPT power converters

Two typical current fed DC/AC power converter topologies for IPT systems are shown in Figure 2.7 [17]. DC inductors are connected in series between the DC power source and power converters to maintain the input current to be nearly constant at high frequency of operation. Instead of employing four switches as in H-bridge converter topology, the push-pull topology employs a splitting transformer to keep the continuous path of current flowing. The push-pull current fed converter has the benefits in terms of reducing the number of switching devices. However, the additional transformer increases the cost, physical size and lowers the efficiency of the whole system.

Usually, a parallel compensation circuit is used in the primary resonant side of current source converter as shown in Figure 2.7. (a). Fixed frequency switching method can be applied in this model. A symmetrical switching diagram for H-bridge and push pull converters is shown in Figure 2.8. If the output current is maintained to be in phase with the output square wave current, the zero current

switching (ZCS) can be achieved as shown in Figure 2.8. This control strategy is simple in configuration and switching principle. However, the output power cannot be controlled by this switching method. In this situation, an additional front end DC power regulator needs to be employed in the secondary side to regulate the output current and power.

Figure 2.7: Current source converters for IPT systems.

Figure 2.8: Symmetrical fixed frequency switching diagram of current source converters.

The phase shift modulation can also be employed in the H-bridge converter to control the magnitude of the output current of the converter; thereby output power can be regulated without the need of another front-end power regulation stage. The switching diagram of the phase shift modulation for the aforementioned converter is shown in Figure 2.9. The hard switching occurs when two upper switches turned on or two lower switches turned off. Assume that the input inductor is large enough so

that the input current is nearly constant. Thereby, the output current has stair-case waveform with the pulse width exactly equal to the phase shift angle between two legs of the converter. By this way, the output power can be achieved from zero to maximum amount when the phase shift angle changes from 0 to π .

An important notice from this switching method is that in the period of overlapping of (S_1 and S_2) or (S_3 and S_4), the primary resonant circuit is free of oscillation, meanwhile the input current will go through the input inductor and a pair of switches in a leg which is switched on. This feature makes the system less efficient due to the power losses through the input inductor.

Figure 2.9: Phase shift modulation switching diagram of H-bridge current source converter.

2.3.2 Voltage-fed DC/AC IPT power converters

The voltage-fed H-bridge and half-bridge power converters are shown in Figure 2.10. The half-bridge converter employs two capacitors in a leg to provide the neutral point of input DC voltage. Compared to the H-bridge converter, a half-bridge converter is simpler in controlling and reduces the number of switching devices, therefore with reduced switching losses. However, the output voltage of the neutral point between two capacitors is normally unbalanced during the switching process of two switches. Another disadvantage of the half-bridge topology is that the additional passive components of half-bridge converter make it more bulky and less reliable. In addition, because the half-bridge converter can only generate the

square waveform with the amplitude of $\pm V_{DC}/2$, the output power capability is much lower than that of the H-bridge converter topology. Therefore, in practice, H-bridge converters are preferable in most of applications.

Figure 2.10: Voltage source converters for IPT systems.

Unlike current source power converters which employ a large inductor in the input side to keep the input current constant, the voltage source configurations comprise a DC power supply connected directly to the power converter. Therefore, the output high frequency voltage must be connected to either a purely resistive load or a current sourced type load. In IPT systems, except for parallel compensation structure, the other topologies including series, LCL, CLCL, CCL, LCCL can be applied. The development of these compensation topologies for IPT systems has been mentioned in the previous section.

Similar to current fed converters, in order to get a high frequency output voltage, the symmetrical inverting and phase shift modulation are usually employed. The output voltage and current waveforms with respect to (w.r.t.) two switching methods are shown in Figure 2.11 and Figure 2.12 (assuming that the series compensation circuits are employed). The root mean square (RMS) value of output voltage w.r.t. the phase shift modulation method can be expressed as follows,

$$V = \frac{2\sqrt{2}}{\pi} V_{DC} \sin\left(\frac{\varphi}{2}\right) \quad (2.1)$$

where φ is the phase shift angle between two legs of the converter, $\varphi = 0 \div \pi$

There are four hard switching states and four soft switching states within an output voltage cycle. To achieve soft switching for this control method, several

auxiliary circuits have been proposed with the addition of passive components such as capacitors, inductors and diodes. The systems therefore become more complicated. Besides, variable frequency switching control method can also be applied to achieve ZCS [159 – 164].

Figure 2.11: Symmetrical fixed frequency switching diagram of voltage source converters.

Figure 2.12: Phase shift modulation switching diagram of H-bridge voltage source converter.

In order to control the output power, another power converter can also be used in the secondary side. If an H-bridge converter is employed, the system becomes bidirectional as shown in Figure 2.13. The power amount, power direction and input power factor can be controlled by adjusting the phase shift angle between two

converters [88 – 91]. More analyses and discussions about this kind of power topology generations will be presented in chapter 4 of this thesis.

Figure 2.13: Typical BIPT system with series-series compensation.

2.4 Direct AC/AC matrix converters

The aforementioned power converter generation is applicable if DC power supplies are available. However, as the DC source is not ubiquitous, an additional AC/DC power conversion stage is usually used to convert AC from a grid supply to DC source. Therefore, the primary-side of IPT systems comprises of two power conversion stages which result in bulky and inefficient system. A direct AC/AC matrix converter is a potential solution to convert power directly from low frequency AC grid to high frequency current which is required by IPT systems.

Figure 2.14: Three-phase to single phase matrix converter for IPT systems

Several matrix converter topologies have been proposed for IPT systems. The development of matrix converter topologies for IPT systems have been firstly introduced in [32]. This paper presented a three-phase to single phase matrix

converter topology to convert power directly from a low frequency AC grid source to high frequency current in order to supply IPT systems. The proposed concept is shown in Figure 2.14.

The high frequency output voltage is synthesized from the boundary of phase to phase input voltage source as shown in Figure 2.15.

Figure 2.15: Input and output profiles of three-phase to single phase matrix converter with phase shift modulation.

Figure 2.16: Output voltage and current waveforms.

To control the output voltage amplitude, phase shift modulation was employed with the output profiles as shown in Figure 2.15 and Figure 2.16.

A similar three-phase to single phase matrix converter topology was presented in [38] in 2008 with the advantages in two step commutation method. In addition, the combination of output high frequency and low frequency selective harmonic pulse patterns to eliminate high order harmonic components of input current is also another advantage of the presented work.

Figure 2.17: Single phase direct AC/AC matrix converter for IPT systems in [30]

Figure 2.18: Switching profiles and output current waveform of the direct AC/AC converter presented in [30].

A single phase direct AC/AC converter as shown in Figure 2.17 is proposed for IPT systems in [184]. The power converter operates at high frequency by natural

circuit oscillation and discrete energy injection control. A variable frequency switching control strategy is employed to achieve ZCS. The output voltage and current waveforms are shown in Figure 2.18.

This topology has the advantage in terms of minimizing the number of switching devices; thereby lowering the cost and physical size. Besides, switching losses are also minimized by ZCS control strategy. However, as the complexity of switching procedures, it is difficult to implement the real time controller. Therefore, although the proposed concept was first introduced in 2006 [184] with simulation results, only in 2012, the comprehensive control scheme was presented with experimental results in [30].

In 2010, a BIPT system which comprises a single phase matrix converter in the primary side was proposed as shown in Figure 2.19.

As the power can be transferred bidirectionally, the proposed topology is suitable for both electric vehicles (EVs) charging and vehicle to grid (V2G). Compared to the direct AC/AC converter topology that was presented above, this matrix converter concept employs more switching devices (four bidirectional switches). However, the switching method is as simple as conventional H-bridge converter. Phase shift modulation can be used to control the output voltage amplitude. The proposed concept has been mathematical modeled and simulated using Matlab/Simulink.

Figure 2.19: Single phase matrix converter based BIPT system [34]

2.5 Multiphase and multilevel converters IPT systems

2.5.1 Multiphase IPT systems

Multiphase IPT systems are investigated to improve the power capability and system efficiency. In addition, the investigation of multiphase IPT systems provides several benefits which are restrictions in single phase systems. First, it overcomes the alignment issues between the transmitter tracks and pick up windings to obtain maximum power transfer, especially in the situations of moving objects such as moving vehicles on street, industrial manufacture chains, self-propelled robots, etc. [55]. Second, in vehicle online charging systems, it provides continuous power transfer with or without onboard batteries during charging along the roadways [55]. Third, when the three phase ac output voltage is rectified to dc voltage, the dc voltage ripple will be equal to one third of that of single phase IPT systems. Thus, the dc filter capacitor volume will be significantly reduced [56].

Several multiphase IPT systems have been reported [55 – 62]. The first three-phase IPT system is introduced in [55] in 2007 as shown in Figure 2.20. It comprises three separate single-phase isolating transformers, which are parts of the LCL network.

Figure 2.20: Three-phase IPT system presented in [55].

Figure 2.21: Phase shift control signals with notches and pulses [55].

The system employs three-phase full-bridge inverter to excite the primary three-phase resonance circuit. Three-phase six-step pulse width modulation with additional notches and pulses in each square waveform control signal is employed to control the magnitude of output ac voltages as shown in Figure 2.21. The system is then verified by the experimental prototype with track current up to 40A/phase at frequency of 38.4 kHz. The presented work shows the feasibility of the three-phase IPT systems. However, a comprehensive study on the behavior of three-phase magnetic coupling is still not sufficiently explored.

In 2011, the complicated mutual inductances of the three-phase IPT systems have been mathematically modeled in [56]. Figure 2.22 shows the winding configuration prototype and Figure 2.23 shows the equivalent circuit for modeling. Three-phase state space model has been investigated for further analyses.

Although these multiphase IPT systems provide many advantages compared to single phase IPT systems, the complication in magnetic coupling among the windings makes it difficult to have a comprehensive study and improve the power scalability of the system. This is the reason why all the presented multiphase systems were verified under very low power level. Anyway, these multiphase systems have drawn a feasible picture for future high power IPT generations.

Figure 2.22: Three-phase IPT charging prototype presented in [56].

Figure 2.23: Equivalent circuit configuration for a three-phase IPT system.

2.5.2 Multilevel converters IPT systems

Many efforts have been made on the development of a high power high efficiency power converter for high power IPT systems [39 - 46]. The development of multilevel converters for IPT systems has been firstly proposed in [39]. A four-level Marx inverter has been employed to excite the primary resonant tank as shown in Figure 2.24.

Figure 2.24: Four-level Marx inverter presented in [39].

Later, in 2014, several advanced cascaded multilevel converters with reduced number of power switches for BIPT systems have been developed as shown in Figure 2.25 [42, 44].

Figure 2.25: BIPT systems presented in [42] and [44]

The presented converters have been compared with the conventional converters in relation to the number of power switches, cost, output voltage and current harmonic distortions, input power factor, power losses and system efficiency to demonstrate the benefits of multilevel converter generations in IPT systems.

These systems pioneered the development of high power IPT systems based on multilevel converter topologies.

2.6 Conclusions

In this chapter, a general overview of IPT circuit topologies was presented with description of fundamental components of IPT systems comprising magnetic coupling structures, resonant compensation circuits and power converters. Different kinds of IPT power converter topologies such as linear amplifiers, class E converters, voltage source converters, current source converters, direct matrix converters, multilevel converters and multiphase IPT systems together with switching control strategies were reviewed with general comparisons. Although each proposed topology and switching method were proven to give different advantages, they have their own drawbacks that need to be improved. Thus, the development of novel high frequency power converters along with control strategies is still being carried out.

The next chapter will present a novel matrix converter topology with reduced number of power switches and power losses. The investigation of optimal efficiency control strategy for BIPT systems will be presented in chapter 4. In chapter 5, different feasible multilevel converter topologies for IPT systems will be presented. The last chapter will give the conclusions and discussions about future works.

A Matrix Converter Topology for IPT Systems

3.1 Introduction

IPT systems have been gaining popularity as they offer numerous advantages over conventional wired power transfer systems in relation to convenience, safety, isolation, operation in hostile conditions and flexibility. Consequently, there is an increased demand for IPT powered applications, such as transportation [26, 114, 168], mobile devices charging, lighting, material handling, mobile control robots [115 – 123], bio-medical implants [124 – 128], etc.

Therefore, development of efficient IPT systems has been receiving increased attention as many attempts have recently been made to improve the overall efficiency. The overall efficiency of an IPT system largely depends on the losses that incur in coupling coils and converters. In case of the former, the losses can be minimized by proper magnetic circuit and coil winding design, whereas in the later, with the reduction of number of conversion stages, switching and conduction losses can be reduced.

A typical IPT system is shown in Figure 3.1 [92, 113]. The primary side comprises an AC source, a rectifier, an inverter and a resonant network to compensate the primary coil inductance. The secondary, mutually coupled to the primary through M , employs a resonant compensation network and a power regulator, which can be a rectifier in the simplest form. As evident, two power conversion stages are invariably required on the primary side to generate a high frequency current in the primary coil. In addition, a DC-link capacitor is also used

to reduce the voltage ripple. Consequently, the system suffers from disadvantages such as lower reliability, increase in power losses, cost and overall physical size.

A matrix converter can be employed to reduce the number of conversion stages and, thereby, to alleviate some of the above disadvantages [169]. The matrix converter concept facilitates the synthesis of a high frequency voltage or current directly from a low frequency supply source, and is ideal for generating a high frequency current, which is required for IPT systems, directly from the supply mains.

There are several matrix converter topologies that have been proposed to deliver the power from line voltage to the primary side inductive coils of IPT systems [30 – 38]. This chapter presents a matrix converter based IPT system that generates a high frequency current with a reduced number of semiconductor switches. Therefore, the power loss of the proposed power converter can be minimized compared to that of a conventional converter topology as shown in the later section in this chapter. Furthermore, SiC MOSFETs are used in the proposed matrix converter as they exhibit efficient performance with reduced power losses. The proposed topology transforms 50-Hz 3-phase utility supply to a single-phase high frequency supply which in turn can be used to directly excite the primary side resonant network of an IPT system. Figure 3.2 shows the IPT system which is driven by the proposed matrix converter based topology. Series compensation is provided for both sides of the IPT system.

Figure 3.1: A typical IPT system with AC power supply.

Figure 3.2: Proposed Matrix Converter based IPT System.

3.2 Topology description and analysis

3.2.1 The proposed matrix converter and commutation strategy

In the proposed topology, a three-phase to single-phase matrix converter with only 4 bidirectional switches is employed to produce a high frequency current in the primary side series resonant tank. The fourth switch is added to control the resonant energy through a zero-voltage state across the resonant tank or v_p . The switching strategy is described in Figure 3.3.

The basic switching rule is that only one bidirectional switch is switched on at any instant. The basic 4-step commutation is employed in this case. So the output voltage will be either equal to zero, or equal to one of three-phase voltages. In case the output voltage is equal to the phase voltage, it should be switched to the phase which is on the boundary of upper and lower voltage profiles as shown in Figure 3.3 (a), i.e. maximum voltage v^+ or minimum voltage v^- . The zero voltage state is controlled by switching on S_4 .

By using the phase shift modulation method [88 – 91, 111 – 113], the output voltage will have a waveform illustrated in Figure 3.3(f). To analyze the fundamental component and higher order harmonics of output voltage v_p , let us define S_x , S_y , v^+ , v^- , f^+ and f^- as shown in Figure 3.3. S_x and S_y are the

square wave signals with resonant frequency and S_y is shifted from S_x by the phase shift angle.

Figure 3.3: Commutation strategy with phase shift modulation method:

- (a) phase input voltage with boundary voltages highlighted;
- (b) S_x , (c) S_y : phase shift control signals;
- (d) f^+ , (e) f^- : positive and negative side control signals;
- (f) primary voltage of resonant load (v_p).

3.2.2 Steady-state harmonic analysis

Assume that the three-phase input voltages are given as follows

$$\begin{aligned}
 v_a(\omega_0 t) &= V_I \sqrt{2} \sin(\omega_0 t) \\
 v_b(\omega_0 t) &= V_I \sqrt{2} \sin(\omega_0 t - \frac{2\pi}{3}) \\
 v_c(\omega_0 t) &= V_I \sqrt{2} \sin(\omega_0 t + \frac{2\pi}{3})
 \end{aligned} \tag{3.1}$$

where ω_0 and V_I are the AC angular frequency and the RMS value of input phase voltages, respectively.

The positive and negative components of boundary input voltage are respectively given by

$$v^+(\omega_0 t) = \begin{cases} V_I \sqrt{2} \sin(\omega_0 t) & \frac{\pi}{6} \leq \omega_0 t \leq \frac{5\pi}{6} \\ v^+(\omega_0 t + \frac{k2\pi}{3}) & , \text{otherwise}; (k \in \mathbb{Z}) \end{cases} \quad (3.2)$$

and

$$v^-(\omega_0 t) = -v^+(\omega_0 t + \frac{2\pi}{3}) \quad (3.3)$$

And using Fourier expansion,

$$v^+(t) = \frac{3\sqrt{6}V_I}{2\pi} \left[1 - \sum_{n=1}^{+\infty} \frac{2}{9n^2 - 1} \cos(3n\omega_0 t - \frac{n\pi}{2}) \right] \quad (3.4)$$

and

$$v^-(t) = -\frac{3\sqrt{6}V_I}{2\pi} \left[1 - \sum_{n=1}^{+\infty} \frac{2^* (-1)^n}{9n^2 - 1} \cos(3n\omega_0 t - \frac{n\pi}{2}) \right] \quad (3.5)$$

Let us define the frequency of square wave for phase shift method be f_T . Therefore f_T is also the resonant frequency in both primary and pick up side of IPT system. So the resonant angular frequency will be

$$\omega_T = 2\pi f_T = \frac{1}{\sqrt{L_1 C_1}} = \frac{1}{\sqrt{L_2 C_2}} \quad (3.6)$$

The phase shift pulse patterns f^+ and f^- as shown in Figure 3.3(d) and Figure 3.3(e) are respectively given by

$$f^+(t) = \frac{\varphi}{2\pi} + \sum_{n=1}^{+\infty} \left[\frac{2}{n\pi} \cos(n\omega_T t - \frac{n\varphi}{2}) \sin(\frac{n\varphi}{2}) \right] \quad (3.7)$$

and

$$f^-(t) = \frac{\varphi}{2\pi} + \sum_{n=1}^{+\infty} \left[\frac{2^* (-1)^n}{n\pi} \cos(n\omega_T t - \frac{n\varphi}{2}) \sin(\frac{n\varphi}{2}) \right] \quad (3.8)$$

where φ is the phase shift angle, $0 \leq \varphi \leq \pi$.

Assume the voltage drop across the filters is neglected. The output voltage appears at the primary side of the resonant tank is given by,

$$v_p(t) = v^+(t)f^+(t) + v^-(t)f^-(t) \quad (3.9)$$

Substituting (3.4), (3.5), (3.7) and (3.8) into (3.9) we get,

$$v_p(t) = \frac{3\sqrt{6}V_I}{\pi} \left\{ \left[1 - \sum_{n=2,4,6\dots}^{\infty} \frac{2}{9n^2-1} \cos(3n\omega_0 t - \frac{n\pi}{2}) \right] \cdot \left[\sum_{k=1,3,5\dots}^{\infty} \frac{2}{k\pi} \cos(k\omega_r t - \frac{k\varphi}{2}) \sin(\frac{k\varphi}{2}) \right] - \left[\sum_{m=1,3,5\dots}^{\infty} \frac{2}{9m^2-1} \cos(3m\omega_0 t - \frac{m\pi}{2}) \right] \cdot \left[\frac{\varphi}{2\pi} + \sum_{l=2,4,6\dots}^{\infty} \frac{2}{l\pi} \cos(l\omega_r t - \frac{l\varphi}{2}) \sin(\frac{l\varphi}{2}) \right] \right\} \quad (3.10)$$

A simpler expression can be derived as follows by omitting the higher order harmonics in each expression (3.4), (3.5), (3.7) and (3.8).

$$v_p(t) \approx \frac{6\sqrt{6}V_I}{\pi^2} \sin(\frac{\varphi}{2}) \left\{ \frac{1}{35} \cos\left[(\omega_r + 6\omega_0)t - \frac{\varphi}{2} \right] + \frac{1}{35} \cos\left[(\omega_r - 6\omega_0)t - \frac{\varphi}{2} \right] \right\} + \frac{3\sqrt{6}V_I \cdot \varphi}{8\pi^2} \sin(3\omega_0 t) \quad (3.11)$$

Simplifying further we have,

$$v_p(t) \approx \frac{6\sqrt{6}V_I}{\pi^2} \sin(\frac{\varphi}{2}) \cos(\omega_r t - \frac{\varphi}{2}) \quad (3.12)$$

Table 3.1: Simulation parameters

Parameter	Symbol	Value	Unit
Input phase voltage (RMS)	V_I	50	V
Input frequency	f_o	50	Hz
Primary and secondary inductance	$L_1 = L_2$	242	μH
Primary and secondary equivalent AC resistance	$R_1 = R_2$	0.5	Ω
Primary and secondary compensator capacitance	$C_1 = C_2$	0.042	μF
Switching frequency	f_T	50	kHz
Coupling coefficient	k	0.2

Figure 3.4: FFT spectra of primary resonant tank voltage.

Figure 3.5: Primary and pick up voltages and currents when phase shift angle $\varphi = 4\pi/5$.

Figure 3.6: Instantaneous voltages and currents from primary and pickup sides when phase shift angle $\varphi = 4\pi/5$.

Figure 3.7: Primary and pick up voltages and currents when phase shift angle $\varphi = \pi/2$.

Figure 3.8: Instantaneous voltages and currents from primary and pickup sides when phase shift angle $\varphi = \pi/2$.

The FFT spectrum of this voltage is illustrated in Figure 3.4. The amplitude of beating harmonic components of v_p which are near the resonant angular frequency ω_r are much less than that corresponds to resonant frequency ω_r . The output voltage and current waveforms are shown in Figure 3.5 – Figure 3.8 with different phase shift angles. Series compensation is used in this simulation on both sides of IPT system with the parameters given in Table 3.1. It is obvious that the output current is nearly sinusoidal.

3.3 Power loss evaluation

One of the most important aspects of a power converter applied for IPT systems is the consideration of its power losses. The power losses of a converter consist of conduction losses and switching losses of diodes and switching devices such as MOSFETs or IGBTs. The power losses of a three-phase to three-phase 9 bidirectional switch matrix converter have been reported in [170 – 171]. This section presents the power losses of the proposed matrix converter with the use of an appropriate commutation method.

3.3.1 Conduction loss

In the proposed topology, the output current flows through the bidirectional switch which has one switch and one diode in any particular direction. Suppose that output current of matrix converter is sinusoidal and the RMS value is I_0 and as MOSFET is employed as switching devices, the conduction losses of MOSFET and DIODE are derived as follows.

Conduction loss of MOSFET

$$P_{cond_MOS} = \frac{1}{T_s} \int_0^{T_s} r_{DS} i_D^2 dt = \frac{1}{2\pi} \int_0^{2\pi} r_{DS} (I_0 \sqrt{2} |\sin x|)^2 dx = r_{DS} I_0^2 \quad (3.13)$$

where r_{DS} is the on-state resistance of MOSFET.

Conduction loss of DIODE

$$P_{cond_D} = \frac{1}{T_s} \int_0^{T_s} v_{AK}(t) i_C(t) dt = \frac{1}{2\pi} \int_0^{2\pi} (V_f + r_D I_0 \sqrt{2} |\sin x|) I_0 \sqrt{2} |\sin x| dx = \frac{2\sqrt{2}}{\pi} V_f I_0 + r_D I_0^2 \quad (3.14)$$

where v_{AK} , V_f and r_D are the anode-cathode voltage, threshold voltage and equivalent on-state resistance of diode, respectively.

3.3.2 Switching loss

To evaluate the switching losses, firstly let us introduce the commutation strategy proposed in this work. The most popular commutation method for matrix converter is the 4-step commutation. There are 2 types of 4-step commutation: 4-step current commutation and 4-step voltage commutation. For the first case, the sign of output current needs to be measured accurately. Since the load current in IPT system is high frequency current, it is difficult to track correctly the sign of load current due to the delay time of signals in the sensors and filters. In this situation, the 4-step commutation based on input voltage proves more advantageous. The basic principle of 4-step voltage commutation of the proposed matrix converter is shown in Figure 3.9, Figure 3.10 and Table 3.2.

Suppose that the load is resistive and all the components are under tuned conditions, and then the output voltage and current waveforms are in phase as shown in Figure 3.12. In this situation, the instantaneous switching current in all the switching is as follows,

$$I_{d_sw} = I_0 \sqrt{2} \cos\left(\frac{\varphi}{2}\right) \quad (3.15)$$

Table 3.2: Switching sequence for 4-step voltage commutation

	From phase X to phase D	From phase D to phase X
$V_x > 0$	$S_{D1}^{ON} \rightarrow S_{X1}^{OFF} \rightarrow S_{D2}^{ON} \rightarrow S_{X2}^{OFF}$	$S_{X2}^{ON} \rightarrow S_{D2}^{OFF} \rightarrow S_{X1}^{ON} \rightarrow S_{D1}^{OFF}$
$V_x < 0$	$S_{D2}^{ON} \rightarrow S_{X2}^{OFF} \rightarrow S_{D1}^{ON} \rightarrow S_{X1}^{OFF}$	$S_{X1}^{ON} \rightarrow S_{D1}^{OFF} \rightarrow S_{X2}^{ON} \rightarrow S_{D2}^{OFF}$

X: A, B, C

Figure 3.9: Single phase diagram for four-step voltage commutation.

Figure 3.10: Switching diagram of 4-step voltage commutation for $V_x > 0$ and switching from phase X to phase D.

Figure 3.11: Switching voltage diagram

Figure 3.12: Output voltage and current diagram

The voltage across each switch is the boundary profile of input phase voltage as seen in Figure 3.11 and the average voltage across each switch and diode is as follows,

$$V_{DS_AV} = \frac{3}{\pi} \int_0^{\pi/3} \sqrt{2}V_i \cos(x)dx = \frac{3\sqrt{6}}{2\pi} V_i \quad (3.16)$$

➤ Switching loss of MOSFET

The energy losses of MOSFET during the transition are

$$E_{SW_MOS} = \int_0^{\tau_{SW}} v_{ds}(t) \cdot i_d(t) dt \quad (3.17)$$

where $v_{ds}(t)$ and $i_d(t)$ are the drain-source voltage and source current of the MOSFET during the transition, respectively; τ_{SW} is the total switching time.

As seen in Figure 3.12, there are four switching states in each resonant cycle. As a result of the four-step commutation method, in any switching state, there are always two switches and two diodes: one turned-on and one turned-off in hard-switching mode. The remaining switches and diodes are in zero-current switching mode. Therefore, the switching power loss of MOSFET can be calculated as follows,

$$P_{SW_MOS} = V_{DS_AV} I_{d_SW} 4f_T \left(\frac{e_{SW_ON} + e_{SW_OFF}}{V_R I_R} \right) = \frac{12\sqrt{3}}{\pi} V_I I_0 f_T \cos\left(\frac{\varphi}{2}\right) \left(\frac{e_{SW_ON} + e_{SW_OFF}}{V_R I_R} \right) \quad (3.18)$$

where e_{SW_ON} and e_{SW_OFF} are the turn-on and turn-off energy losses of MOSFET; V_R and I_R are the reference drain-source voltage and source current of MOSFET, respectively, which are provided by the manufacturer.

➤ Switching loss of Diode

The turn-on loss of the diode can be neglected because it is very small compared to the reverse recovery loss of the diode when it is turned-off. With the identical derivation as MOSFET, the switching loss of the diode is given as follows,

$$P_{SW_D} = \frac{12\sqrt{3}}{\pi} V_I I_0 f_T \cos\left(\frac{\varphi}{2}\right) \frac{Q_{RR}}{I_{R_D}} \quad (3.19)$$

where Q_{RR} is the diode reverse recovery charge and I_{R_D} is the reference current of the diode, which is provided by the manufacturer.

Therefore, the total power loss of the proposed matrix converter is given as follows,

$$P_{LOSS_M} = P_{cond_MOS} + P_{cond_D} + P_{SW_MOS} + P_{SW_D} \quad (3.20)$$

3.3.3 Power loss function (PLF)

An equivalent circuit of the topology presented in Figure 3.2 is shown in Figure 3.13.

Figure 3.13: The equivalent circuit of the proposed series-series compensated IPT system

The equivalent circuit inductances are given by

$$\begin{aligned} L_a &= L_1 - M \\ L_b &= L_2 - M \end{aligned} \quad (3.21)$$

where L_a and L_b are the leakage inductances of the primary and the pickup windings respectively and the mutual inductance $M = k\sqrt{L_1 L_2}$ where k is the coupling factor between the two windings or inductances. It is obvious from Figure 3.13 that the impedance referred to the voltage source side is given by,

$$Z_1 = \frac{M^2 \omega^2}{j\omega L_2 - \frac{j}{\omega C_2} + R_2 + R_L} + j\omega L_1 - \frac{j}{\omega C_1} + R_1 \quad (3.22)$$

When IPT operates at the resonance angular frequency of ω_r ,

$$Z_1(\omega_r) = \frac{M^2 \omega_r^2}{R_2 + R_L} + R_1 \quad (3.23)$$

The input power from the primary side of inductive coils is given by,

$$P_1 = \frac{1}{2} \frac{V_{pm}^2}{Z_1(\omega_r)} = \frac{1}{2} \frac{V_{pm}^2}{\frac{M^2 \omega_r^2}{R_2 + R_L} + R_1} \quad (3.24)$$

The output current of matrix converter which is also the current of primary side winding is given by,

$$I_0 = \frac{|i_{pm}|}{\sqrt{2}} = \frac{V_{pm}}{\sqrt{2} Z_1} = \frac{V_{pm}/\sqrt{2}}{\frac{M^2 \omega_r^2}{R_2 + R_L} + R_1} \quad (3.25)$$

The output power is given by

$$P_{out} = \frac{1}{2} R_L |i_{sm}|^2 = \frac{1}{2} R_L \frac{V_{pm}^2}{\left(\omega_T M + \frac{R_1(R_2 + R_L)}{\omega_T M} \right)^2} \quad (3.26)$$

where V_{pm} is the amplitude of voltage $v_p(t)$, $V_{pm} = \frac{6\sqrt{6}V_L}{\pi^2} \sin(\frac{\varphi}{2})$

The efficiency of the IPT system through the coils with series-series compensated topology presented in Figure 3.13 is given by,

$$\eta_i = \frac{|i_L|^2 R_L}{|i_L|^2 R_L + |i_p|^2 R_1 + |i_s|^2 R_2} = \frac{1}{1 + \frac{R_2}{R_L} + \frac{R_1}{R_L} \left(\frac{R_2 + R_L}{\omega_T M} \right)^2} = f(M) \quad (3.27)$$

It is obvious that the efficiency is proportional to the mutual inductance. In another words, the increase in distance between two winding coils makes the efficiency smaller.

Let us define the power loss function of the power converter as,

$$PLF = \frac{\text{Proposed Converter Power Loss}}{\text{Converter Output Power}} = \frac{P_{LOSS_M}}{P_1} \quad (3.28)$$

Figure 3.14: Power loss function of the proposed converter.

Figure 3.14 shows the dependence of PLF on the phase shift angle and the coupling factor. It can be seen from Figure 3.14 that the power loss of the converter is inversely proportional to the phase shift angle and the coupling factor and hence to the mutual inductance. That means the power loss will be excessive when the converter operates at a very low power range (very small phase shift angle). Therefore, this IPT topology is suitable for high power range applications.

The overall efficiency of the system is given as follows,

$$\eta_{OVERALL} = \frac{\text{Load Power}}{\text{Three Phase Input Power}} = \frac{\eta_i}{1 + \text{PLF}} \quad (3.29)$$

Figure 3.15 shows the dependence of overall efficiency on the phase shift angle and the coupling factor.

Figure 3.15: Overall efficiency of the proposed system.

It can be seen that the overall efficiency is also a function of the phase shift angle and coupling factor. The efficiency decreases significantly with the decrease of phase shift angle and coupling factor. Therefore, it would be beneficial to operate the IPT system with a small air-gap and a high phase shift angle that corresponds to transfer of high power.

Figure 3.16: Power loss distribution of IPT system.

The power loss distribution of the proposed system is shown in Figure 3.16. For a given set of system parameters, the power losses through the coils play a major role in the total losses. The conduction loss ratio is almost independent on the phase shift angle and coupling factor while the switching loss ratio tends to zero as the phase shift angle tends to π .

3.4 Power loss comparison with the conventional IPT converter

Figure 3.17: Conventional power converter for IPT systems with 3 phase input voltages.

Assume that the conventional power converter feeds the primary side of inductive coils as shown in Figure 3.17 and all the components are adjusted to be the identical to the proposed system. Power losses of this topology are the sum of power losses of the rectifier and the H-bridge converter.

3.4.1 Power loss of the rectifier

Since the switching frequency of diodes in the rectifier is at grid frequency, the switching loss can be neglected. To calculate the conduction loss, let us consider the effect of DC link current. As seen in Figure 3.12 in the previous section, the H-bridge converter operates in the null state when the output voltage is zero and it causes I_{DC} to be zero and equal to the input side resonant current in the active state from $\frac{\pi - \varphi}{2} \rightarrow \frac{\pi + \varphi}{2}$ in the first half of switching cycle. Each diode will conduct DC current I_{DC} in one-third of input voltage cycle. Therefore, the conduction loss of rectifier can be calculated as follows,

$$\begin{aligned}
 P_{cond_D_RECT} &= \frac{6}{T_s} \int_0^{T_s} \frac{1}{3} (V_{f_R} + r_{D_R} i_{DC}(t)) i_{DC}(t) dt \\
 &= \frac{2}{\pi} \int_{\frac{\pi - \varphi}{2}}^{\frac{\pi + \varphi}{2}} (V_{f_R} + r_{D_R} I_0 \sqrt{2} \sin x) I_0 \sqrt{2} \sin x dx \\
 &= \frac{4\sqrt{2}}{\pi} V_{f_R} I_0 \sin\left(\frac{\varphi}{2}\right) + \frac{2}{\pi} r_{D_R} I_0^2 (\varphi + \sin \varphi)
 \end{aligned} \tag{3.30}$$

where V_{f_R} and r_{D_R} are the threshold voltage and equivalent on-state resistance of diodes in rectifier side, respectively.

3.4.2 Power loss of the H-bridge converter

Assume that the H-bridge converter employs the phase shift modulation method which is the most popular modulation method in literature [88 – 91] for driving the high frequency current to the primary side inductive coil. In this case, each body diode will conduct the input side resonant current from $0 \rightarrow \frac{\pi - \varphi}{2}$ and each MOSFET will conduct the rest of the half switching cycle. Therefore,

➤ Conduction loss of diode

$$\begin{aligned}
 P_{cond_D_H} &= \frac{4}{T_s} \int_0^{T_s} (V_f + r_D i_D(t)) i_D(t) dt = \frac{2}{\pi} \int_0^{\frac{\pi - \varphi}{2}} (V_f + r_D I_0 \sqrt{2} \sin x) I_0 \sqrt{2} \sin x dx \\
 &= \frac{2\sqrt{2}}{\pi} V_f I_0 \left[1 - \sin\left(\frac{\varphi}{2}\right) \right] + \frac{1}{\pi} r_D I_0^2 (\pi - \varphi - \sin \varphi)
 \end{aligned} \tag{3.31}$$

➤ **Conduction loss of MOSFET**

$$P_{cond_MOS_H} = \frac{4}{T_s} \int_0^{T_s} r_{DS} i_D^2(t) dt = \frac{2}{\pi} \int_{\frac{\pi-\varphi}{2}}^{\pi} r_{DS} (I_0 \sqrt{2} \sin x)^2 dx = \frac{1}{\pi} r_{DS} I_0^2 (\pi + \varphi + \sin \varphi) \quad (3.32)$$

➤ **Switching loss of H-bridge**

There are 2 switches turned on, 2 switches turned off, 2 diodes turned on and 2 diodes turned off in hard-switching mode during any resonant cycle. Similar to the proposed matrix converter topology, the switching current of the H-bridge converter is $I_0 \sqrt{2} \cos(\frac{\varphi}{2})$. Thus, the switching loss of the H-bridge converter is as follows,

$$P_{SW_H} = 2V_{DC} I_0 \sqrt{2} \cos(\frac{\varphi}{2}) f_T (\frac{e_{SW_ON} + e_{SW_OFF}}{V_R I_R} + \frac{Q_{RR}}{I_{R_D}}) \quad (3.33)$$

where $V_{DC} = \frac{3\sqrt{6}}{\pi} V_{I_H}$ is the output voltage of the full bridge rectifier; V_{I_H} is the RMS value of the input phase voltage of the rectifier. The peak value of the output voltage of H-bridge converter is given as follows [90],

$$V_{pm_H} = \frac{4}{\pi} V_{DC} \sin(\frac{\varphi}{2}) = \frac{12\sqrt{6}}{\pi^2} V_{I_H} \sin(\frac{\varphi}{2}) \quad (3.34)$$

It can be seen from (3.12) and (3.34) that the conventional converter produces an output voltage two times higher than the proposed matrix converter for a given input voltage. In order to compare the power losses between the converter topologies, let us keep the output power equal, and then the input voltage of the conventional converter becomes half of that of the proposed converter, i.e. $V_{I_H} = \frac{1}{2} V_I$. Therefore, the power loss of H-bridge converter will become,

$$P_{SW_H} = \frac{6\sqrt{3}}{\pi} V_I I_0 \cos(\frac{\varphi}{2}) f_T (\frac{e_{SW_ON} + e_{SW_OFF}}{V_R I_R} + \frac{Q_{RR}}{I_{R_D}}) \quad (3.35)$$

Finally, the total power loss of the conventional power converter is given as follows,

$$P_{LOSS_H} = P_{cond_D_RECT} + P_{cond_D_H} + P_{cond_MOS_H} + P_{SW_H} \quad (3.36)$$

In order to compare the power losses between two converter topologies, let us define a factor Γ as follows,

$$\Gamma(\varphi, M) = \frac{\text{Proposed Converter Power Loss}}{\text{Conventional Converter Power Loss}} = \frac{P_{LOSS_M}}{P_{LOSS_H}} \quad (3.37)$$

Figure 3.18: Power loss ratio of two power converter topologies.

Figure 3.18 shows the dependence of Γ upon to the phase shift angle and coupling factor. It can be seen from Figure 3.18 that the power loss of conventional converter is much higher than the proposed converter in case of large phase shift angles.

When the phase shift angle reaches to its maximum value (π), the power loss of conventional converter doubles the proposed converter's power loss. This proves the advantage of the proposed converter compared to the conventional converter.

3.5 Hardware implementation and experimental verification

In order to verify the feasibility of the proposed converter, an experiment system has been implemented using a combination of DSP for controlling and FPGA for commutating and high switching speed of SiC MOSFETs (CMF10120D) and ultrafast diodes (BYV410X-600). Table 3.3 shows the parameters of SiC MOSFETs and diodes used for the calculation of power losses in the above section. Compared to the traditional Silicon (Si) devices, Silicon Carbide (SiC) devices provide numerous advantages such as low power loss, high switching speed and low on-state resistance for high voltage, high current operation. Moreover, SiC semiconductors work better at higher temperatures compared to Si devices [171 – 178]. An identical set of parameters as given in Table 3.1 for the simulation part are used in the experiment setup.

In the literature, several geometrical structure of coupling coils have been investigated as mentioned in section 2.2.1 of this thesis. In this experiment, in order to simplify the verification, a flat circular spiral air-core coupling coil is built as shown in Figure 3.20. A comprehensive study on the coupling coils is out of scope of this thesis.

Litz wire is selected to build the coupling coils. Litz wire comprises of multiple individually insulated magnet strands twisted or braided into a uniform pattern in order to mitigate the skin and proximity effect at high frequency so that the ac losses will be reduced.

Figure 3.19: The schematic of hardware configuration.

Figure 3.20: Hardware prototype of the IPT system.

Table 3.3: Switching device parameters

	Parameter	Symbol	Value	Unit
CMF10120D	Drain-Source On-State Resistance	r_{DS}	0.26	Ω
	Turn-On Switching Loss	e_{SW_ON}	260	μJ
	Turn-Off Switching Loss	e_{SW_OFF}	120	μJ
	Reference Drain-Source Voltage	V_R	800	V
	Reference Source Current	I_R	10	A
BYV410X-600	Anode-Cathode On-State Resistance	$r_D = r_{D_R}$	0.08	Ω
	Forward Threshold Voltage	$V_f = V_{f_R}$	1.2	V
	Reverse Recovery Charge	Q_{RR}	28	nC
	Reference Anode-Cathode current	I_{R_D}	1	A

Figure 3.21: Input phase voltage and current.

Figure 3.22: FFT of input phase current.

Figure 3.23: Input phase voltage and output voltage of converter.

Figure 3.24: Converter output voltage and current and load current with phase shift angle $\phi = 4\pi/5$.

Figure 3.25: Turn on transition waveform

Figure 3.26: Turn off transition waveform

The output voltage of the converter is shown in Figure 3.23 for a half of input voltage cycle. The output voltage and current of the converter together with the load current are shown in Figure 3.24 to show that the output voltage and current of the converter are in phase as in Figure 3.12 and the load current leading the converter current by 90 degrees. Figure 3.25 and Figure 3.26 show the switching transition waveforms of drain-source voltage and source current of MOSFETs. It shows good commutation and there is no short circuit during the commutation. However, due to the limitation of PCB design techniques, the output voltage and current waveforms quality is poor as there are ringing during the switching.

An output power of 300 W has been delivered to the resistive load with an overall efficiency of ~85% when the coil gap is 12 cm. The overall efficiency in the experimental results is lower than that given in the analytical results. One of the major reasons is the poor switching quality as mentioned above. In addition, power losses in connecting wires, internal power source, input filter and compensation capacitor could lower the system efficiency. By optimizing the design further, the overall efficiency can be made even higher.

3.6 Conclusions

A SiC matrix converter based IPT system has been proposed to improve the overall efficiency of system through the reduction of power conversion stages. A series compensated IPT system has directly been excited from the supply mains to generate a high frequency current in the resonant tank. A mathematical model for the proposed system has also been presented with simulated results. The experimental results show the feasibility of the proposed power converter for IPT systems.

The power loss comparison of the proposed IPT converter and the conventional IPT converter demonstrated the benefit of the proposed converter in term of efficiency. However, in order to evaluate the overall aspects of a converter to apply for IPT systems, it is important to consider some other criteria such as the compactness, control complexity, EMI effect, protection issues and reliability. It is

obvious that, with the reduction of one power conversion stage, especially the DC link, the proposed topology has also the benefit in term of the compactness. For the rest of criteria mentioned above, it is necessary to improve the PCB design techniques and control algorithms to enhance the system performance.

An Efficiency Optimization Scheme for Bidirectional IPT Systems

4.1 Introduction

IPT is a well-established technology that allows power transfer from one system to another without any physical contacts. It offers numerous advantages such as convenience, safety, isolation, flexibility and free maintenance. Applications of IPT range from low power such as biomedical implant, mobile device charging to medium to high power such as home appliances, material handling systems, street lighting systems, and charging of electric vehicles (EVs) [114 – 128]. A typical Unidirectional IPT (UIPT) system, as shown in Figure 4.1, transfers power only from one side to the other. This system is beneficial in terms of simplicity in designing and controlling. However, the disadvantage of this system is the dependency of efficiency on the load as discussed in detail in this chapter. In addition, this system is not suitable for applications with regenerative energy capability or active loads. BIPT systems are considered to be the best choice to solve this issue. A typical BIPT system is shown in Figure 4.3. Several studies carried out recently demonstrated the benefits of BIPT systems, especially for EV charging systems [88 – 94, 179]. In literature, the developments of BIPT systems focus on analyzing the characterization of this system. A generalized steady state model of BIPT systems is developed in [89] with a LCL compensation circuit. Synchronization of the primary and pick up converters is proposed in [90] and a closed loop controller is implemented in [91]. These systems demonstrate the

applicability of BIPT systems. However, one of the most important aspects of BIPT systems, efficiency, has not been considered carefully in the literature.

As the aforementioned in the previous chapter, the power losses of an IPT system comprise of the losses that incur in converters and in coupling coils. In case of the former, many studies have been reported in relation to the development of converter topologies which are applicable for IPT systems [30 – 62, 180 – 184], including the proposed matrix converter that has been presented in chapter 3. In case of the latter, studies focus on the optimization of the proper magnetic circuit and coil winding designs [14, 147]. This chapter proposes a phase shift modulation to minimize the coil losses by selecting the proper phase shift angle of the primary and secondary side converters of the BIPT topology. In addition, the converter power losses are also analyzed to obtain the overall efficiency of BIPT systems. From these analyses, a solution for designing an efficient BIPT system is considered. The analysis is based on the series-series (SS) compensation circuit for the primary and pickup sides which is the most established compensation circuit model. A hardware laboratory prototype of 0.5-kW power rating BIPT system is implemented to investigate the behavior of the proposed concept. The experimental results show the efficacy of the proposed system.

Figure 4.1: Typical Unidirectional IPT system

4.2 Typical IPT system

4.2.1 Typical Unidirectional IPT system

A typical Unidirectional IPT (UIPT) system is shown in Figure 4.1. An H-bridge converter is employed in the primary side while an H-bridge diode rectifier is used to convert high frequency current from the secondary resonant circuit to DC load

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which is represented by a resistor. Inductive coil resistances of the primary and secondary windings are R_1 and R_2 respectively. The efficiency of the UIPT SS compensated circuit topology has been presented in Chapter 3 as follows,

$$\eta = \frac{|i_L|^2 R_{Leq}}{|i_L|^2 R_{Leq} + |i_p|^2 R_1 + |i_s|^2 R_2} = \frac{1}{1 + \frac{R_2}{R_{Leq}} + \frac{R_1}{R_{Leq}} \left(\frac{R_2 + R_{Leq}}{\omega_T M} \right)^2} = f(R_{Leq}) \quad (4.1)$$

The equivalent load resistance R_{Leq} can be expressed in terms of DC load resistance as follows,

$$R_{Leq} = \frac{8}{\pi^2} R_L \quad (4.2)$$

As η is a function of R_{Leq} it can be maximized by tuning the load resistance.

By using,

$$\frac{d\eta}{dR_{Leq}} = 0 \quad (4.3)$$

Maximum efficiency can be determined as follows,

$$\eta_{\max} = \frac{1}{1 + \frac{2R_1 R_2}{(\omega_T M)^2} + \frac{2R_1}{(\omega_T M)^2} \sqrt{R_2^2 + \frac{R_2}{R_1} (\omega_T M)^2}} \quad (4.4)$$

And

$$R_{Leq} \Big|_{\eta=\eta_{\max}} = \sqrt{R_2^2 + \frac{R_2}{R_1} (\omega_T M)^2} \quad (4.5)$$

Figure 4.2: The dependency of efficiency on the load resistance.

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Figure 4.2 shows the dependency of efficiency on the load resistance with the x-axis representing the normalized load resistance ($R_L/\omega M$) when the primary and secondary side coupling coil resistances are identical ($R_1 = R_2$). The simulation parameters are identical to the BIPT system simulation which is given in Table 4.1. It is obvious that the efficiency of the UIPT system significantly decreases with the change in load. This is the disadvantage of the UIPT systems which can be improved by BIPT systems as shown in the next section.

4.2.2 Typical BIPT system

Figure 4.3: A typical BIPT system.

A typical BIPT system as shown in Figure 4.3 (a) employs an H-bridge converter in the primary side to generate the high frequency current to the primary coupling coil/track from DC power supply. H-bridge converters are well-known as a solution for most of the IPT systems due to their simplicity and effectiveness. Another H-bridge converter employed in the secondary side with the output can be connected to active loads such as Electric Vehicles which are able to consume or regenerate power.

High frequency current is transferred wirelessly through the air-gap to the secondary winding which is coupled to the primary winding via loose electromagnetic coupling. The primary and the secondary circuits are identical with tuned series-series capacitive compensation circuits in order to operate the system at resonance frequency.

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The amount of power transmission is controlled by the phase shift angle of each converter (φ_1, φ_2) while the power direction is adjusted by the relative phase shift angle between the primary and secondary side converters (θ) as shown in Figure 4.3 (b).

In recent literature [89, 90], the phase shift angle of the primary side converter in a BIPT system is adjusted to control the limit of input current while the output power is controlled by adjusting the phase shift angle of the pickup side converter. This study presents a novel algorithm in obtaining the phase shift angle in order to minimize coil losses.

With the phase shift modulation strategy as shown in Figure 4.3 (b), the fundamental components of the input and output side voltages can be given as follows,

$$v_p(t) = \frac{4}{\pi} V_{DC1} \sin\left(\frac{\varphi_1}{2}\right) \cos(\omega_T t) \quad (4.6)$$

$$v_s(t) = \frac{4}{\pi} V_{DC2} \sin\left(\frac{\varphi_2}{2}\right) \cos(\omega_T t + \theta) \quad (4.7)$$

where φ_1 and φ_2 are the phase shift angles between the two legs of primary and secondary converters respectively, and θ is the phase shift angle between primary and secondary converters.

The voltages, v_{pi} and v_{si} , that are induced in the primary and pickup coils respectively, are given by,

$$v_{pi} = j\omega M i_2 \quad (4.8)$$

$$v_{si} = j\omega M i_1 \quad (4.9)$$

where M is the mutual inductance between the primary and secondary coils. The mutual inductance M is a function of the coupling coefficient between the two coils as follows,

$$M = k \sqrt{L_1 L_2} \quad (4.10)$$

where k is the coupling coefficient between primary and secondary windings.

From Figure 4.3 (a), the input and output currents are,

$$i_1 = \frac{v_p - v_{pi}}{R_1 + j\omega L_1 + \frac{1}{j\omega C_1}} \quad (4.11)$$

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$$i_2 = \frac{v_s - v_{si}}{R_2 + j\omega L_2 + \frac{1}{j\omega C_2}} \quad (4.12)$$

The fundamental components of v_p and v_s given in (4.11) and (4.12) can be represented in phasor form as follows,

$$v_p = V_{pm} \angle 0^\circ \text{ and } v_s = V_{sm} \angle \theta \quad (4.13)$$

IPT operates at the resonance angular frequency of ω_r , where $\omega_r^2 = 1/L_1 C_1 = 1/L_2 C_2$. Substituting (4.8), (4.9) and (4.13) into (4.11), and (4.12), the input and output currents are given respectively as follows,

$$i_1 = \frac{R_2 V_{pm} - j\omega M V_{sm} \angle \theta}{\omega^2 M^2 + R_1 R_2} \quad (4.14)$$

$$i_2 = \frac{-j\omega M V_{pm} + R_1 V_{sm} \angle \theta}{\omega^2 M^2 + R_1 R_2} \quad (4.15)$$

The primary and secondary active and reactive powers can be calculated by,

$$\begin{aligned} P_1 &= \frac{1}{2} \operatorname{Re} \{ v_p \cdot i_1^* \} \\ &= \frac{1}{2} \frac{V_{pm} (R_2 V_{pm} + \omega M V_{sm} \sin \theta)}{\omega^2 M^2 + R_1 R_2} \end{aligned} \quad (4.16)$$

$$\begin{aligned} Q_1 &= \frac{1}{2} \operatorname{Im} \{ v_p \cdot i_1^* \} \\ &= \frac{1}{2} \frac{\omega M V_{pm} V_{sm} \cos \theta}{\omega^2 M^2 + R_1 R_2} \end{aligned} \quad (4.17)$$

$$\begin{aligned} P_2 &= \frac{1}{2} \operatorname{Re} \{ v_s \cdot i_2^* \} \\ &= \frac{1}{2} \frac{V_{sm} (R_1 V_{sm} - \omega M V_{pm} \sin \theta)}{\omega^2 M^2 + R_1 R_2} \end{aligned} \quad (4.18)$$

$$\begin{aligned} Q_2 &= \frac{1}{2} \operatorname{Im} \{ v_s \cdot i_2^* \} \\ &= -\frac{1}{2} \frac{\omega M V_{pm} V_{sm} \cos \theta}{\omega^2 M^2 + R_1 R_2} \end{aligned} \quad (4.19)$$

The reactive power components at both sides of the system can be minimized by keeping the phase shift angle between the primary and the secondary side converters to be either $+90^\circ$ or -90° . When $\theta = +90^\circ$, power will be transferred from the primary side to the secondary side, while power flow is reversed when $\theta = -90^\circ$.

4.3 Proposed phase shift modulation strategies for IPT systems

With the descriptions of BIPT system in the previous section, when $\theta = +90^\circ$, power is delivered from primary to secondary side, and the forward direction efficiency is given as follows,

$$\eta_{for} = \frac{|P_2|}{P_1} = \frac{V_{sm}(\omega M V_{pm} - R_1 V_{sm})}{V_{pm}(\omega M V_{sm} + R_2 V_{pm})} = \frac{\xi(\omega M - R_1 \xi)}{\omega M \xi + R_2} \quad (4.20)$$

where $\xi = V_{sm}/V_{pm}$ is the ratio of secondary and primary output voltages of the converters.

Similarly, when $\theta = -90^\circ$, power is delivered from secondary to primary side and the efficiency in the reverse direction is as follows,

$$\eta_{rev} = \frac{|P_1|}{P_2} = \frac{V_{pm}(\omega M V_{sm} - R_2 V_{pm})}{V_{sm}(\omega M V_{pm} + R_1 V_{sm})} = \frac{\omega M \xi - R_2}{\xi(\omega M + R_1 \xi)} \quad (4.21)$$

It is obvious from (4.20) and (4.21) that the BIPT system efficiency is a function of the converter output voltage ratio. By differentiating the efficiency as a function of voltage ratio,

$$\frac{d\eta}{d\xi} = 0 \quad (4.22)$$

Figure 4.4: The dependency of coil efficiency on the converter voltages

(when $R_1 = R_2$)

Figure 4.5: A PI controller for the proposed BIPT system.

From (4.20) and (4.22), the optimum voltage ratio to get maximum efficiency in the forward direction is as follows,

$$R_1 \omega M \xi_{for_opt}^2 + 2R_1 R_2 \xi_{for_opt} - R_2 \omega M = 0 \quad (4.23)$$

$$\Leftrightarrow \xi_{for_opt} = \frac{-R_1 R_2 + \sqrt{R_1^2 R_2^2 + R_1 R_2 \omega^2 M^2}}{R_1 \omega M} \quad (4.24)$$

From (4.20) and (4.23), the maximum efficiency in the forward direction can be achieved as follows,

$$\begin{aligned} (23) \quad & \Leftrightarrow R_1 \xi_{for_opt} (\omega M \xi_{for_opt} + R_2) = R_2 (\omega M - R_1 \xi_{for_opt}) \\ & \Leftrightarrow \frac{\omega M - R_1 \xi_{for_opt}}{\omega M \xi_{for_opt} + R_2} = \frac{R_1}{R_2} \xi_{for_opt} \\ & \rightarrow \eta_{for_opt} = \frac{\xi_{for_opt} (\omega M - R_1 \xi_{for_opt})}{\omega M \xi_{for_opt} + R_2} = \frac{R_1}{R_2} \xi_{for_opt}^2 \\ & = 1 - \frac{2R_1 R_2}{R_1 R_2 + \sqrt{R_1^2 R_2^2 + R_1 R_2 \omega^2 M^2}} \end{aligned} \quad (4.25)$$

Similarly, the optimum voltage ratio and maximum efficiency in the reverse direction are as follows,

$$\xi_{rev_opt} = \frac{R_1 R_2 + \sqrt{R_1^2 R_2^2 + R_1 R_2 \omega^2 M^2}}{R_1 \omega M} \quad (4.26)$$

$$\eta_{rev_opt} = 1 - \frac{1}{1 + \frac{\omega^2 M^2}{2(R_1 R_2 + \sqrt{R_1^2 R_2^2 + R_1 R_2 \omega^2 M^2})}} = \eta_{for_opt} \quad (4.27)$$

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Assuming $(R_1, R_2) \ll \omega M$, which is satisfied in most situations, from (4.24) and (4.26), it is seen that ξ_{for_opt} and ξ_{rev_opt} can be approximated to be $\sqrt{R_2/R_1}$ which becomes unity if both inductive coils are identical.

Figure 4.4 shows the dependence of efficiency on the voltage ratio. It is obvious that the efficiency decreases with the change in voltage ratio. This property is similar to the UIPT system which is described in the previous section. However, the efficiency of the BIPT system can be maximized by adjusting the voltage ratio as shown in (4.24) or (4.26) and it is independent on the load.

From (4.6) and (4.7), the calculations of the phase shift angle used in the primary side and the pickup side converters are as follows,

$$\sin\left(\frac{\phi_1}{2}\right) = \frac{\pi}{4} \frac{V_{pm,ref}}{V_{DC1}} \quad (4.28)$$

$$\sin\left(\frac{\phi_2}{2}\right) = \frac{\pi}{4} \frac{V_{sm,ref}}{V_{DC2}} = \frac{\pi}{4} \frac{\xi V_{pm,ref}}{V_{DC2}} \quad (4.29)$$

where $V_{pm,ref}$ and $V_{sm,ref}$ are the desired output voltage amplitude of the primary and secondary H-bridge converters.

A closed loop controller can be applied for the given system to minimize coil losses by choosing an appropriate voltage ratio between the primary side and secondary side converters.

From (4.28) and (4.29), it is obvious that the following condition should be satisfied to ensure that the phase shift angles of the primary and secondary side converters are feasible,

$$V_{pm,ref} \leq \min\left(\frac{4V_{DC1}}{\pi}, \frac{4V_{DC2}}{\pi\xi_{opt}}\right) \quad (4.30)$$

where ξ_{opt} is calculated from either (4.24) or (4.26).

Figure 4.5 presents a PI controller with the phase shift generator block using the set of equations (4.28), (4.29) and either (4.24) or (4.26) to calculate the phase shift angle of both converters. Communication block is added to transfer the data between the primary and secondary-side converter controllers.

Figure 4.6: Output voltages and currents of converters.

The BIPT system in Figure 4.3 is simulated using the controller in Figure 4.5 with the parameters in Table 4.1. Figure 4.6 shows the simulation results of input and output voltages and currents of the given system during the control process. Figure 4.7 shows the power response of the controller with zero steady state error and zero over shooting. The efficiencies of the coupling coils are maintained at around 99% for almost full range of the output power as shown in Figure 4.8 in 2D and Figure 4.9 in 3D graphs.

Figure 4.7: Power response of the PI controller.

Figure 4.8: Calculated coil efficiency based on proposed control scheme.

Figure 4.9: The variation of efficiency with the voltage ratio and output power range.

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Table 4.1: Simulation and experimental parameters

Parameter	Simulation	Experiment
Coils inductance $L_1 = L_2$ (μH)	183	183.3
Equivalent AC resistances $R_1 = R_2$ (Ω)	0.1	0.14
Compensation capacitances $C_1 = C_2$ (μF)	0.06	0.06
Switching frequency f_T (kHz)	48	48
Coupling coefficient k	0.38	0.38
Power rating (kW)	6	0.5
Air gap (mm)	NA	60

4.4 Overall efficiency optimization analysis

In the previous section, the analyses of the coil loss minimization have been derived. The phase shift modulation has been investigated to optimize the efficiency in the resonant sides. However, the overall power losses of BIPT systems are not only coil losses but also power converter losses which comprises of switching losses and conduction losses. Therefore, an analysis of overall BIPT system efficiency is derived in this section.

In order to simplify the analyses, assuming that the power is transferred from the primary side to the secondary side and the relative phase shift angle θ is kept at $+90^\circ$ to minimize the reactive power. Since the system is symmetrical, in the case where power is transferred from secondary side to primary side, the derivation of efficiency of the proposed system is completely similar.

From (4.6) and (4.7), the amplitudes of primary and secondary converter output voltages are given as follows,

$$V_{pm} = \frac{4}{\pi} V_{DC1} \sin\left(\frac{\phi_1}{2}\right) \quad \text{and} \quad V_{sm} = \frac{4}{\pi} V_{DC2} \sin\left(\frac{\phi_2}{2}\right) \quad (4.31)$$

With the definition of ξ in the previous section, from (4.31) we have,

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$$V_{sm} = \frac{4}{\pi} V_{DC2} \sin\left(\frac{\varphi_2}{2}\right) = \xi V_{pm} = \frac{4\xi}{\pi} V_{DC1} \sin\left(\frac{\varphi_1}{2}\right) \quad (4.32)$$

From (4.32), the relationship between the phase shift angle of primary and secondary side converters can be derived as follows,

$$\varphi_2 = 2 \arcsin\left(\frac{\xi V_{DC1} \sin\left(\frac{\varphi_1}{2}\right)}{V_{DC2}}\right) \quad (4.33)$$

From (4.16), (4.18), (4.31) and (4.32), the primary and secondary active powers can be rewritten as follows,

$$P_1 = \frac{8}{\pi^2} V_{DC1}^2 \sin^2 \frac{\varphi_1}{2} \frac{R_2 + \omega M \xi}{\omega^2 M^2 + R_1 R_2} \quad (4.34)$$

$$|P_2| = \frac{8\xi}{\pi^2} V_{DC1}^2 \sin^2 \frac{\varphi_1}{2} \frac{\omega M - R_1 \xi}{\omega^2 M^2 + R_1 R_2} \quad (4.35)$$

Figure 4.10: Primary side H-bridge converter and output waveforms.

With the above assumptions, it is obvious from (4.14) and (4.15) that the output voltages and currents of both converters, (v_p, i_1) and (v_s, i_2) , are in phase. Suppose the effect of higher order harmonics is negligible then the output currents of converters are found to be sinusoidal as shown in Figure 4.10.

For the given assumption ($\theta = +90^\circ$), from (4.14) and (4.15), the RMS value of input and output converter currents are respectively given as follows,

$$I_1 = \frac{1}{\sqrt{2}} V_{pm} \frac{R_2 + \omega M \xi}{\omega^2 M^2 + R_1 R_2} = \frac{2\sqrt{2}}{\pi} V_{DC1} \sin \frac{\varphi_1}{2} \frac{R_2 + \omega M \xi}{\omega^2 M^2 + R_1 R_2} \quad (4.36)$$

$$I_2 = \frac{1}{\sqrt{2}} V_{pm} \frac{\omega M - \xi R_1}{\omega^2 M^2 + R_1 R_2} = \frac{2\sqrt{2}}{\pi} V_{DC1} \sin \frac{\varphi_1}{2} \frac{\omega M - \xi R_1}{\omega^2 M^2 + R_1 R_2} \quad (4.37)$$

To evaluate the overall efficiency of the system, let us calculate the power losses of the converters. The power losses of the IPT H-bridge converter consist of

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conduction losses and switching losses of diodes and switching devices which have been presented in chapter 3 from (3.31) to (3.33).

Therefore, the total power losses of primary-side and secondary-side converters are respectively given as follows,

$$\begin{aligned}
 P_{loss_p} = & \frac{2\sqrt{2}}{\pi} V_f I_1 \left[1 - \sin\left(\frac{\varphi_1}{2}\right) \right] + \frac{1}{\pi} r_D I_1^2 (\pi - \varphi_1 - \sin \varphi_1) \\
 & + \frac{1}{\pi} r_{DS} I_1^2 (\pi + \varphi_1 + \sin \varphi_1) \\
 & + 2V_{DC1} I_1 \sqrt{2} \cos\left(\frac{\varphi_1}{2}\right) f_T \left(\frac{e_{SW_ON} + e_{SW_OFF}}{V_R I_R} + \frac{Q_{RR}}{I_{R_D}} \right)
 \end{aligned} \tag{4.38}$$

And

$$\begin{aligned}
 P_{loss_s} = & \frac{2\sqrt{2}}{\pi} V_f I_2 \left[1 - \sin\left(\frac{\varphi_2}{2}\right) \right] + \frac{1}{\pi} r_D I_2^2 (\pi - \varphi_2 - \sin \varphi_2) \\
 & + \frac{1}{\pi} r_{DS} I_2^2 (\pi + \varphi_2 + \sin \varphi_2) \\
 & + 2V_{DC2} I_2 \sqrt{2} \cos\left(\frac{\varphi_2}{2}\right) f_T \left(\frac{e_{SW_ON} + e_{SW_OFF}}{V_R I_R} + \frac{Q_{RR}}{I_{R_D}} \right)
 \end{aligned} \tag{4.39}$$

The input power supply and output power can be calculated as follows,

$$P_{in} = P_1 + P_{loss_p} \tag{4.40}$$

$$P_{out} = |P_2| - P_{loss_s} \tag{4.41}$$

The overall efficiency of the given system is given as follows,

$$\eta_{overall} = \frac{P_{out}}{P_{in}} = f(V_{DC1}, V_{DC2}, \varphi_1, \xi) \tag{4.42}$$

It is obvious that the overall efficiency is a function of input and output DC voltages (V_{DC1} , V_{DC2}), voltage ratio (ξ), primary and secondary converter phase shift angles (φ_1 , φ_2) where secondary phase shift angle is a function of primary phase shift angle and voltage ratio as shown in (4.33).

Figure 4.11 shows the overall efficiency as a function of voltage ratio and output power with different input and output DC voltage ratios. Figure 4.12 shows the overall efficiency variation with each power surface clearly. Figure 4.13 shows the power loss percentage of the given BIPT system while Figure 4.14 and Figure 4.15

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show the variation of the overall optimal efficiency with output power and DC voltage ratio respectively. From Figure 4.11 - Figure 4.15, the following conclusions can be drawn:

1. The optimal efficiency takes place when the output AC voltage ratio $\xi = V_{DC2}/V_{DC1}$. From (4.33), it is obvious that the phase shift angles of both converters are identical. This is the concept to control the power converters to achieve the maximum efficiency.
2. With the same AC voltage ratio (ξ) and DC voltage ratio (V_{DC2}/V_{DC1}), the higher the output power requirement, the higher the overall efficiency that can be achieved for a given system.
3. For the given system, the converter loss percentage is much higher than the coil loss percentage. The coil losses are independent of output power level and DC voltage ratio while converter losses are largely dependent on the output power and DC voltage ratios.
4. Figure 4.14 presents the optimal overall efficiency as a function of output power and DC voltage ratio where AC voltage ratio is kept to be equal to DC voltage ratio as in the first conclusion above. Figure 4.15 presents the maximum overall efficiency boundary (maximum output power) as a function of DC voltage ratio. It is obvious from Figure 4.15 that the maximum overall efficiency of the given system takes place when $V_{DC1} = V_{DC2}$ ($\sim 97.5\%$). When the DC voltage ratio changes from 0.5 to 2, the overall efficiency decreases by an amount of 0.6%.

(a) $V_{DC2} = V_{DC1}$

(b) $V_{DC2} = 0.5 * V_{DC1}$

(c) $V_{DC2} = 2 * V_{DC1}$

Figure 4.11: Overall efficiency of BIPT system versus AC voltage ratio and output power for different DC voltage ratios.

(a) $V_{DC2} = V_{DC1}$

(b) $V_{DC2} = 0.5 * V_{DC1}$

(c) $V_{DC2} = 2 * V_{DC1}$

Figure 4.12: Overall efficiency of BIPT system versus AC voltage ratio for different output power surfaces $P_1 < P_2 < P_3$ and DC voltage ratios.

(a) $V_{DC2} = V_{DC1}$

(b) $V_{DC2} = 0.5 * V_{DC1}$

(c) $V_{DC2} = 2 * V_{DC1}$

Figure 4.13: Converter and coil loss percentage versus AC voltage ratio for different output power surfaces $P_1 < P_2 < P_3$ and DC voltage ratios.

Figure 4.14: Optimal overall efficiency versus output power for different DC voltage ratios.

Figure 4.15: Variation of optimal overall efficiency with the DC voltage ratio.

4.5 Hardware installation and experimental verification

In order to verify the efficacy of the proposed optimized phase shift modulation strategy, an experiment system has been implemented using FPGA Spartan 3E card for controlling and modulating high switching speed SiC MOSFETs (C2M0080-120D) and SiC diodes (CPW5-0650-Z030B) as shown in Figure 4.16. Table 4.2 shows the parameters of SiC MOSFETs and diodes used for the calculation of power losses in the previous section. The advantages of SiC devices have been mentioned in section 3.5 of the previous chapter.

In this experimental setup, in order to improve the coupling coefficient which is proportional to the overall efficiency of the IPT system, a circular E-core structure is built for the coupling coils as shown in Figure 4.16. The resonance circuit parameters as well as switching frequency are given in Table 4.1.

Figure 4.17 shows the voltage and current waveforms of the proposed BIPT system in the situation where both converter phase shift angles as well as both converter DC voltages are maintained identical to obtain the maximum efficiency. It is obvious that the voltages and currents in both the primary and secondary side are in phase and the current waveforms are almost sinusoidal as in the assumption in Figure 4.10. However, due to the effect of voltage spikes (as shown in Figure 4.17), high order harmonic distortion, conduction losses of connection wires, filter losses and the delay in the driver circuit, the experimental overall efficiency of the proposed BIPT system is lower than that in the simulation results. By optimizing the design further, the overall efficiency could be higher. Figure 4.18 and Figure 4.19 show the variation of the overall efficiency with the desired power and AC voltage ratio. These experimental results confirm the conclusions drawn in the previous section.

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Figure 4.16: Hardware prototype of the proposed BIPT system.

Table 4.2: Switching device parameters (SiC MOSFETs and Diodes)

	Parameter	Symbol	Value	Unit
C2M0080120D	Drain-Source On-State Resistance	R_{DS}	0.08	Ω
	Turn-On Switching Loss	e_{SW_ON}	290	μJ
	Turn-Off Switching Loss	e_{SW_OFF}	130	μJ
	Reference Drain-Source Voltage	V_R	800	V
	Reference Source Current	I_R	20	A
CPW5-0650-Z030B	Anode-Cathode On-State Resistance	R_D	0.02	Ω
	Forward Threshold Voltage	V_f	1.37	V
	Reverse Recovery Charge	Q_{RR}	0	nC
	Reference Anode-Cathode current	I_{R_D}	30	A

(a) delivering 450 W with 92% efficiency (b) delivering 300 W with 90% efficiency

Figure 4.17: Experimental results – Voltage and current waveforms.

Figure 4.18: Experiment results – Efficiency versus input power

Figure 4.19: Experiment results – Efficiency versus AC voltage ratio

4.6 Conclusions

In this chapter, an optimized phase shift modulation to minimize coil losses for BIPT systems has been theoretically analyzed and modelled using Matlab. In addition, a comprehensive study on the impact of power converters on the overall efficiency of the system is also presented. A closed loop controller is proposed to optimize the overall efficiency of the BIPT system. Theoretical results are presented in comparison to both simulations and measurements of a 0.5 kW prototype to show the benefits of the proposed concept. Results convincingly demonstrate the applicability of the proposed system offering high efficiency over a wide range of output power.

Multilevel Converter Topologies for Bidirectional IPT Systems

5.1 Introduction

In IPT systems, power converters are employed to generate high frequency currents, typically from 10 kHz to a few MHz. The primary side current is used to power the primary side inductive coils/tracks. Inductively coupled coils on the secondary side receive the high frequency currents. Thereafter, one or more power converters are employed to convert from high frequency current to DC or low frequency AC forms, depending on load requirements [29].

Nowadays, the application of IPT systems is widespread in low and medium power levels such as biomedical implants, transportation, material handling, street lighting, as well as in high power applications such as electric trains, industrial manufacture chains, which can reach hundreds of kilowatts to megawatts [114 – 128].

Many converter topologies have been proposed for IPT systems such as conventional H-bridge converter [88 – 96, 101], half-bridge converter [97 – 99], buck-boost converter [100] and direct matrix converter [30 – 38]. As there are some limitations in power switch characteristics in terms of voltage and current ratings, operating temperature, etc., the aforementioned conventional power converters become unsuitable for high power IPT generations.

Therefore, research on power converters used in high power IPT systems has become an important issue. Towards the aim of increasing power scalability and the overall efficiency, several attempts have been made in terms of using multiphase IPT systems [55 – 62], multi-input/multi-pickup IPT systems [154 – 155], parallel

connected power converter IPT system [167] and matching transformer power converter IPT system [114]. In [114], the power reaches up to 1 MW for high speed train applications. Generally, these kinds of power converter topologies have a common point as using multiple channel of coupling windings to increase the power level of IPT systems. However, some other aspects such as high voltage and current stress on the switching devices, high order harmonic distortion of the output voltage and current, the system compactness, have not been comprehensively considered yet. In addition, the need of multiple coupling windings structure gives rise more complicated and bulky in designing and implementing the systems.

Multilevel converter is known to be one of the most popular converter topologies, suitable for medium to high power applications as they offer many benefits such as operation at higher voltage and power levels, their lower dv/dt , output voltage and current distortions, and their capability of generating smaller common mode voltage. Thus, the use of multilevel converters has become more popular recently in various industrial high power applications such as laminators, mills, conveyors, fans and pumps [52 – 54].

Three different multilevel converter topologies in use are neutral point-clamped (NPC), flying capacitor (FC) and cascaded multilevel converters [52]. In addition, the combination of these topologies gives rise to other hybrid multilevel converter topologies.

Multilevel converters have been proposed for IPT systems in [39] with “Marx” multilevel inverter, in [45] with the conventional cascaded multilevel converter, and with an effort to reduce number of power switches, a modified hybrid multilevel converter (MH-MLC), which was first introduced in [47], has been proposed for BIPT systems in [42].

Along with the additional switching devices of the multilevel converter topologies are employed, the control complexity, voltage imbalance and EMI issues due to high operating voltage are major challenges. However, the advent of the commercial products in relation to high power multilevel converters demonstrated the feasibility of these topologies [54].

This chapter will present some feasible multilevel converter topologies together with modulation strategies that can be employed for high power IPT systems with

the aims of reducing the power losses and improving the output voltage and current characteristics.

Conventional cascaded multilevel converters (CC-MLC), neutral point clamped multilevel converters (NPC-MLC) and flying capacitor multilevel converters (FC-MLC) will be presented first. Thereafter, modified cascaded multilevel converters (MC-MLC) with reduced number of switching devices and modified H-bridge multilevel converters (MHB-MLC) with minimum number of power switches will be presented. Case study and comparison of different converter topologies will be provided to show the advantages and disadvantages of each topology.

Along with multilevel converter topologies, two modulation control methods will be presented in this chapter. Depending on the compensation circuit configurations, a suitable modulation control method is adopted to minimize the power losses through the power converters. Specifically, selective harmonic elimination (SHE) modulation method will be employed for BIPT systems which are compensated by LCL (or CLCL, LCCL) networks to improve the output current waveform quality [44]. Thereby, the power losses caused by the high order harmonic components and the switching losses through the power converter will be mitigated. If series compensation circuit is used in IPT systems, SHE modulation method is unnecessary as the output current waveform is nearly sinusoidal. In this situation, an advanced modulation strategy will be proposed to minimize switching losses through the power converters.

5.2 Cascaded multilevel converter

5.2.1 Topology description

The most significant benefit of the cascaded multilevel converter is its inherent ability to connect multiple low power modules to increase the overall power rating. This is useful in the case of multiple low power sources such as PV cells.

Figure 5.1 shows the conventional cascaded multilevel converter (CC-MLC) topology and the phase shift modulation input gate drive signals and output profiles. By using an H-bridge converter in the pickup side, the power flow can be made bidirectional. BIPT systems have numerous advantages compared to unidirectional

Figure 5.1: Cascaded multilevel converter based BIPT system.

IPT systems as discussed in chapter 4. The power direction and input VA rating can be controlled by the phase shift angle θ between primary and secondary converters. The series-series (SS) compensation circuit is employed for both resonant sides of the IPT system.

Several types of multilevel converter topologies as well as modulation strategies have been proposed for IPT systems [42, 44, 45]. The modulation strategies focused on eliminating the high order harmonic components of the output voltage. This modulation strategy is beneficial in case LCL resonant circuit is employed. More details about this modulation strategy will be presented in section 5.4. In this section, an advanced modulation strategy will be carried out for SS resonant circuit of IPT systems as shown in Figure 5.1. As a result in chapter 4, output current waveforms of SS resonant configuration which was powered by the H-bridge converter are already nearly sinusoidal. Therefore, the elimination of high order harmonic components of SS resonant converters is not very necessary.

An important consideration to apply the multilevel converters to IPT systems is the power loss that consists of converter switching loss and conduction loss. This section proposes an advanced phase shift modulation strategy to control the CC-MLC with minimized switching losses.

In relation to conduction losses, depending on the number of cascaded modules, the conduction loss of multilevel converter is larger than that of the conventional H-bridge converter if the switching devices employed are the same. However, in the situation where the input DC voltages of all multilevel converter modules are low, the OptiMOS power MOSFET with extremely low on-state resistance could be applied. Thus, the conduction loss of the multilevel converter can be kept even lower than that of the conventional converter. To illustrate this assumption, we compare 12 modules of CC-MLC with conventional H-bridge converter. The CC-MLC and the conventional H-bridge converter employ the OptiMOS MOSFET (IPB025N10N: 100V, 180A, 2.5 m Ω) and the high speed SiC MOSFET (C2M0080120D: 1200V, 32A, 80 m Ω), respectively. From this example, one can see that the conduction loss of CC-MLC is lower than that of the conventional converter for the same system voltage rating.

5.2.2 Derivation of the proposed phase shift modulation strategy

Firstly, let us consider the switching loss of the H-bridge converter with the phase shift modulation method.

Assume that the output voltage and current of the H-bridge converter are in phase and the output current is sinusoidal as shown in Figure 5.2. The instantaneous switching current of the H-bridge converter is $I_0\sqrt{2}\cos(\varphi/2)$. Thus the switching loss is zero when phase shift angle between converter legs is equal to π . Furthermore, to achieve zero switching loss when the phase shift angle is zero, (S_1 , S_3) and (S_2 , S_4) should be turned on and off for the whole duration required, respectively. This is the idea to derive the modulation strategy with minimum switching loss described in the following section.

Figure 5.2: H-bridge converter output waveforms.

5.2.3 The proposed phase shift modulation strategy

Let us consider the CC-MLC that consists of n -modules of H-bridge converters as shown in Figure 5.1 (a). The output voltage of each H-bridge converter and the output voltage of CC-MLC are shown in Figure 5.1 (b).

The output voltage of each module is given as follows,

$$v_i(t) = \frac{4V_{DC1,i}}{\pi} \sum_{k=1,3,5,\dots}^{+\infty} \left[\frac{1}{k} \cos(k\omega_t t) \sin\left(\frac{k\phi_{1,i}}{2}\right) \right] \quad (5.1)$$

And the output voltage of CC-MLC is given as follows,

$$v_p(t) = \sum_{i=1}^n v_i(t) \quad (5.2)$$

The amplitude of the fundamental component of output voltage is as follows,

$$V_{pm} = \frac{4}{\pi} \sum_{i=1}^n V_{DC1,i} \sin\left(\frac{\phi_{1,i}}{2}\right) \quad (5.3)$$

The phase shift angle is calculated so that the switching loss is minimized. Let us define $V_{pm,ref}$ as the desired amplitude voltage of the CC-MLC which is obtained from the closed loop controller. Assume that the input DC voltages are sorted as follows,

$$0 = V_{DC1,0} < V_{DC1,1} \leq V_{DC1,2} \leq \dots \leq V_{DC1,n} \quad (5.4)$$

The phase shift angle of each converter is defined using the following rules,

$$\begin{aligned} 1. \text{ If } \frac{4}{\pi} V_{DC1,i} < V_{pm,ref} \leq \frac{4}{\pi} V_{DC1,i+1} \\ \Rightarrow \begin{cases} \sin \frac{\phi_{1,i+1}}{2} = \frac{V_{pm,ref}}{\frac{4}{\pi} V_{DC1,i+1}} \\ \phi_{1,j} = 0 \text{ for } j \neq i+1 \\ i = 0, 1, \dots, (n-1) \\ j = 1, 2, \dots, n \end{cases} \end{aligned} \quad (5.5)$$

$$\begin{aligned}
 & 2. \text{ If } \frac{4}{\pi} \sum_{i=k}^n V_{DC1,i} < V_{pm,ref} \leq \frac{4}{\pi} \sum_{i=k-1}^n V_{DC1,i} \\
 & \Rightarrow \begin{cases} \sin \frac{\varphi_{1,k-1}}{2} = \frac{V_{pm,ref} - \frac{4}{\pi} \sum_{i=k}^n V_{DC1,i}}{\frac{4}{\pi} V_{DC1,k-1}} \\ \varphi_{1,i} = \pi \quad \text{for } k \leq i \leq n \\ \varphi_{1,i} = 0 \quad \text{for } i < (k-1) \\ k = 1, 2, \dots, n \end{cases}
 \end{aligned} \tag{5.6}$$

The first rule describes the case where normalized reference voltage $\frac{\pi}{4} V_{pm,ref}$ is less than the largest input DC voltage while the second rule describes the case where normalized voltage is greater than all the input DC voltages. Figure 5.3 illustrates these rules in greater clarity.

(a) For the rules in (5.5)

b) For the rules in (5.6)

Figure 5.3: Illustration of advanced phase shift modulation strategy.

With the proposed modulation scheme, at any point in time, only one module is controlled with phase shift angle: $0 < \phi_{1,i} < \pi$, while phase shift angles of other modules are either zero or π . Therefore, as mentioned in section 5.2.2, the switching loss of these modules is zero.

5.2.4 Simulation results

The BIPT system in Figure 5.1 is simulated with the parameters in Table 5.1. Three H-bridge modules are employed in the primary side converter. Figure 5.4 shows the simulation results of the proposed system. It is obvious from Figure 5.4 that the input and output currents are nearly sinusoidal and are in phase with the input and output voltages, respectively. In this simulation, there are two H-bridge modules operate at maximum phase shift angle and one module operate at $0 < \phi < \pi$. Therefore, at the moment of zero input current, the primary H-bridge modules which have maximum phase shift angle will achieve soft switching.

Figure 5.4: Instantaneous voltages and currents of the converters.

Table 5.1: Simulation parameters of CC-MLC

Parameter	Symbol	Value
DC input voltage	$V_{DC1,1}$	150 V
	$V_{DC1,2}$	140 V
	$V_{DC1,3}$	160 V
DC output voltage	V_{DC2}	500 V
Coils inductance	$L_i = L_1 = L_2 = L_o$	183 μ H
Equivalent AC resistances	$R_i = R_1 = R_2 = R_o$	0.1 Ω
Compensator capacitances	$C_1 = C_2$	0.06 μ F
Switching frequency	f_T	48 kHz
Coupling coefficient	k	0.38

5.3 NPC and FC multilevel converters

5.3.1 NPC multilevel converter

Figure 5.5: Single phase Neutral Point Clamped multilevel converter

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In this section, the applicability of the three-level and five-level NPC and FC converters for BIPT systems will be discussed. Among three basic multilevel converter topologies: cascaded H-bridge, NPC and FC, the NPC converter which was proposed 30 years ago is most widely used in industrial applications [54]. In contrast to cascaded multilevel converters, NPC and FC converters accept high input DC voltages up to 6 kV with several commercial products as shown in [54]. The DC voltage is split into multi levels by several series-connected bulk capacitors. In NPC converters, the output neutral point is clamped by the diodes. Meanwhile, the output neutral point in FC converters is clamped by the flying capacitors themselves.

Figure 5.6: Phase shift modulation for 3-level NPC converter.

Figure 5.5 (a) shows the NPC three-level converter. The output voltage v_{an} has three states: $V_{dc}/2$, 0, and $-V_{dc}/2$. For output voltage level $V_{dc}/2$, S_1 and S_2 need to be turned ON; for the “0” level, S_1' and S_2 need to be turned ON; and S_1' , S_2' need to be turned ON for the output voltage level $-V_{dc}/2$. The two switches S_1' and S_2' are known as the complementary switch pairs with S_1 and S_2 , i.e. there are two complementary pair (S_1 and S_1') and (S_2 and S_2'). In this topology, with phase shift

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modulation, the output voltage v_{an} has similar form to the output voltage of H-bridge converter. The modulation strategy is illustrated in Figure 5.6.

A five-level NPC converter is shown in Figure 5.5 (b). The output voltage v_{an} has five states which are a result of the combination of switches as follows,

- $v_{an} = V_{dc}/2$ if all upper switches $S_1 - S_4$ are turned ON.
- $v_{an} = V_{dc}/4$ if three upper switches $S_2 - S_4$ and one lower switch S_1' are turned ON.
- $v_{an} = 0$ if S_3, S_4, S_1' and S_2' are turned ON.
- $v_{an} = -V_{dc}/4$ if the upper switch S_4 and three lower switches $S_1' - S_3'$ are turned ON.
- $v_{an} = -V_{dc}/2$ if all lower switches $S_1' - S_4'$ are turned ON.

The four complementary switch pairs are (S_1, S_1') , (S_2, S_2') , (S_3, S_3') , and (S_4, S_4') . The modulation strategy for five-level NPC converter can be seen in Figure 5.7. It is configured such that S_3 and S_4 are shifted by $\pi/2$ from S_2 and S_1 respectively and then are inverted.

Figure 5.7: Phase shift modulation for 5-level NPC converter.

5.3.2 FC multilevel converter

Figure 5.8: Flying capacitor multilevel converter

Figure 5.9: Phase shift modulation for 3-level Flying Capacitor converter.

Figure 5.8 (a) shows a FC three-level converter. Similar to the NPC three-level converter, the output voltage of the FC three-level converter v_{an} has three states: $V_{dc}/2, 0, \text{ and } -V_{dc}/2$. For voltage level $V_{dc}/2$, two upper switches S_1 and S_2 need to be turned ON. For the “0” level, there are two combinations: either pair (S_1, S_1') or $(S_2,$

S_2') need to be turned ON. For the voltage level $-V_{dc}/2$, two lower switches S_1' and S_2' need to be turned ON.

A FC five-level converter is shown in Figure 5.8 (b). Unlike NPC five-level converter, the switching state combination of FC five-level converter has more redundant states for an output voltage state. There are three combinations for the phase-leg to the neutral point voltage level $v_{an} = \pm V_{dc}/4$ and six combinations for $v_{an} = 0$. The switching state combinations are summarized in Table 5.2.

Figure 5.10: Phase shift modulation for 5-level Flying Capacitor converter.

In the FC multilevel converter, the most important consideration is the balancing of flying capacitor voltage. The flying capacitor voltage is balanced when the charging and discharging time is equal with identical current. Figure 5.9 and Figure 5.10 show the phase shift modulation strategy as well as output voltages and currents of three-level and five-level FC converters, respectively. It is clear that the

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time for charging and discharging the capacitors is equal with an identical current amount. Do note that the dead time of commutation process does not affect the equalization of charging and discharging duration.

With the proposed phase shift modulation, the control signal of each switch will be a square waveform, resulting in heat distribution balancing of the power switches which is a drawback with respect to NPC converters.

Similar to cascaded multilevel converters, the advanced modulation strategy can be applied for the NPC and FC converters to minimize the switching losses when the SS compensation circuit is employed.

Both the NPC and FC converters are simulated by Matlab/Simulink with the BIPT circuit configuration as shown in Figure 5.11 and the simulation parameters as shown in Table 5.3.

Table 5.2: Switching combinations for flying capacitor 5-level converter.

v_{an}	$S_1,$ $\overline{S_4}'$	$S_2,$ $\overline{S_3}'$	$S_3,$ $\overline{S_2}'$	$S_4,$ $\overline{S_1}'$
$V_{dc}/2$	1	1	1	1
$V_{dc}/4$	1	1	1	0
	0	1	1	1
	1	0	1	1
0	1	1	0	0
	0	0	1	1
	1	0	1	0
	1	0	0	1
	0	1	0	1
	0	1	1	0
$-V_{dc}/4$	1	0	0	0
	0	0	0	1
	0	0	1	0
$-V_{dc}/2$	0	0	0	0

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In Table 5.2, the symbols $\overline{S_x}$ is the inverted state of S_x where $x = 1, 2, 3, 4$.

Figure 5.11: IPT system circuit configuration for simulation.

Table 5.3: Simulation parameters

Parameters		Value	Unit
Input and output DC voltage ($V_{DC1} = V_{DC2}$)	Three-level	1000	V
	Five-level	2000	
Coils inductance ($L_1 = L_2$)		33.8	μH
Equivalent AC resistance ($R_1 = R_2$)		0.05	Ω
Compensated capacitance ($C_1 = C_2$)		0.3	μF
Switching frequency		50	kHz
Mutual inductances		10	μH
Output power rating	Three-level	120	kW
	Five-level	280	

5.4 Modified H-bridge multilevel converter

5.4.1 Topology description

In the previous sections, three typical multilevel converter topologies for IPT systems have been presented. The presented topologies as well as modulation strategies have the advantages in terms of high power capability and low power losses. All aforementioned IPT systems are compensated by SS resonant networks with high quality output current.

However, in the situation where the LCL compensation circuit is employed, the output current quality is very poor so it significantly affects the power efficiency of the whole systems. LCL compensation network are known to have several advantages and has been employed in many IPT systems [88 – 96, 137 – 146]. Numerous attempts have been made to improve the output current quality of LCL compensation circuit configuration [152, 153]. This section will present a simple way to solve this issue.

Figure 5.12: BIPT system with 7-level modified H-bridge converter topology (upper) and output voltage waveforms (lower).

A modified H-bridge multilevel converter (MHB-MLC) topology which is first introduced in [50] with minimum number of power switches will be selected for BIPT systems as shown in Figure 5.12. The converter topology comprises six unidirectional switches (MOSFETs or IGBTs) together with the free-wheeling diodes and two DC power supplies. The primary converter output voltage, i.e., v_p , receives one of the following values: $0, \pm E_{p1}, \pm E_{p2}, \pm (E_{p1} + E_{p2})$. By setting $E_{p2} = 2E_{p1}$ (or $E_{p1} = 2E_{p2}$), the output voltage achieves seven different levels. In addition, at any instant of switching sequence, only three switches conduct the current. Therefore, the conduction losses will also be mitigated.

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Due to the limitation of voltage and current rating of switching devices, this converter topology is not suitable for high power IPT applications. In case of high power requirement, higher level converter topologies which are constructed from the basic 7-level converter topology as shown in Figure 5.13 can be employed.

Figure 5.13: General multilevel converter.

5.4.2 Modulation strategy

Table 5.4 shows the output voltages for different switching states of the power converter in Figure 5.12. States “1” and “0” indicate the ON and OFF states of the power switches, respectively. The 4th state gives two options to get zero voltage state in the output. The transition between two options in the 4th state can be completely soft switched by choosing appropriate switching instants. The transition between the following pairs of states (1st and 2nd), (3rd and 4th), (4th and 5th) and (6th and 7th) comprise one hard switching while the transition between two pairs of states (2nd and 3rd) and (5th and 6th) comprise two hard switching. In addition, at any instant of the output voltage cycle, there are three switches conducting the current.

Assume that the primary and secondary converter output voltage profiles are as shown in Figure 5.12. The DC power supplies for two converters are as follows,

$$E_{p2} = 2E_{p1} = 2E_p \quad \text{and} \quad E_{s2} = 2E_{s1} = 2E_s \quad (5.7)$$

The amplitudes of the k^{th} order harmonic component of the primary and secondary side output voltages are respectively as follows,

$$V_{pm}^{(k)} = \frac{4E_p}{k\pi} \left[\sin\left(\frac{k\varphi_{p1}}{2}\right) + \sin\left(\frac{k\varphi_{p2}}{2}\right) + \sin\left(\frac{k\varphi_{p3}}{2}\right) \right] \quad (5.8)$$

$$V_{sm}^{(k)} = \frac{4E_s}{\pi} \left[\sin\left(\frac{k\varphi_{s1}}{2}\right) + \sin\left(\frac{k\varphi_{s2}}{2}\right) + \sin\left(\frac{k\varphi_{s3}}{2}\right) \right] \quad (5.9)$$

Table 5.4: Switching states and output voltages.

No.	S ₁₁	S ₁₂	S ₁₃	S ₁₄	S ₁₅	S ₁₆	V _p
1	1	0	1	0	0	1	E _{p1} +E _{p2}
2	0	1	1	0	0	1	E _{p2}
3	1	0	0	1	0	1	E _{p1}
4	0	1	0	1	0	1	0
	1	0	1	0	1	0	
5	0	1	1	0	1	0	-E _{p1}
6	1	0	0	1	1	0	-E _{p2}
7	0	1	0	1	1	0	-(E _{p1} +E _{p2})

Two high order harmonic components which most significantly affect the distortion of the output currents are the third and the fifth orders. In order to eliminate undesirable harmonic components, the following equation has to be satisfied,

$$\begin{cases} \sin\left(\frac{\varphi_{p1}}{2}\right) + \sin\left(\frac{\varphi_{p2}}{2}\right) + \sin\left(\frac{\varphi_{p3}}{2}\right) = \frac{\pi V_{pm,ref}}{4E_p} \\ \sin\left(\frac{k\varphi_{p1}}{2}\right) + \sin\left(\frac{k\varphi_{p2}}{2}\right) + \sin\left(\frac{k\varphi_{p3}}{2}\right) = 0; k = 3, 5. \end{cases} \quad (5.10)$$

$$\text{and } \begin{cases} \sin\left(\frac{\varphi_{s1}}{2}\right) + \sin\left(\frac{\varphi_{s2}}{2}\right) + \sin\left(\frac{\varphi_{s3}}{2}\right) = \frac{\pi V_{sm,ref}}{4E_s} \\ \sin\left(\frac{k\varphi_{s1}}{2}\right) + \sin\left(\frac{k\varphi_{s2}}{2}\right) + \sin\left(\frac{k\varphi_{s3}}{2}\right) = 0; k = 3, 5. \end{cases} \quad (5.11)$$

where $V_{pm,ref}$ and $V_{sm,ref}$ are the desired voltages of the primary and secondary side converters.

Solutions for the given two equations will be modulation phase angles for the MHB-MLC.

5.4.3 Case study

To evaluate the advantages of the MHB-MLC topology, the comparison of different multilevel converter topologies which were applied for BIPT systems in the literature will be carried out.

Table 5.5 shows the comparison about the number of switching devices (N_{sw}) and conducting devices ($N_{conduct}$) of the CC-MLC presented in [45, 46], the multilevel converter presented in [44] and MHB-MLC in this section. It is noted that the number of switching devices and conducting devices of the NPC-MLC and FC-MLC presented above is the same as that of the CC-MLC. Assume that all the converter topologies comprise identical number of power supplies ($N_E = 2n$) and therefore they generate the same maximum number of output voltage steps (N_{step}). It is obvious from Table 5.5 that at any instant of the output voltage cycle, the number of total switching devices and the number of switching devices which conduct the current of the MHB-MLC are the least. It results in the reduction of switching and conduction losses of the power converter which is the most important aspect in IPT systems.

5.4.4 Simulation results

A 16 kW BIPT system as shown in Figure 5.12 is simulated to demonstrate the feasibility of the presented topology. The simulation parameters are shown in Table 5.6. Figure 5.14 and Figure 5.15 show the output voltage and current waveforms of the conventional BIPT system fed by H-bridge converter and the MHB-MLC BIPT topology with the SHE modulation method, respectively. Both cases are simulated using the identical BIPT system parameters given in Table 5.6. It is obvious that the output current of the MHB-MLC topology is almost sinusoidal since the third and the fifth harmonic components have been eliminated. This result leads to the reduction of total harmonic distortion of output voltage and current. Therefore, the power losses caused by the high order harmonic component are also reduced. In addition, the output voltage and current are in phase. This leads to unity input and output power factors with aforementioned features. All over the above statements,

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the effectiveness of the presented system is evidently improved in term of efficiency.

Figure 5.14: Output waveforms of the conventional systems

Figure 5.15: Output waveforms of the MHB-MLC BIPT system.

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Table 5.5: Comparison of different multilevel converter topologies for BIPT systems.

	CC-MLC presented in [45, 46]	Multilevel converter presented in [44]	MHB-MLC
N_{step}	$2^{2n+1}-1$	$2^{2n+1}-1$	$2^{2n+1}-1$
N_E	$2n$	$2n$	$2n$
N_{SW}	$8n$	$4n+4$	$4n+2$
$N_{conduct}$	$4n$	$2n+2$	$2n+1$

Table 5.6: Simulation parameters.

Parameter	Symbol	Value
DC power supplies	$E_{p1} = E_{s1}$	250 V
	$E_{p2} = E_{s2}$	500 V
Coil inductance	$L_i = L_1 = L_2 = L_o$	33.8 μ H
Coil inductance	$L_i = L_1 = L_2 = L_o$	33.8 μ H
Equivalent AC resistance	$R_i = R_1 = R_2 = R_o$	0.05 Ω
Compensator capacitance	$C_1 = C_2$	0.3 μ F
Switching frequency	f_T	50 kHz
Coupling coefficient	k	0.38
Power rating	16 kW	

5.5 Modified cascaded multilevel converter

5.5.1 Topology description

A multilevel converter topology with a reduced number of switching components has been first introduced in [48] to lower the power losses, cost and physical size. This section presents a cascaded topology based on the modified multilevel converter in [48] for high power – low to medium voltage IPT applications as shown in Figure 5.16. In the primary side of the BIPT system, a modified cascaded multilevel converter (MC-MLC) is employed. It comprises of multiple modified H-Bridge modules which are multilevel converters themselves. These H-Bridge converters are constructed using basic units which are switched – DC power supplies.

Due to the voltage rating of switching devices, maximum output voltage of each H-Bridge module will be limited. To solve this issue, a cascaded configuration is employed to increase the voltage and power capability.

A combination of advanced modulation and SHE modulation strategies can be employed in this situation to minimize the power losses of the system and improve the system characteristics. In Figure 5.16, SS compensation circuit is presented. However, both SS and LCL compensation circuit are suitable to be employed in this situation. In case of high power – high voltage applications, another cascaded multilevel converter can be employed in the secondary side as shown in Figure 5.16.

Figure 5.16: MC-MLC based SS resonant circuit BIPT system and its output voltage waveforms.

5.5.2 Modulation strategy

Similar to the modulation strategy presented in the CC-MLC section, in order to minimize the switching losses and conduction losses of power converters, the advanced modulation strategy for MC-MLC will be presented as follows.

Let us define the maximum fundamental components of output voltages of each module as shown in Figure 5.16 as $v_{1,max}$, $v_{2,max}$, ... and $v_{m,max}$; $v_{pm,ref}$ is the desired input voltage amplitude of the resonance circuit. Without loss of generality, assume that

$$v_{1,\max} \leq v_{2,\max} \leq \dots \leq v_{m,\max} \quad (5.12)$$

The modulation strategy of converter modules is defined using the following rules,

$$1. \text{ If } v_{i,\max} < v_{pm,\text{ref}} \leq v_{i+1,\max} \quad (i = 1 \div (m-1)) \quad (5.13)$$

Then the $(i+1)^{\text{th}}$ module will be selected to excite the resonance circuit.

The remaining modules will be idled.

$$2. \text{ If } \sum_{i=k}^m v_{i,\max} < v_{pm,\text{ref}} \leq \sum_{i=k-1}^m v_{i,\max} \quad (5.14)$$

Then the set of modules from $(k-1)$ to m will be selected to excite the resonance circuit. The remaining modules (from 1 to $(k-2)$) will be idled.

With the above set of rules, at any instant, the number of switching components that conduct the current will be minimized. In addition, the four main switches in the H-bridge of each module can be operated in fully soft switching mode. Therefore, the conduction and switching losses through power converters will be minimized in case the loads have a large range of demanded power.

Inside each module, the SHE method can be employed to determine the phase shift angle of staircase output voltage waveform. This is useful in the case where LCL compensation circuit is employed as mentioned in the previous sections. Depending on the required output power and the number of high order harmonic components need to be eliminated, the corresponding PV panels (or DC power supplies) will be switched ON. The other PV panels will be in OFF state.

5.5.3 Case study

Assume that each H-Bridge module comprises n basic units and the converter comprises m H-Bridge modules as shown in Figure 5.16. The number of output voltage steps (N_{step}), switching devices (MOSFETs or IGBTs) (N_{SW}) and DC power sources (N_E) are given as follows,

$$N_{\text{step}} = 2m(n+1)+1 \quad (5.15)$$

$$N_{\text{SW}} = m(2n+4) \quad (5.16)$$

$$N_E = m(n+1) \quad (5.17)$$

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With the same number of switching devices, the number of output voltage steps can be even higher than that in the MC-MLC topology if input DC voltage sources are asymmetrical [49].

Table 5.7 shows the comparison of the MC-MLC with the CC-MLC in [45, 46] and the MH-MLC presented in [42] for BIPT systems. In this comparison table, it is assumed that the CC-MLC and the MH-MLC in [42] comprise n modules and the MC-MLC comprises n basic units and only one module is employed ($m = 1$).

It is obvious from Table 5.7 that to achieve the same number of output voltage steps, the MC-MLC needs a fewer number of power switches compared to CC-MLC topology. In another words, the THD quality of output voltage is significant improved. Therefore, the power losses due to high order harmonic components will also be mitigated.

In addition, compared to the CC-MLC in [45, 46] and MH-MLC in [42], the MC-MLC has the advantage in term of lower conduction losses of power switches since the number of power switches of the MC-MLC conducting the current in a whole cycle is less than that of the others.

Table 5.7: Comparison of different multilevel converter topologies for BIPT systems.

	CC-MLC in [45, 46]	MH-MLC in [42]	MC-MLC with $m=1$.
N_{step}	$2n+1$	$2n+3$	$2n+3$
N_{SW}	$4n$	$2n+4$	$2n+4$
N_E	n	$n+1$	$n+1$
$N_{conduct}^*$	$2n$	$n+2$	$k+2$ **

* $N_{conduct}$ is the number of switching devices conducting the current when the converter operates at any instant.

** k is the number of active DC power supplies in the module.

5.5.4 Simulation results

Figure 5.17 shows the simulation results where input and output voltages and current waveforms are depicted. The simulation circuit configuration is identical to that in Figure 5.16 with identical converters in both sides. The number of modules

in each side are $m = 2$ and each module comprises $n = 1$ basic unit. Therefore, the output voltage has nine steps. The first three high order harmonic components (the third, the fifth and the seventh orders) are selected to be eliminated. The DC power supplies are all set to be identical as 250 V and the resonant circuit parameters are as shown in Table 5.1 in the previous section. The given BIPT system is designed with a rating power of 40 kW. It is obvious from Figure 5.17 that the output currents of the MC-MLC BIPT system are almost sinusoidal and in phase with the output voltages.

Figure 5.17: Simulation results: Voltage and current waveforms.

5.6 Conclusions

This chapter has shown the applicability of some multilevel topologies for a wide range of output power. Three basic multilevel converters and two modified multilevel converters have been presented together with the modulation strategies,

while case studies and simulation results show that the presented circuit topologies are feasible and applicable.

Cascaded topology is suitable in the situation where several individual DC sources are available such as PV cells. NPC and FC topologies are more suitable in the situation where compact high voltage source is available such as high power high voltage wind generators. A modified cascaded multilevel converter topology for IPT systems with reduced number of switching devices while maintaining the number of output voltage steps to be equal to that of the conventional cascaded multilevel converter is presented. All these four multilevel converter topologies are introduced for IPT systems with the aim to achieve higher voltage level for high power IPT applications such as fast charging systems.

In order to overcome the drawback of LCL compensation circuit topology in IPT systems, an advanced H-bridge 7-level converter which has the lowest number of power switches is considered with the selective harmonic elimination modulation strategy to improve the output voltage and current characteristics. In term of the number of switching devices per number of output voltage steps, the MHB-MLC has the least among all the presented converter topologies. However, since the MHB-MLC is presented with the aim of reducing the output current distortion, it is not suitable for high power-high voltage applications.

Overall, depend on the applications and the compensation circuits, appropriate converter topologies and modulation strategies will be employed. With the increasing adoption of IPT technologies in high power applications, these converter topologies promise to play an important role in high power and high efficiency IPT systems.

Conclusions and Future Works

6.1 Conclusions

A comprehensive study of efficient circuit topologies and modulation strategies for IPT systems was undertaken in this thesis. Attention was paid to the three important aspects as follows:

- To develop new power converter topologies for IPT systems with reduced power losses and electronic components.
- To investigate the control algorithms with optimized efficiency of IPT systems.
- To develop high frequency power converter topologies together with modulation strategies to supply high power IPT systems by taking into account the power efficiency.

In Chapter 1, a general introduction to IPT systems was presented. Motivations of a convenient and efficient power transmission generation have been shown to demonstrate the need of IPT systems. An overview of IPT technology achievements in both research aspects and commercialization aspects was presented.

In Chapter 2, a general overview of IPT technologies was presented. Fundamental components of IPT systems including magnetic coupling structures, power conversion types, switching techniques and compensation circuit configurations have been briefly described. Thereafter, several commonly used types of high frequency power converters including DC/AC voltage and current source converter, direct AC/AC matrix converter and multilevel converter topologies have been presented. The advantages and constraints of each converter topology have been carefully addressed.

A novel matrix converter topology was proposed for IPT systems in Chapter 3. A comprehensive study on the overall efficiency of the proposed system has been carried out by theoretical analyses, computer simulation and experimental verification results. It has been found that the power loss of the proposed topology is much less than that of the conventional two stage AC-DC-AC power converter in the case of high power requirements.

In Chapter 4, an optimized phase shift modulation strategy was proposed to maximize the overall efficiency of BIPT systems. It has been shown that a very important aspect of BIPT system, efficiency optimization, has been completely solved by theoretical analyses, simulation and experimental results. The proposed concept is a prerequisite for designing an efficient BIPT system.

The need of high power IPT systems for rapid growth in industrial applications is the reason for the high power BIPT systems which was presented in Chapter 5. Several multilevel converter topologies including conventional and advanced cascaded multilevel converters, NPC and FC multilevel converters, advanced H-bridge multilevel converter topologies together with advanced modulation strategies have been presented to increase the power scalability and minimize the power losses. Case studies and comparisons have been carried out to show the benefits of presented topologies.

6.2 Publications

Two journal and five conference papers have been published as an outcome of this thesis as follows:

1. Nguyen X. B., Vilathgamuwa D. M., Foo G., Peng Wang, Ong A., Madawala U. K., Nguyen T. D., “An Efficiency Optimization Scheme for Bidirectional Inductive Power Transfer Systems,” *IEEE Transaction on Power Electronics*, vol. 30, no. 11, pp. 6310 – 6319, 2015.
2. Nguyen X. B., Vilathgamuwa D. M., Madawala U. K., “A SiC-based matrix converter topology for inductive power transfer system,” *IEEE Transactions on Power Electronics*, vol. 29, no. 8, pp. 4029 – 4038, 2014.

3. Nguyen X. B., D. M. Vilathgamuwa, G. Foo, Wang Peng, Andrew Ong, “A Modified Cascaded Multilevel Converter Topology for high power BIPT systems with the reduction of power electronic components and losses,” Accepted for presentation in *IEEE Conference on Power Electronics and Drive Systems (PEDS 2015)*, June 2015, Australia.
4. Nguyen X. B., D. M. Vilathgamuwa, G. Foo, Wang Peng, Andrew Ong, “A Multilevel Converter Topology based Bidirectional Inductive Power Transfer System With Improved Characteristics,” Accepted for presentation in *IEEE Conference on Power Electronics – ECCE Asia (ICPE 2015)*, June 2015, Korea.
5. Nguyen X. B., D. M. Vilathgamuwa, G. Foo, A. Ong, P.K. Sampath, U. K. Madawala, “Cascaded multilevel converter based bidirectional inductive power transfer (BIPT) system,” in *IEEE International Power Electronics Conference (IPEC-Hiroshima - ECCE-ASIA)*, pp. 2722 – 2728, May 2014, Japan.
6. Nguyen X. B., D. M. Vilathgamuwa, G. Foo, A. Ong, U. K. Madawala, “Efficiency optimization for bidirectional IPT system,” in *IEEE Transportation Electrification Conference and Expo (ITEC)*, pp. 1 – 5, June 2014, USA.
7. Nguyen X. B., D. M. Vilathgamuwa, U. K. Madawala, “A matrix converter based Inductive Power Transfer system,” in *IEEE Conference on Power & Energy (IPEC)*, pp. 509 – 514, Dec. 2012, Hochiminh City, Vietnam.

6.3 Suggestions for future works

This thesis focused on the development of efficient circuit topologies for IPT systems including power converters and modulation strategies. Several power converter topologies and modulation strategies have been proposed to achieve the research objectives. Theoretical analyses, computer simulations and experimental verifications have been successfully undertaken. However, as mentioned in Chapter 2, there are many more areas of IPT systems that need to be comprehensively understood and investigated. Several suggestions for future works are listed as follows.

➤ Study on multiphase power converters for multiphase IPT systems

As discussed in Chapter 2 of this thesis, the multiphase IPT systems have been found to be feasible for the new generations of power transmissions as they offer numerous advantages. However, up to date, the development of multiphase IPT systems is only with full-bridge three-phase inverter [55, 56] which is limited by power rating and power controllability. A comprehensive study on the development of multiphase high frequency power converters is an important and interesting research topic in IPT area.

➤ **Comprehensive study on the characteristics of different compensation circuit configurations**

There are several compensation circuit configurations that have been presented in Chapter 2. Each of these circuit structures has different characteristics which offer advantages with respect to different applications. Thus, a comprehensive study on the characteristics of these compensation circuit configurations is also an important aspect that needs to be clarified.

➤ **Study on the communication between primary and secondary side power converters of BIPT systems**

BIPT systems offer many benefits in terms of bidirectional power controllability, especially for EVs charging applications. However, one of the biggest drawbacks of this topology is the communication between primary and secondary power converters. In order to control the power direction as well as the transmitted power amount, the phase shift angle (θ) between two converters needs to be synchronized. In case of EVs, however, the primary converter is underground, whereas the secondary converter is in the car. Synchronization algorithms for the primary transmitter and secondary pick-up are definitely interesting topic for future works.

➤ **Adaptive control algorithms with systems parameter identification**

IPT systems are considered as high order systems that are difficult in controlling, particularly when variable parameters are considered. Several control algorithms such as PID controller, PID fuzzy controller, PID tuning using genetic algorithm, sliding mode controller, etc., have been proposed for controlling this system. A comprehensive study on this aspect is a potential

future work. The application of system parameter identification and adaptive control techniques can be considered.

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