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Contactless Slipring Systems for Wireless Power Transfer in Rotary Applications

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A thesis submitted in partial fulfilment of the requirements for the degree of

Doctor of Philosophy in Electrical and Computer Engineering

The University of Auckland

October 2014
To my loving family
Abstract

Inductive Power Transfer (IPT) technology has motivated considerable research and development in the past two decades. This new technology can be used in various wireless power transfer applications with different specifications, necessities, and restrictions. One example application is to deliver electrical power from a static frame to a rotating shaft. This is generally achieved by using mechanical slipring assemblies. However, because of the physical contacts between the stationary brushes and rotating metal rings, frequent maintenance is required due to mechanical and electrical wear and tears. This would significantly increase the system operational cost in applications such as wind turbine pitch control. An alternative solution is to use magnetically coupled coils to achieve contactless power transfer to rotary loads based on inductive power transfer technology.

In this thesis three types of contactless slipring systems are proposed and developed based on pulsating, travelling, and rotating magnetic field principles. Full theoretical analysis, computer simulations, and practical experiments have been conducted on both the magnetic field coupling structures and power converter driving circuits to evaluate the proposed systems.

Existing single-phase contactless sliprings based pulsating magnetic field have been improved with new pot-core and through-hole magnetic coupling structures, and it is shown that they can increase the magnetic coupling coefficients by up to 45%. A poly-phase system based on axial travelling magnetic field principle is proposed, and shown to be able to transfer about 3.7 times more power than a single-phase system. Another poly-phase system is proposed based on rotating magnetic field principle, and shown to be able to increase the power density by 55% compared to the existing counterpart single-phase system.

An improved autonomous current-fed push-pull resonant converter is developed in this thesis to drive the single-phase contactless sliprings. It is shown that the proposed converter can increase the operating frequency to MHz level with full resonance and Zero Voltage Swathing (ZVS). Two new poly-phase current-fed push-pull resonant converters with ZVS operation are proposed and developed to drive the poly-phase contactless sliprings. The former uses an autonomous converter as the driving phase and allows the other phases to follow with a pre-determined phase delay; and the latter is based on full-autonomous operation of all the phases without any additional phase control. Experimental results have
demonstrated that these converters can generate good quality currents to excite the magnetic coils of the sliprings to achieve efficient contactless power delivery, which are practically useful for rotary applications such as wind turbine pitch control.
Acknowledgments

There are many people whom I would like to thank. Each of them has provided me with support and encouragement during this long and hard but joyful journey.

First and foremost, I would like to express my special thanks to my supervisor Dr Aiguo Patrick Hu for his insightful guidance during my doctoral studies. Patrick’s continuous supervision, as well as his valuable ideas and conceptual assistance have always been helpful and motivating.

Next, special thanks go to Dr Nirmal Nair for his encouraging guidance, suggestions, and very helpful proofreading.

Big thanks go to the University of Auckland for offering me the Doctoral Scholarship, and PowerbyProxi Ltd for the financial support and the upcoming job offered. Life would indeed have been a struggle without their financial support over the course of this thesis.

I would also like to thank my peers and friends at The University of Auckland, power electronics, and power systems research groups, all of them have been very supportive and helpful whenever I needed them. I would like to thank all the lab technicians for their technical assistance throughout my degree. I would like to thank the department administration office managers for their help at numerous occasions.

Last but not the least, sincere gratitude goes to my parents, Zemam & Hasna, my brothers and my sisters, for their full support and encouragement. Without their caring and help I would not have been able to come this far, thank you!

Ali Abdolkhani
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# Nomenclature

## Acronym

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<thead>
<tr>
<th>Acronym</th>
<th>Definition</th>
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<tbody>
<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>CLC</td>
<td>Capacitor-Inductor-Capacitor connection</td>
</tr>
<tr>
<td>CPT</td>
<td>Capacitive Power Transfer</td>
</tr>
<tr>
<td>CSS</td>
<td>Contactless Slipring System</td>
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<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DPF</td>
<td>Displacement Power Factor</td>
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<tr>
<td>EMC</td>
<td>Electromagnetic Compatibility</td>
</tr>
<tr>
<td>EMI</td>
<td>Electromagnetic Interference</td>
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<tr>
<td>ESR</td>
<td>Equivalent Series Resistance</td>
</tr>
<tr>
<td>FEM</td>
<td>Finite Element Model</td>
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<tr>
<td>FPGA</td>
<td>Field-programmable Gate Array</td>
</tr>
<tr>
<td>IGBT</td>
<td>Insulated Gate Bipolar Transistor</td>
</tr>
<tr>
<td>IPT</td>
<td>Inductive Power Transfer</td>
</tr>
<tr>
<td>LCL</td>
<td>Inductor–Capacitor-Inductor connection</td>
</tr>
<tr>
<td>MOSFET</td>
<td>Metal Oxide Silicon Field Effect Transistor</td>
</tr>
<tr>
<td>PI</td>
<td>Proportional and integral control</td>
</tr>
<tr>
<td>PLECS</td>
<td>Piecewise Linear Electrical Circuit Simulation</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase-Locked Loop</td>
</tr>
<tr>
<td>PV</td>
<td>Photovoltaic</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse Width Modulation</td>
</tr>
<tr>
<td>rms</td>
<td>root mean square</td>
</tr>
<tr>
<td>RF</td>
<td>Radio frequency</td>
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<tr>
<td>SMPS</td>
<td>Switch Mode Power Supply</td>
</tr>
<tr>
<td>TET</td>
<td>Transcutaneous Energy Transfer</td>
</tr>
<tr>
<td>THD</td>
<td>Total Harmonic Distortion</td>
</tr>
<tr>
<td>VCO</td>
<td>Voltage Controlled Oscillator</td>
</tr>
<tr>
<td>VHDL</td>
<td>VHSIC hardware description language</td>
</tr>
<tr>
<td>VLF</td>
<td>Very Low Frequency (3-30 kHz)</td>
</tr>
<tr>
<td>ZVD</td>
<td>Zero voltage detection</td>
</tr>
<tr>
<td>ZCS</td>
<td>Zero Current Switching</td>
</tr>
<tr>
<td>ZVS</td>
<td>Zero Voltage Switching</td>
</tr>
<tr>
<td>2D</td>
<td>Two Dimensional</td>
</tr>
<tr>
<td>3D</td>
<td>Three Dimensional</td>
</tr>
<tr>
<td>3φ</td>
<td>Three phase</td>
</tr>
<tr>
<td>C</td>
<td>Capacitor (Farads)</td>
</tr>
<tr>
<td>C_p</td>
<td>Primary resonant capacitance</td>
</tr>
<tr>
<td>C_s</td>
<td>Secondary resonant capacitance</td>
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## Symbol

<table>
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<td>3φ</td>
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<td>Primary resonant capacitance</td>
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<tr>
<td>C_s</td>
<td>Secondary resonant capacitance</td>
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</table>
F - Frequency (Hz)
$f_{sw}$ - Switching frequency
$f_{zvs}$ - Zero voltage switching frequency
H - Magnetic Field Intensity (A/m)
I - Current magnitude (Amperes)
$I_{in}$ - Primary input current
$I_L$ - Load current
k - Magnetic coupling coefficient
L - Self-inductance (Henrys)
$I_p$ - Primary resonant coil current
$L_d$ - Input DC inductance
$L_p$ - Primary resonant inductance
$L_s$ - Secondary resonant inductance
M - Mutual inductance (Henrys)
N - Number of turns
$N_1$ - Number of turns in primary coil/conductor or track
$N_2$ - Number of turns in secondary or pick-up coil
P - Power (W)
$P_L$ - Load power
$P_{loss}$ - System power loss
Q - Quality factor
R - Resistance (Ω)
$R_{cp}$ - Primary resonant capacitor ESR
$R_{cs}$ - Secondary resonant capacitor ESR
$R_{ds(ON)}$ - MOSFET ON state resistance
$R_{Lp}$ - Primary coil ESR
$R_{Ls}$ - Secondary coil ESR
$R_{load}$ - Load resistance
$R_{L1}$ - ESR DC inductor L1
$R_{L2}$ - ESR DC inductor L2
$S_u$ - Uncompensated pick-up output power
T - Mains period
$T_{sw}$ - Switching period
T - Time (s)
V - Voltage magnitude (Volts)
$V_{ds}$ - Drain to source voltage of MOSFET
$V_{dc}$ - Input DC voltage
$V_G$ - Gate source voltage
$V_{load}$ - Load voltage
$V_{ref}$ - Referenced voltage
$V_{rms}$ - Resonant tank root mean square voltage
$V_{oc}$ - Voltage induced at secondary
$V_{pk}$ - Resonant tank peak voltage
$V_z$ - Zener voltage
$\omega$ - Angular frequency (radian/s)
$\omega_r$ - Resonant frequency (radian/s)
$\omega_m$ - Mains frequency (radian/s)
$Z_p$ - Resonant tank input impedance
$Z_{sr}$ - Secondary impedance reflected to the primary
$Z_s$ - Secondary impedance
Chapter 1: Introduction

1.1 An outline to contactless power transfer

Contactless power transfer is to transfer electrical power from one point to another through an air gap without any direct electrical contacts. This technology has been used for applications as in EV’s (Electric Vehicles), consumer electronics, biomedical, etc. where conventional wires are inconvenient, hazardous, unwanted or impossible [1-7]. Supplying electrical power using electric plugs for battery charging, sliding contacts in rotary applications, supplies for trolley buses may cause electric shock, short circuit, sparking, etc. This makes the system unsafe and unreliable with reduced lifespan. Increasing the lifespan, reliability and low-maintenance operation can be achieved by eliminating the cables, mechanical sliprings as well as plugs and sockets.

In the early days of electromagnetism before the electrical-wire grid was deployed, serious interest and effort was devoted (most notably by Nikola Tesla) towards the development of schemes to transfer energy over long distances without any carrier medium. These efforts appear to have met with little success. Radiative modes of Omni-directional antennas (which work very well for information transfer) are not suitable for power transfer, because a vast majority of power is wasted into free space. Today, we face a different challenge than Tesla. Since the existing electrical-wire grids carry power almost everywhere, contactless power transfer technology creates new possibilities to supply portable devices with electrical energy which has been used in many different applications. A range of applications has been considered and several approaches have been proposed based on the individual requirements. Currently the main areas of contactless power transfer applications can be categorized as following [8]:

- Industrial (Operation in harsh environment Ex. mining, next to explosive gases)
- Automotive (Electric cars and general battery charging)
- Aerospace (Transferring energy to moving parts)
- Consumer Electronic (Charging a cell phone or a laptop wirelessly)
- Biomedical (Inductive interface to power implantable biomedical devices).
For rotary applications, mechanical sliprings are widely used to transfer electrical power to a rotatable part. However, they need frequent maintenance due to wear and tear caused by friction between stationary brushes and rotating metal rings. This would increase the system operational cost significantly in applications such as wind turbine pitch control [9]. Therefore, there is a need to develop an alternative contactless solution with low maintenance and long life cycle. A new contactless power transfer solution based on IPT (Inductive Power Transfer) technology has been proposed [2, 10-12], in which the primary and the secondary coils of a traditional transformer are wound on separate magnetic cores can overcome the mechanical slipring’s shortcomings. The advantages of this new IPT-based method of transferring power to rotating equipment without using electrical contacts have motivated many researchers in the past. They have found that the magnetic coupling structure and winding layouts, the air gap between the two parts, the inductors misalignment and thermal issues are some of the factors that affect the electrical behaviour of the system. Different approaches, either by improving the magnetic coupling structure [13-18]; or leakage inductance compensation (resonant techniques) [19-25]; have been researched on to overcome the problem of low magnetic coupling coefficient and high leakage inductances in this field, and increase the system power transfer capability. Despite all the progress made so far in the field, still more research is needed in contactless sliprings systems in order to improve the systems overall performance.

1.2 Introduction to contactless slipring system

Contactless slipring system is a new alternative to the widely used mechanical slipring assemblies. A typical contactless slipring is constructed by winding the primary and the secondary coils into separate halves of a ferrite core; these two halves can be arranged to face each other or to be coaxial with each half mounted on one of the primary and the secondary parts. As the electrical contacts eliminated in such system, magnetic link is used to provide a reliable and efficient transfer of electrical power across an air gap. The magnetic flux couples from one half of the core to the other half provide a mutual inductance that couples energy from the primary to the secondary side. In fact a typical contactless slipring functions like a power transformer, but makes use of a rotating secondary side to accommodate the equipment's rotation. In a contactless slipring system, because of the structure symmetry, there is a complete overlap between the primary and secondary coils for any angle of rotation.
This is in combination with a uniform fixed air gap, results in constant electrical properties and therefore, this system is not affected by the rotation.

Although mechanical slipring systems can be used for power transfer to rotary objects, they need regular maintenance due to friction, wear & tear, and jittery contacts. By comparison, a contactless slipring system has none of these limitations. The reliability and long life cycle, in combination with minimum maintenance requirement, are the main advantages of contactless slipring systems. In this thesis the term CSS is used for Contactless Slipring System which normally refers to a loosely coupled system designed for delivering power to a rotatable object. A CSS is an IPT-based system and can be considered as a loosely coupled power transfer system using modern power conversion, control, and magnetic coupling techniques to achieve contactless power transfer from a static frame to a rotatable electrical load across an air gap efficiently.

The fundamental theory of contactless slipring systems is governed by Faraday and Ampere’s laws as shown in Figure 1-1. Based on Ampere’s law, current $I$ generates a magnetic field $H$. Some of this magnetic field links the secondary power pickup coil and according to Faraday’s law causes a voltage $V$ to be induced.

![Figure 1-1: Fundamental theory of a contactless slipring system.](image)

Based on Faraday’s law, the induced voltage within the pickup coil can be produced by one of the following methods:

1. Stationary coil and time-varying magnetic field (transformer induction).
2. Relative movement between coil and static magnetic field (motional induction).
Chapter 1: Introduction

3. Relative movement between coil and time-varying magnetic field (both transformer and motional induction).

In this thesis, different types of contactless slipring systems are researched on and developed which their basic principle falls into one of the above induction types. In Chapter 3 single-phase systems are presented which their basic principle is the same as the transformer type of induction. Whereas, poly-phase systems are presented in Chapter 6 and Chapter 7 which their principles are the motional type of induction.

1.3 General features of contactless slipring system

As compared to the mechanical slipring assemblies, contactless slipring system has the following main features:

**Contactless:** Because of the air gap between the primary and secondary sides of a CSS, the secondary can move, rotate or keep still relative to the primary. This freedom of rotation makes it possible to deliver power to a rotatable electric load without direct electrical contacts.

**Free of noise:** A contactless slipring system transfers power that is free of noise as compared to the mechanical sliprings which associated with noise due to the contact between the stationary brushes and rotating part.

**Low maintenance:** Since there is no direct contact between the two sides, a CSS requires low maintenance compared to mechanical sliprings which require frequent maintenance to maintain their performance.

**Safe and reliable operation:** Mechanical sliprings are associated with wear and tear which can cause unwanted arcing and damage from dirt in field conditions. In comparison, CSS eliminates the arcing and it is inherently resistant to dirty environments.

1.4 Suitable applications of contactless slipring system

**Wind turbine pitch control system.** An important application of the CSS is in wind turbine pitch control applications. This control system is mounted on a rotary shaft of the turbine and it is used to control the blade angle according to the speed of the wind. A wind turbine pitch control system is shown in Figure 1-2. Currently mechanical sliprings are used to supply power to this system. The operational cost would be very high in case of need for cleaning or
brush replacement. Therefore, replacing the mechanical slipring system with a contactless method would be very advantageous for wind turbine applications [26, 27].

**Doubly-fed induction generator (DFIG).** The CSS technology can also be used for other wind turbine applications replacing the mechanical slipring assemblies on the rotor of the generator as shown in Figure 1-2. DFIG is an induction generator with a multi-phase wound rotor and a multi-phase slipring assembly with brushes to access the rotor windings. It is important to replace the multi-phase slipring assembly with the contactless method in order to avoid all the concerns associated with the mechanical sliding contacts as detailed in [9, 28]. However, it may be required to design the contactless slipring system for low frequency and using laminated iron magnetic cores.

![Figure 1-2: Wind turbine pitch control system of a DFIG.](image)

**Excitation systems of synchronous machines.** A CSS can be used in other applications such as synchronous machines excitation systems transferring power to the rotor winding. Using such a method for excitation systems has great advantages as compare to the currently used solutions. It eliminates the direct contact and frequent maintenance of the brush-system and it is capable of transferring power to the standstill shaft with a faster response compared to the brushless rotating-diodes system [29].

**Robotics applications.** Transferring power across a robot joint by the means of mechanical sliprings and flexible cables is undesirable [30]. Contactless sliprings offer a very good solution to eliminate direct mechanical contacts or cable twisting to achieve 360 degrees freedom of rotation.
Chapter 1: Introduction

**Radar applications.** One of the other important applications of the CSS is to transfer electrical power from the static frame to an airborne frame of radar. Same as other presented applications, usually power transfer is achieved by the use of mechanical slipring assemblies. If this can be achieved wirelessly using a contactless slipring system, it would have great benefits of removing physical contacts between the stationary and revolving parts and minimizing the maintenance [3, 31].

1.5 **Advances and challenges**

1.5.1 **Historical achievements**

Transferring power from a static to a rotary frame across an air gap is not new and it has been a challenge for many researchers over the years since the 1970s. Different alternatives have been proposed for the magnetic coupling structure and the power electronics around it. In 1970, a rotating transformer was proposed to transfer power from rotating photovoltaic panels to a satellite by Landsman [10]. One year later in 1971, a U.S. patent proposed a rotating transformer structure with a single air gap used to transfer signal and power to a rotating part [11]. In 1982, author of [32] proposed a rotating transformer with a closed air gap and not open directly to the free space. He claimed that the air gap of the transformer which is located within the magnetic core provides a self-shielding effect in order to minimize the stray flux leakage and EMI (Electromagnetic Interference). In the same period of time in 1982, a new air-bag system based on rotating transformer is proposed and tested for 20 kHz to have the benefits of completely noiseless operation [18]. Kyung-Ho Kim in [33] calculated the magnetic coupling coefficient and electrical parameters of contactless power supply by 3D finite element method by varying the secondary side position at 1 kHz operating frequency. He has found that one way of reducing the leakage flux is to design the system with segmented magnetic cores.

Authors of [2] with the co-operation of Moog Electric company presented two adjacent and coaxial designs for a rotating transformer using pot-core geometry. Based upon the design method, a 2 kW system driven by a full-bridge zero-voltage-switched power converter is constructed and tested for 100 kHz.

US patent number 7098551 (Aug, 2006) proposed an asynchronous machine to transfer power to the rotating parts of the wind turbine installation [34]. The rotor is connected to the rotating part, while the stator is connected to the non-rotating part of the wind power
installation. Due to the relative movement between the rotor and the rotating stator field, magnetic field is induced in the rotor windings to form a voltage source. In this condition the asynchronous machine is operating in a generating mode. The induced AC voltage in the rotor windings then can be used for desired purpose in the rotating part of the wind turbine installation.

US patent number 2007/0120634, designed a new structure model of rotating transformer with multiple secondary windings. Several windings are included in the secondary where each one of them can be connected to a separate load. An alternating current with 1 kHz in the primary produces an alternating electromagnetic field having flux lines in the secondary oriented in parallel to the axis of rotation. An aluminium cylinder as an electromagnetic shield is mounted on the outer surface of the shaft. The shield isolates the shaft from the alternating electromagnetic field to prevent eddy currents heating the shaft [35].

Papastergiou, K. D. proposed in [3], rotating transformers with two adjacent and coaxial arrangements for airborne radar power supply applications as shown in Figure 1-3. The rotating transformer prototypes are built on Ferro-cube pot cores with a diameter of 66 mm (P66/56). The transformers were designed to handle 1 kW of power at an input voltage of 330 V to operate with bipolar square current waveforms at 100 kHz. The transformers tested with various air gap lengths ranging from 0.25–2 mm. By comparing the two winding arrangements, he stated that the coaxial layout results in reduced leakage inductance for the same number of winding turns and magnetic-core size. The primary and secondary windings’ resistance of the adjacent windings arrangement (5.8 and 0.7 Ω, respectively, measured at 100 kHz) are higher than those of the coaxial (1.8 and 0.37 Ω) due to the increased proximity losses (more winding layers). As presented in [3, 17] the two pot-core type of rotating transformers have achieved a power efficiency of about 90% with 1 mm air gap, 1 kW of power at 100 kHz.
MORADEWICZ, AJ in [19], developed a pot-core type 3 kW laboratory model of Inductive Contactless Energy Transfer (ICET) as shown in Figure 1-4. He applied a series resonant circuit and assuring zero current switching in order to minimize total losses of the system. The developed ICET system is controlled by high efficiency and fast FPGA-based controller and protection system. The resonant frequency is adjusted by extreme regulator which follows instantaneous value of primary peak current. The transformer tested for 10mm air gap with an overall efficiency of 93% has been presented in [19].

![Rotatable transformer with adjustable air gap.](http://powerbyproxi.com/)

PowerbyProxi Ltd team, a young and pioneer company in development of contactless power transfer technology, installed their first contactless slipring system with coaxial magnetic coupling in 2007 for one of the wind turbine pitch control in Spain. Figure 1-5 shows one of their slipring products (Proxi-Ring240) which transfers 240 Watts output power over 15 mm air gap [26].

![Contactless slipring system (Proxi-Ring240).](http://powerbyproxi.com/)
1.5.2 Present challenges

1.5.2.1 Theoretical developments

The aim of a contactless slipring system is to provide power to a rotary frame across a gapped magnetic structure. Its theoretical development relies on both magnetic and power electronics together as an integrated system. In the case of magnetic structure, designing a magnetic coupling structure with a small air gap would result in high magnetic coupling coefficient and increased power transfer capability. Modelling and representing the magnetic circuit and associate its geometrical characteristics with its electrical behaviour is very important as: 1) to enable predicting the circuit performance, and 2) to provide the insight needed to achieve an optimized design. Furthermore, the magnetic structure of a contactless slipring system combines the magnetic properties of both an ideal transformer and an inductor. There are more room for theoretical improvements in magnetization, magnetizing inductance, leakage inductance and their connection with the structure geometry and AC losses that are critical in power electronic designs. In contactless slipring systems, in order to reduce the skin and proximity effects associated with the coils, multi-strands woven Litz wire is often used. Modelling and developing functional analysis of such a phenomenon associated with Litz wire is of high importance for the development of an efficient CSS system.

Power electronics on the other hand, covers a large area including electronics, control and communications. Analysis and modelling of switch-mode nonlinear circuits are the main concern. Like most other power electronic applications, the further development of IPT depends largely on some fundamental advances in switch-mode nonlinear theories. Moreover, the loose magnetic coupling between the primary the secondary coils of an IPT power supply is more difficult to analyse than a traditional closely coupled transformer. This further increases the circuit complexity so that proper compensation and control have to be taken into consideration in the design [4, 36-38].

1.5.2.2 Technical limitations

Because of the air gap, designing a contactless slipring system poses some unusual design constraints compared to the traditional compactly coupled design. The relatively large gap in the magnetic circuit results in a low primary magnetizing inductance and high leakage inductances. Eddy currents caused by fringing flux, can be formed in the magnetic material near the air gap and cause power losses and EMI. Operating at high frequencies presents unique design problems due to the increased core loss, leakage inductance, and winding
capacitance. This is because physical orientation and spacing of the windings determine the leakage inductance and winding capacitance which are distributed throughout the windings in the magnetic structure [39-42]. Experienced SMPS (Switch-Mode Power Supply) designers know that SMPS success or failure depends heavily on the proper design and implementation of the magnetic components. Inherent parasitic elements in high frequency IPT applications cause a variety of circuit problems including: high losses, high voltage spikes necessitating snubbers or clamps, poor cross regulation between multiple outputs, noise coupling to input or output, restricted duty cycle range, etc. [43-45].

Some major constrains associated with the design and practical implementations are:

a. **Meeting the power demand.** It is often difficult to deliver the required power to a load via contactless sliprings due to limited space on the shaft and specific power flow regulations.

b. **Switching speed.** Operating at a higher frequency can help reduce the size of a contactless slipring. However the switching speed of the switches is one of the major constraints. The most suitable switching devices for IPT applications seem to be IGBTs (Insulated Gate Bipolar Transistors) with commercial products up to the power level of 3kV/2kA, and a switching frequency up to 80 kHz. Power MOSFETs (Metal Oxide Silicon Field Effect Transistors) can switch at a speed up to MHz levels, but their voltage levels are too low for high power IPT applications [46].

c. **Efficient operation.** Due to various copper and ferrite losses, achieving high system power efficiency is a challenging task.

d. **Maximum temperature rise.** This is an important factor to keep the system operating in an acceptable range of temperature, especially when the system is used in a specified temperature environment.

e. **System size/weight.** The size and weight are limiting factors in designing a contactless slipring system. The conversion process in power electronics requires the use of magnetic components that are usually the heaviest and bulkiest items of the circuit. The design of such components has an important influence on the overall system size/weight, power conversion efficiency, and cost.
f. **System stability and control.** It is always important to have a stable system in practical applications. Variable frequency controlled IPT power supplies can become unstable if not designed properly [47, 48].

g. **Cost effectiveness.** Existing contactless sliprings are more costly than traditional mechanical sliprings due to the complicated power electronic circuitry and magnetic coupling design. It is a challenging task to improve the system design to bring the cost down for practical applications.

h. **Compliance.** Final practical slipring products need to pass EMC (Electromagnetic Compatibility) and safety standards, which can be a challenging engineering design task.

The above mentioned challenges may interact with each other, making the system optimization very difficult. Trade-offs often needs to be made depending on practical constraints and requirements.

### 1.6 Research objectives and scope

The objectives of this thesis are:

1) to improve the existing single-phase contactless slipring systems by proposing new magnetic coupling structures and power converters.

2) to propose and develop new poly-phase magnetic coupling structures and power electronics driving circuits based on moving magnetic field principles.

This research involves the following key aspects:

1. **Magnetic field coupling.** This phase focuses on the study of the magnetic, electrical and mechanical properties of the magnetic coupling structure. Initially existing systems in the literature and industry are thoroughly studied and analysed to determine what improvements can be made. Then, proposing improved/new magnetic coupling structures for contactless slipring systems based on pulsating, travelling and rotating types of magnetic field theories. Three Dimensional Finite Element Method (3D-FEM) simulations of the proposed magnetic coupling structures will be completed at this stage using JMAG Studio.10 package.

2. **Power converters.** The second phase of the research aims at the strategy for most appropriate power converters to be integrated with the developed magnetic coupling
structures. Therefore, various types of converter topologies will be studied and analysed in order to develop either new or hybrid model of converter topologies for driving the proposed magnetic structures in the previous phase. This will cover single-phase as well as poly-phase converters topologies. Simulation models of the proposed power converters using MATLAB, PLECS and LTspice-IV packages will be conducted at this stage.

3. **System implementation and evaluation.** The full contactless slipring systems including both magnetic couplings and power converters will be built for practical testing and evaluation. The results will be fully analysed to assess the system performance and give suggestions for future work.

### 1.7 Outline of the thesis

The following chapters of the thesis are arranged as follows:

Chapter 2 is a general overview on the technologies involved in power transfer for rotary applications. It covers the basics of possible methods that can be used for power transfer in rotary applications. The emphasis is given to IPT-based contactless solution which is the subject of this thesis. The fundamental parts of the contactless IPT-based method is covered with more details including the magnetic structure characteristics, associated losses, primary inverting technologies, and review of power flow regulation methods.

In Chapter 3 the single-phase pulsating magnetic field-based systems are studied and basic magnetic coupling theory is introduced. Theoretical calculations, measurements, and finite element analysis are conducted in order to obtain the electrical properties of the magnetic structure. In addition, improvements made to model the physical structure by adding the ferrite losses to the equivalent model which results in a higher accuracy. Several methods of measuring the magnetic properties of a practical model are presented. In addition, a new method of determining the magnetic coupling coefficient at operating conditions is proposed and derived. Various improved magnetic coupling structures are proposed and compared with the existing structures.

Chapter 4 proposes an improved current-fed single-phase primary power converter for driving the single-phase contactless sliprings. The proposed converter is able to operate up to MHz levels with ZVS operation autonomously without any control circuitry required. Its basic principle including autonomous gate operation, start-up and steady state are analysed. A
Chapter 1: Introduction

complete simulation study of the proposed converter is undertaken and experimentally verified on the practical contactless slipring models.

Chapter 5 presents system-based study of the pulsating magnetic field-based systems presented in Chapter 3. It covers a full 3D-FEM simulation and experimental analysis of their power transfer capability and a comparison between different systems is conducted. Two methods of shielding the shaft from the stray flux lines using aluminium and flexible ferrite materials are presented.

Chapter 6 proposes a new poly-phase contactless slipring system based on axial travelling magnetic field. It covers the basic principle of generating travelling magnetic field and its capability of power transfer. A new index ‘per pole mutual inductance’ is introduced which simplify the analysis of a complex mutually coupled poly-phase system to a single-phase basis. The 3D-FEM simulation and experimental studies on the magnetic structure are undertaken to investigate the system’s magnetic characteristics properties and compared with an existing single-phase counterpart.

Chapter 7 proposes a new poly-phase contactless slipring system based on rotating magnetic field principle. It covers the basic principle of the rotating magnetic field, modelling and analysis of a poly-phase system based on reflected impedance from the secondary onto the primary side. Two primary design layouts and three secondary power pickups are proposed for this system. The 3D-FEM simulation and experiments are undertaken to investigate the system’s overall performance in relation to an equivalent existing single-phase system.

Finally, Chapter 8 summarises the conclusions of this thesis and suggests future work. The contributions of the research are highlighted, and the major advantages and disadvantages of different IPT-based contactless sliprings for rotary applications are presented.
Chapter 2: An overview of power transfer technologies in rotary applications

2.1 Mechanical slipring assemblies

In rotary applications, mechanical sliprings are the widely used solution to transfer power to an electrical load mounted on a rotary shaft. Mechanical sliprings are electromechanical devices that consist of rotary (rotor) and stationary (stator) parts as shown in Figure 2-1. They allow power to be transferred from their rotors to stators and vice versa. This is accomplished by either 1) holding the centre core stationary while the brushes and housing rotate around it, or 2) holding the brushes and housing stationary while the centre core is allowed to rotate. Mechanical sliprings are available in a wide variety of configurations, types and materials to fit most applications. Modern slipring technology has evolved to the point where very expensive materials or processes are used to extend the life cycle of the sliprings, such as using gold contacts and multi fibre brushes [49]. Mercury-wetted sliprings, noted for their low resistance and stable connection, using different principles which replaces the sliding brush contacts with a pool of liquid metal molecularly bonded to the contacts. During rotation the liquid metal maintains the electrical connection between the stationary and the rotating contacts.

However, the use of mechanical slipring assemblies is often undesirable; their inherent friction due to the contact between the stationary brushes and moving part cause wear and tear over time. Eventually, this is subjected to frequent maintenance or even full brush replacement. The replacement of the brush assemblies can significantly increase the operational costs in applications such as wind turbine pitch control system [9]. Moreover, the use of mechanical slipring assemblies introduces the danger of short circuits from the accumulation of electrically conductive particles between the rings. The sparking of mechanical sliprings assemblies make them unsuitable for explosive environments. In the case of liquid slipring systems, use of mercury poses safety concerns, as it is a toxic substance. If a slipring application involves food manufacturing/processing or any other use where contamination could be a serious threat, leakage of the mercury and the resultant contamination could be extremely serious.
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Figure 2-1: Mechanical slipring assemblies.

2.2 Capacitive power transfer

Figure 2-2 shows the basic configuration of a typical Capacitive Power Transfer (CPT) system. A power source is converted to a high frequency AC voltage and supplied to two primary metal plates. When two secondary metal plates are placed in proximity, electric fields are formed between them and displacement current is generated. As a result, power can be transferred without direct electrical contacts to supply the load and it also allows for some freedom of movement between the primary and the secondary plates [50].

As compared to IPT principle, CPT has several advantages. Firstly, as magnetic field can not penetrate through metals, IPT can not be used in situations where metal barriers exist between power sources and loads. Even when power transfer is not blocked, power losses of an IPT system can be very high if metal pieces are placed around the magnetic field.
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Secondly, use of electromagnetic field as the energy carrying medium causes EMI problems. Furthermore, standing power losses can be high because IPT systems are current driven and high magnitude currents are often required even under no load or light load conditions.

However, until now capacitive coupling is used in low power range (e.g. sensor supply systems) whereas inductive coupling allows transferring power from a few mW up to several kW. The authors of [51] proposed the use of aerodynamic fluid bearings to maximize the capacitive coupling and achieve high output power. However, in addition to proper sealing requirement, this method again relying on the relative velocity between the bearing surfaces to entrain the fluid into the bearing clearance and generate a pressure profile which defeats the main feature of a contactless power transfer system.

2.3 Contactless slipring based on semi-solid material

A type of contactless slipring system using semi-solid material as an intermediate is proposed in [52] for wind turbine applications as shown in Figure 2-3. The system includes a stationary conductive part, a rotating conductive part, and a conductive semi-solid material. The semi-solid material is used to transfer electric current from the rotating side to the non-rotating side [52]. A sealing mechanism is required for such a system in order to avoid the semi-solid material leakages. The conductive semi-solid material is suggested to be silver-impregnated grease, carbon-impregnated grease, metallic-alloy impregnated grease, conductive oil or conductive powder.

![Figure 2-3: Contactless sliprings with semi-solid material.](image-url)
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It can be observed that because of the presence of the semi-solid material and requirement of a proper sealing mechanism, this way of power transfer is not totally a contactless solution for power transfer. This means that there still is an electrical contact between the two parts which discards the main feature of contactless power transfer in the basis of with no direct electrical contacts.

### 2.4 Brushless exciter-based method

Generally synchronous generators require direct current to energize their magnetic field mounted on the rotor from a separate source called an ‘exciter’. Two brushed and brushless types of exciters are used for this purpose [53]. In the brushless type, the brushes and sliprings are replaced by a set of rotating diodes assembly as shown in Figure 2-4. The stationary part is fed either by a permanent magnet or a DC current to produce a stationary magnetic field in the air gap between the two sides. When, the armature coil of the exciter rotates, an emf is generated which can be rectified and fed to the load mounted on the rotating shaft and can be expressed by:

\[ e = Bvl \]  

(2.1)

where \( B \) is the magnitude of the flux density, \( l \) is the total length of the rotating coil and \( v \) is the relative velocity between the coil and the magnetic field.

Based on this principle, the power transferred to the load is zero when the shaft is standstill \((v = 0)\). Since this system is subjected to high speed rotation to generate the required voltage in the rotary coil; in the case of wind turbine applications with a low speed (about 20~30 rpm), this system is incapable of transferring the required power to the load.

![Figure 2-4: Block diagram of a brushless exciter system.](image-url)
2.5 IPT-based contactless slipring system

To overcome the shortcomings associated with mechanical sliprings as well as the possible contactless solutions; IPT-based contactless method is the suitable way that requires no electrical contact between the rotating and stationary parts for kW level. In this system, inductive components are used to offer a reliable transfer of electrical power to the rotating part across a sufficiently large air gap. This new contactless technology has been proposed in which the primary and the secondary windings of a traditional transformer are wound on separate magnetic cores [2, 12, 13, 25]. Magnetic flux provides the coupling from one half of the core to the other, providing the mutual inductance that couples energy from the primary to the secondary side. Figure 2-5 shows the block diagram of a typical inductively coupled slipring system. It consists of a primary side AC/DC/AC resonant converter which converts the rectified AC power into high frequency AC power. This converted AC power fed to a primary coil which physically separated from the secondary coil, while magnetically coupled together. Therefore, the secondary side can be movable (linearly or/and rotating) giving flexibility, mobility and safeness for supplied loads. In the secondary side, the induced high frequency AC power is converted and controlled by a secondary converter to meet the requirements specified by the load parameters. The following sub-sections present the function of each part of the IPT-based contactless slipring system shown in Figure 2-5.

![Figure 2-5: General structure of IPT-based contactless slipring system.](image)

### 2.5.1 Primary power converter

Because of the existence of air gaps between the two sides in a contactless slipring system, the required power transferred to the load is not sufficient at low frequencies (50/60 Hz). Therefore, use of high frequency power conversion is a necessity in order to make a stronger magnetic induction between the two coils. To achieve this, a primary converter is employed in the primary side to generate a high frequency alternating current flows through the primary
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coil generating an alternating electromagnetic field. In fact, a high frequency power converter improves the power transfer characteristics by fast changing rate of the magnetic field. In terms of physical size, by operating at a higher frequency, a given core and coil configuration is able to transfer more power without reaching saturation; thus, the magnetic coupling structure can be physically built more compact.

2.5.2 Contactless magnetic structure

To allow for the freedom of relative rotation between the primary and the secondary sides, a contactless magnetic structure is essential for contactless slipring systems. Generally, for high power and a small air gap, magnetic structures with magnetic cores in primary and secondary sides are appropriate. In contrast, for a large air gap and low power, coreless structures are preferred. A special case is a sliding transformer which can have a construction for linear or circular movement. The final configuration of contactless slipring systems depends also strongly on the number of loads to be supplied. In such cases, a transformer with multi secondary coils is used.

2.5.3 Secondary power converter

The time varying magnetic field generated by the primary coil induces an electromotive force in the secondary coil which forms a voltage source of the secondary power supply. Since the magnetic coupling of an IPT system is normally very loose compared to normal transformers, the induced voltage source is usually unsuitable to be used to drive the load directly. Thus, a power conditioner with proper circuit tuning and conversion is required to control the output power according to the load requirements [54].

2.5.4 System tuning

Tuning the primary and the secondary inductances of a contactless slipring system is important for several advantages: 1) Compensating the primary inductance makes the power supply able to drive the given limited voltage ratings of the employed switches, 2) Constructing a resonant tank for resonant converters which is required to enable soft-switched operation of the switches to reduce switching losses and EMI, 3) Prevent the harmonic propagation in the circuit due to its filtering function and 4) Compensating the secondary leakage inductance improves the power transfer capability.
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2.5.4.1 Primary tuning

Compensating the primary inductance improves the power factor and reduces the required Volt-Ampere rating of the primary power source. The compensation topologies are considered, namely series compensation, parallel compensation and series-parallel compensation as shown in Figure 2-6. In series compensation, the voltage across the capacitance compensates the voltage drop of the primary equivalent reactance, making the required supply voltage to reduce. In parallel compensation, the current going through the capacitance compensates the current in the primary winding. Hence the required supply current is reduced. Series-parallel compensation realizes the reduction of both in voltage and in current [55].

In Figure 2-6 in a series compensation topology, the output current from the inverter bridge (I_1) is equal to the primary current which is passing through the primary inductance (I_p). This signifies that the primary current in series compensation circulates through the inverter bridge, in turn, causes significant power dissipation in the switching network. Moreover, in series compensation, the voltage is boosted because of the added voltage across the tuning capacitor. This results in an increased voltage across the primary coil (V_p) that is higher than the inverter output voltage (V_1). This is beneficial as it allows the power supply to drive a high primary inductance with a desired primary current. The parallel compensation on the other hand, has a current increase property. The primary current (I_p) is greater than the power converter output current (I_1). This is because the reactive current is circulating inside the resonant tank, and only the real current is flowing through the inverter bridge. As a result, in a parallel tuned power supply design, lower current rated switches may be used. The value of the primary compensation capacitor is often designed to fully compensate the primary inductance at the primary operating angular frequency (\( \omega \)). In this case, the primary compensation capacitance is determined by:

\[
C_p = \frac{1}{\omega^2 L_p} \quad (2.2)
\]

Note that the capacitance of (2.2) is calculated for either a series or a parallel tuning topologies. In the case of a composite topology, it is a combination of the two topologies. One simple way is to use \( C_{p2} \) to fully cancel the primary leakage inductance (\( L_p-M \)), and \( C_{p1} \) to fully tune the remaining portion of the primary self-inductance (\( M \)):
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\[
C_{p1} = \frac{1}{\omega^2 M} 
\]

(2.3)

\[
C_{p2} = \frac{1}{\omega^2 (L_p - M)} 
\]

(2.4)

![Diagram of primary tuning topologies](image)

**Figure 2-6: Primary tuning topologies.**

### 2.5.4.2 Secondary tuning

In a loosely coupled contactless slipring system, because of the large internal reactance, the power transferred from the primary to the secondary side is not sufficient for the load. For an uncompensated secondary inductance, the maximum output power occurs when the load resistance \(R_L\) is equal to the internal reactance \(\omega L_s\), and can be expressed by \([56]\):

\[
P_{o,max} = \frac{1}{2} V_{oc} I_{sc} 
\]

(2.5)

where, \(V_{oc}\) is the open-circuit voltage at the secondary terminals and \(I_{sc}\) is the short-circuit current of the secondary side.

In order to improve the power transfer capability of the system, normally the secondary side is tuned to the primary frequency. There are two basic tuning topologies used in the secondary side, namely series compensation and parallel compensation as shown in Figure 2-7. An equivalent resistance (\(R_L\)) represents the load on the secondary side. In Figure 2-7(a); the equivalent reactance and its compensation capacitance cancel each other at the resonant frequency, making the output voltage independent of the load and equal to the secondary open-circuit voltage. Likewise, in parallel compensation the equivalent admittance and its compensating capacitance cancel each other at the resonant frequency, which makes the output current independent of the load and equal to the secondary short circuit current as illustrated in Figure 2-7(b). For maximum power transfer, the secondary resonant frequency
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is normally designed to equal the nominal frequency \( (\omega_0) \). In this case, the secondary compensation capacitance is determined as:

\[
C_s = \frac{1}{\omega_0^2 L_s}
\]  

\( \text{(2.6)} \)

\[ \text{Series (Z=0)} \]

\[ \text{Parallel (Z=\infty)} \]

*Figure 2-7: Secondary side tuning topologies.*

### 2.5.5 Equivalent circuit and power transfer capability

#### 2.5.5.1 T-equivalent model

The fundamental flux linkages in a contactless slipring system is through a secondary coil of inductance \( L_2 \) mutually coupled to a primary coil of inductance \( L_1 \) and carrying current \( I_1 \) through a mutual inductance \( M \), as shown in Figure 2-8. The mutual inductance \( M \) is a function of geometry and can be found by simulation, measurement or modelling the physical structure [17, 57]. Typical IPT systems are modelled using the conventional transformer model with an equivalent T-circuit as shown in Figure 2-9 [58]. It should be noted that normally the air gap in the path of the flux leads to low flux densities which tend to make the core loss very small, so the core loss component may be neglected from the model without significant loss of accuracy.
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Figure 2-8: A CSS with two mutually coupled coils.

Figure 2-9: T-equivalent circuit.

From the above model, the power transfer capability of a contactless slipring is determined by two fundamental parameters $V_{oc}$ (when $R_L=\infty$), and $I_{sc}$ (when $R_L=0$) obtained from two basic open-circuit and short-circuit tests as [59]:

$$V_{oc} = j\omega M I_0 = j\omega M I_1$$  \hspace{1cm} (2.7)

$$I_{sc} = I_1 \cdot \frac{j\omega M}{R_2 + j\omega[M + (L_2 - M)]}$$  \hspace{1cm} (2.8)

In contactless slipring systems, in order to reduce the winding losses, a woven Litz wire is normally used with low ESR; therefore, in (2.8) $R_2$ can be neglected giving the short-circuit current as [59]:

$$I_{sc} = \frac{M}{L_2} I_1$$  \hspace{1cm} (2.9)

From the above equations, the uncompensated power $S_u$ of the secondary side can be expressed as follows [59]:

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\[ S_u = V_{oc} I_{sc} = \omega \frac{M^2}{L_2} I_1^2 \]  

(2.10)

In practice, the maximum power transferred to the load side without compensation is \( S_u/2 \) given by [60]:

\[ P_{2,\text{max}} = \frac{1}{2} V_{oc} I_{sc} = \frac{S_u}{2} \]  

(2.11)

The power presented by (2.10) can be boosted up \( Q_S \) times, using resonant techniques. As detailed in the previous section, a capacitance can be added to the circuit in series or parallel to tune the secondary power pick-up circuit to the primary resonant frequency \( (\omega_0 = 2\pi f_0) \). The maximum power transferred to the load side for a certain \( Q_S \) then can be given as below [60]:

\[ P_2 = Q_S S_u = Q_S \omega_0 \frac{M^2}{L_2} I_1^2 \]  

(2.12)

where, \( Q_S \) is the secondary circuit quality factor and it is different depending on the secondary tuning topology as given by:

\[
Q_s = \begin{cases} 
\text{Seriestuning}: & \frac{\omega_0 L_2}{R_L} \\
\text{Parallel tuning}: & \frac{R_L}{\omega_0 L_2} 
\end{cases}
\]  

(2.13)

It should be noted that for the parallel tuned secondary, the secondary coil appears as a current source, while for a series tuned secondary it acts as a voltage source. The tuned output of the secondary coil is then rectified and regulated to a constant output voltage using an appropriate switch-mode controller according to the load requirements. However, based on the tuning topology, different power regulation circuitry such as boost, buck, etc. is used on the secondary side [37]. The quality factor \( Q_S \) is determined by the secondary circuit parameters as well as the equivalent output load which is related to the operation of the switch-mode controller [61, 62]. It is an important parameter which needs to be chosen accurately during design. Although higher \( Q_S \) increases the power transfer capability, however for several reasons in practice, the highest \( Q_S \) reachable is limited to about 10 [60]: Firstly, the secondary operational bandwidth is inversely proportional to \( Q_S \). Secondly, higher \( Q_S \) would be obtained at the expenses of the higher secondary VA rating so that practical size
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and thermal ratings also naturally limit $Q_S$ in commercial systems. Thirdly, practical component tolerances and aging affect the sensitivity of the circuit and make power transfer at high $Q_S$ impractical.

2.5.5.2 Mutual inductance model

The mutual inductance model uses the concepts of induced and reflected voltages to describe the coupling effect between the secondary and the primary networks. Both the induced and reflected voltages are expressed in terms of the mutual inductance. It has been shown that even though the power transfer in IPT systems happens in similar fundamentals to that of a conventional transformer; there are a few reasons that make it difficult to model the system using the conventional equivalent circuit [63]. Instead the magnetic interactions between the primary and the secondary sides are described using a mutual inductance model as shown in Figure 2-10 [5, 64-66]. Note that winding losses are assumed to be very small and not considered in the presented model.

The reasons for using the mutual inductance model are:

1) The magnetic coupling between the primary and the secondary coils is considerably lower than a conventional tightly coupled system.

2) Since there is no closed magnetic core, a major portion of the magneto-motive force will be situated across a substantial air gap. In such cases it is advantageous to fully utilise the primary by operating at the rated current of the coil, and letting the voltage vary as a function of the load. The primary power source thus needs to supply a controlled current to the primary coil [60].

3) In a mutual model, the primary and the secondary sides can be designed largely independently.

4) In a mutual inductance model, the leakage inductance is not modelled separately from the mutual portion for circuit analysis as in other models such as T-equivalent circuit. This is a major advantage for IPT-based contactless power transfer systems where the leakage inductance is too large to be ignored.
In the mutual inductance model, the reflected voltage is used to represent the total effect of the secondary on the primary side. However, the effect of the secondary must be considered together with the inherent properties of the primary coil in the analysis of the primary network. The reflected voltages generally determined by the current flowing within the secondary coil and the mutual inductance between the primary and the secondary as given by [66]:

$$V_r = -j\omega M I_2$$  \hspace{1cm} (2.14)

The effect of the secondary can be represented by the equivalent reflected impedance by dividing the reflected voltage by the primary current as follows [66]:

$$Z_r = \frac{V_r}{I_1} = \frac{-j\omega M I_2}{I_1}$$  \hspace{1cm} (2.15)

The secondary current flowing through the secondary coil ($I_2$) on the other hand is defined by the input impedance seen by the open circuit voltage as:

$$I_2 = \frac{V_{oc}}{Z_2} = \frac{j\omega M I_1}{Z_2}$$  \hspace{1cm} (2.16)

Combining (2.15) and (2.16), the equivalent reflected impedance of the secondary back on to the primary is expressed as:

$$Z_r = \frac{\omega^2 M^2}{Z_2}$$  \hspace{1cm} (2.17)

where the secondary impedance is:

$$Z_2 = R_L + j\omega L_2$$  \hspace{1cm} (2.18)
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The overall impedance of the secondary side for both series and parallel compensated secondary side can be expressed as [67]:

\[ Z_{s} (\text{series}) = j\omega L_s + \frac{1}{j\omega C_s} + R_L \] (2.19)

\[ Z_{s} (\text{parallel}) = j\omega L_s + \frac{1}{j\omega C_s + \frac{1}{R_L}} \]

\[ = \frac{R_L}{1 + (\omega C_s R_L)^2} + j \left[ \omega L_s - \frac{\omega C_s R_L^2}{1 + (\omega C_s R_L)^2} \right] \] (2.20)

Since the secondary network depends on the selected compensation topology, the reflected resistance and reactance at the resonant frequency onto the primary side is different as given in Table 2-1 [64].

<table>
<thead>
<tr>
<th>Tuning Topology</th>
<th>Reflected Resistance</th>
<th>Reflected Reactance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Series</td>
<td>( \frac{\omega_0^2 M^2}{R_L} )</td>
<td>0</td>
</tr>
<tr>
<td>Parallel</td>
<td>( \frac{M^2 R_L}{L_2^2} )</td>
<td>( - \frac{\omega_0 M^2}{L_2} )</td>
</tr>
</tbody>
</table>

It can be seen from Table 2-1 that the reflected impedance for the series-tuned secondary coil is pure resistive, while the parallel-tuned secondary coil reflects both resistive and capacitive components. This is one of the major differences between the series and parallel compensated secondary inductance. The reflected impedance can then be compared and combined with the impedance of the primary coil. The power transfer from the primary to the secondary is simply the reflected resistance (real part of the reflected impedance) multiplied by the squared primary current as given by [64]:

\[ P_s (\text{series _tuning}) = \left[ \frac{\omega_0^2 M^2}{R_L} \right] I_i^2 \] (2.21)

\[ P_s (\text{parallel _tuning}) = \left[ \frac{M^2 R_L}{L_2^2} \right] I_i^2 = \frac{M^2}{L_2^2} R_L I_i^2 \] (2.22)
2.6 Contactless slipring system characteristics

2.6.1 Effect of air gap on the system

Whenever an air gap is introduced into a magnetic path, the magnetic field lines will fringe outward as they cross the air gap as shown in Figure 2-11. These fringing fields increase the effective cross-sectional area $A_g$ of the air gap. One way to correct this effect is by adding the length of the gap to the dimensions of the core centre-pole cross-section. To avoid what could be a significant error, the inductance calculation must be based upon the effective gap area rather than the actual centre-pole area. A range of experimental methods have been developed for correcting this effect. For a core with a rectangular centre-pole (cp) with cross-section dimensions $(a$ and $b$), the effective gap area, $A_g$ is approximately [68]:

$$A_{cp} = a \times b ;$$

$$A_g \approx (a + l_g) \times (b + l_g)$$

(2.23)

Similarly, for a round centre-pole with a diameter $D_{cp}$:

$$A_{cp} = \frac{\pi}{4} D_{cp}^2 ;$$

$$A_g \approx \frac{\pi}{4}(D_{cp} + l_g)^2$$

(2.24)

The above correction methods should be implemented when modelling a magnetic circuit of a contactless slipring system, to achieve the desired inductances. This has been presented in Chapter 3 (Section 3.5).

![Fringing magnetic field](image)

Figure 2-11: Fringing flux lines around the air gaps.
The above mentioned fringing field is possibly to cause two problems: 1) the development of eddy currents in the nearby turns, and 2) the electromagnetic emissions. The generated eddy currents oppose the main current flow in the coil, and as a result reduce the effective cross-sectional area of the conductor and accordingly increase the effective resistance of the conductor. This leads to excessive heating of the portion of the coils that is next to the air gap. Therefore, the generated eddy currents are responsible for high temperature by causing localized heating in the winding turns that are next to the air gap which may reduce the overall efficiency of the system. Moreover, due to magnetic field variations, variable electric field would be created and the result is an electromagnetic emission around the air gap [69].

It is worth noting that the area around the air gaps is very responsive to metal and ferrite objects, which may change the reluctance of the magnetic circuit, and the equivalent inductances of the primary and the secondary coils.

2.6.2 Effect of rotation on the system

Since in a contactless slipring system, one side is stationary and the other may rotate, it is important to design a system that is not affected by the rotation. This is because the rotation speed is not controlled or determined by the slipring system and it may be slow as in the case of wind turbine applications or fast as in excitation systems of power generators.

From Eqns. (2.7) and (2.9), it can be observed that controlling $V_{oc}$ and $I_{oc}$ is highly related to the mutual inductance between the two sides [70]. Thus, in some of the IPT applications such as electrical vehicle charging systems, the magnetic coupling between the primary and the secondary side is not fixed and varies with the relative position of the vehicle. This results in a variable mutual inductance ($M$) and variable electrical properties of the outputs for the system. Therefore, a special control needs to be designed for the output power flow in EV’s applications [61, 62]. In the case of contactless slipring systems, because of the unique circular geometry, there is a complete overlap between the primary and the secondary coils for any angle of rotation which results in a fixed air gap between the two sides. This signifies that the mutual inductance between the primary and the secondary sides is constant and not affected by the relative rotation. Hence, the power transfer property from the primary to the secondary does not vary as the secondary rotates.
2.6.3 Windings inductances

Due to the existence of air gap between the primary and the secondary sides, contactless slipring system reveals a low magnetic coupling coefficient. This results in storing a portion of the total supplied energy into the magnetic structure itself (mainly within the air gap) rather than transferring it to the secondary pickup coil. There are two basic energy storing elements; the magnetizing inductance and the leakage inductance. The magnetizing inductance of such a contactless system is much lower than that of a compactly coupled transformer. As a result, in addition to the reflected load current, the primary winding carries an increased magnetizing current which increases the primary side conduction losses. The air gap could be reduced in order to increase the magnetizing inductance and avoid the high magnetizing current. Moreover, because of the availability of alternative magnetic-flux paths (air space) rather than the ideal core path, contactless slipring exhibits increased leakage inductance in both sides. The increased leakage inductance, which appears in series with the coil, causes a voltage drop, resulting in a reduced voltage gain [3].

2.6.4 Windings capacitances

The parasitic winding capacitances need to be considered when the system is operating at a very high frequency. The coils capacitance can cause three possible problems: 1) drive the system into unwanted resonance, 2) produce large primary current spikes when operating from a square wave source, and 3) produce electrostatic coupling to other circuits. During operation, different voltage slopes arise almost everywhere. These slopes are caused by a large variety of capacitance throughout the system, due to the coil configurations and how they are placed throughout the magnetic structure. These capacitances can be categorized into four groups: 1) capacitance between turns, 2) capacitance between layers, 3) capacitance between the primary and the secondary coils, and 4) stray capacitance [68].

Turn-to-Turn capacitance is due to the stored energy in the form of electric field between two neighboring turns being in proximity. Usually this type of capacitance is small and its effect can only be seen at very high frequencies and can be reduced by using bank winding techniques or changing the wire insulation to one with a lower dielectric constant. The capacitance between layers of the coils is the major portion of the overall parasitic capacitance of the system. The layer capacitance could be minimized by dividing the primary and secondary coils into sections (sandwiched design) and increasing the amount of insulation between coils. Coil-to-Coil capacitance forms between the primary and the
secondary coils and it can be reduced by increasing the amount of insulation between the coils. This will decrease the amount of capacitance, but again, this will increase the leakage inductance. Adding the known Faraday’s Shield between the primary and the secondary can noticeably reduce this type of capacitance. However, in a contactless slipring system application because of the required physical separation, it is difficult to add extra materials in the air gap between the coils. Stray capacitance is very important to minimize, as it can generate asymmetry currents and could lead to high common mode noise. Stray capacitance is similar to coil-to-coil capacitance except that the capacitance is between the system’s coils and other coils next to the surrounding circuitry. Stray capacitance can be minimized by using a balanced winding, or using a copper shield over the entire coil.

### 2.6.5 Windings resistances

The coil’s resistance is the product of two components. The first is the usual DC resistance of the wires, and the second is the result of the high frequency effects and the resulting resistance is termed effective resistance. The effective resistance increases due to the proximity and skin effects and can be controlled by reducing the number of primary and secondary turns and by carefully selecting the wire. The coil resistance losses have a direct impact on the efficiency as compared to the inductances and windings capacitances which affect the system efficiency indirectly. The conduction losses can be significant due to the increased primary current and winding resistances ($I^2R$). The effective resistance can be calculated using Hurley’s rule as given by [3]:

$$R_{eff} = R_{dc} + \frac{\phi}{3} \Delta^4 R_{dc} \left[ \frac{I'_{rms}}{\omega I_{rms}} \right]^2$$  \hspace{1cm} (2.25)

where $I_{rms}$ is the rms value of the current and $I'_{rms}$ is the rms value of the derivative of the current, $R_{eff}$ is the effective resistance of the coil, $R_{dc}$ is the DC resistance and:

$$\phi = \frac{5p^2 - 1}{15}$$  \hspace{1cm} (2.26)

where $p$ is the number of coils layers and:

$$\Delta = \frac{d}{\delta_0}$$  \hspace{1cm} (2.27)
where \( d \) is the thickness of the layer and:

\[
\delta_0 = \sqrt{\frac{2}{\omega_0 \mu_0 \sigma}}
\] (2.28)

where \( \omega_0 \) is the angular frequency of the current, \( \mu_0 \) is the magnetic permeability of free space and \( \sigma \) is the electrical conductivity of the conductor.

Note that the above mentioned relationships are for a single strand type of wire and not a multi-strands wire such as Litz wire. In the case of Litz wire, the coil resistances can be calculated from Cheng’s method as presented in [71].

### 2.6.6 Losses associated with the magnetic coupling structure

Two types of power losses are associated with the magnetic coupling structure of a CSS: 1) core losses, and 2) coils losses (conduction losses) as detailed below:

#### 2.6.6.1 Core losses

Typically, core losses are divided into two groups: hysteresis and eddy current losses. Hysteresis loss increases with higher frequencies as more cycles are undergone per unit time and is associated with the applied voltage and its frequency. While, eddy currents loss is proportional to the square of the applied voltage and it is independent of frequency. In the case of core loss due to magnetic hysteresis, all ferromagnetic materials tend to retain some degree of magnetization after disclosure to an external magnetic field. This tendency to stay magnetized is called hysteresis. And it takes a certain amount of energy to overcome this opposition to change every time when the magnetic field produced by the primary coil changes polarity (twice per AC cycle). This type of loss can be mitigated by choosing a core with low hysteresis (with a ‘thin’ B-H curve), and designing the core for minimum flux density (large cross-sectional area). The hysteresis losses per unit volume are approximated by the following relationship [72]:

\[
P_h = k \cdot f_{sw}^a \cdot \left( \frac{\Delta B}{2} \right)^d \left( \frac{W}{m^3} \right)
\] (2.29)

where \( k, a, \) and \( d \) are material related constants, \( f_{sw} \) is the switching frequency and \( \Delta B \) is the flux excursion.
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The second part of the core loss is due to magnetic effects of the core material. This type is resistive power dissipation due to the induced eddy currents in the core material. Usually, when a low resistivity material is placed in a variable magnetic field, electric currents are induced in the material similar to the currents induced in the secondary coil. These induced currents tend to circulate through the cross-section of the core perpendicular to the main flux path. Their circular motion gives them their unusual name: like eddies in a stream of water that circulates rather than move in straight lines. These ‘eddy currents’ must overcome an electrical resistance as they circulate in the core. To overcome the resistance offered by the core, they dissipate power in the form of heat. Hence, there will be a source of power loss in the core that is difficult to eliminate and largely depends on the conductivity of the material and the magnetic field density. Thus eddy currents can be minimized by selecting magnetic core materials that have low electrical conductivity. Ferromagnetic materials such as ferrites have a considerably low electrical conductivity making them more suitable for high frequency applications. The eddy current losses in a volume like that of Figure 2-12 are approximated by the following relationship [72]:

\[
P_{eddy} = \frac{B^2 f_{sw}^2 d^2}{\rho} \quad \left( \frac{W}{m^3} \right)
\]

where \(\rho\) is the resistivity of the material and \(f_{sw}\) is the switching frequency.

![Figure 2-12: Eddy currents created in a magnetic material.](image)

2.6.6.2 Conduction losses

In practice, there will be always power dissipated in the form of heat through the resistance of current-carrying conductors. Increasing the gauge of the wire can minimize this loss, but only
with extensive increases in cost, size, and weight. Generally, the conduction losses of the coils classified into two groups: 1) the DC, and 2) the AC losses. The DC losses are the result of current I (rms value of the current) flowing through the wire with resistance R, power \(I^2R\) is dissipated and accordingly, developing heat in the wire. On the other hand, normally due to the varying magnetic field around the coils, the current distribution in the conductors is not uniform. The self-inductance of the conductor, as well as the magnetic field created by adjacent turns of the same coil, redistributes the current flow within the wire and reduces the effective cross-sectional area of the conductor and accordingly increases the AC resistance. These effects are known as skin and the proximity effects and they become stronger at higher operating frequencies as detailed below:

### 2.6.6.2.1 The skin effect

When a wire is carrying a DC current, the current is distributed uniformly over all the effective area of the wire (left side of Figure 2-13). This current distribution is non-uniform when the wire is carrying an alternating current as the intensity of the current and the magnetic field in a conductor change. The change in the magnetic field creates an electric field which opposes the change in current intensity. This opposing electric field is called counter EMF. The counter EMF is strongest at the centre of the conductor, and forces the conducting electrons to the outside of the conductor, as shown in the right side of Figure 2-13. As a result the self-induced back-EMF will be greater towards the centre of the conductor and causing the current density to be less at the centre than the conductor surface. This extra concentration at the surface is known as ‘skin effect’ and results in an increase in the effective resistance of the conductor and thereby reduces the available cross-sectional area for electron flow.

![Figure 2-13: Current distribution in a wire for DC current (left) and AC current (right).](image-url)
In fact, the skin effect is the tendency of high frequency current density to be highest at the surface of a conductor and then to decay exponentially toward the center. Because of this skin effect, larger power losses is caused for a given rms value of AC current than the losses when the same value of DC current is flowing through the conductor as shown in Figure 2-14. Increased AC resistance due to the skin effect can be mitigated by using specially woven Litz wire. The AC current density $J$ in a conductor decreases exponentially from its value at the surface $J_s$ according to the depth $d$ from the surface and expressed as:

$$J = J_s e^{-d\delta}$$

(2.31)

where $\delta$ is called the skin depth. The skin depth is thus defined as the depth below the surface of the conductor at which the current density has fallen to 1/e (about 0.37) of $J_s$. Typically it is well approximated as:

$$\delta = \frac{2\cdot\rho}{\omega \cdot \mu}$$

(2.32)

where $\rho$ is the resistivity of the conductor, $\omega$ is the angular frequency of the current and $\mu = \mu_r \mu_0$ is the absolute magnetic permeability of the conductor.

### 2.6.6.2.2 The proximity effect

When another conductor is brought into proximity to one or more other nearby conductors, such as within a closely wound coil of wire, the distribution of current within the first conductor will be constrained to smaller regions. This is shown in Figure 2-15 in blue. Field intensity is no longer uniform around each conductor surface. The resulting current crowding is termed the ‘proximity effect’. This crowding gives an increase in the effective resistance and increases with the frequency. Consequently, increases the AC resistance of the conductor which results in higher $I^2R$ losses in the wire.
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2.7 Review of power converter topologies

2.7.1 Fly-back converters

This topology is the simplest converter with the least components for construction and usually used to drive coupled inductors. As a result of coupled inductors, the fly-back stores energy into the magnetic component and releases it to the secondary coil. Figure 2-16 shows a basic circuit of the fly-back converter that can be considered for contactless power transfer applications. However, because of the voltage stress on the main switch as well as incomplete utilization of the magnetic core (due to the unidirectional flux excursion); this topology is usually considered for low power applications [73].

2.7.2 Resonant converters

Resonant converters are the mostly used in contactless power transfer applications for their several advantages [54]. Soft-switching feature is the main advantage of this family which greatly reduces switching losses and EMI. Soft-switching techniques enable high-frequency operation and consequently reduce the overall system size. Resonant converters are
commonly used in applications with a long air gap in the range of millimetre to several centimetres between two parts and they can deliver power in the range of kW [24]. Using sine-waves, the secondary can be tuned to the primary frequency and picks up power from the primary coil with low harmonic content and reduced EMI. Resonant family comprises current-fed and voltage-fed converters in two basic push-pull and full-bridge topologies [74]. In fact, the two upper switches of a full-bridge converter are replaced by two input inductors or a splitting transformer in a push-pull topology [75, 76]. Figure 2-17 shows a commonly used current-fed push-pull resonant converter for contactless power transfer applications [77]. The current-fed converter has the advantages of maintaining high frequency operation, since only the real current is sourced through the switches and the high reactive primary current is confined to the resonant tank. In voltage-fed converters, the primary inductor can be increased beyond the input voltage of the power supply which makes them preferable for electrical vehicle applications with long primary tracks. However, high resonant current flows through both the switch and the track, which lowers the system’s efficiency.

![Figure 2-17: Current-fed push-pull converter.](image)

Figure 2-18 (a) illustrates an LCL converter topology [78]. This converter has the benefits of a current source property when driving and presents unity Displacement Power Factor (DPF) to the input stage [79]. An improved current-fed CLC converter is proposed in [54] based on current-fed parallel-series resonant converter with an additional ‘π’ network introduced at the output of the supply feeding the primary and the load as shown in Figure 2-18(b). This converter overcomes the disadvantages associated with the typical current-fed converter such as possible primary current variation, frequency instability, etc. and improves the system performance.
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2.7.3 Self-resonant converters

One of popular member of this category is the class-E converter proposed in [80] for biomedical power supplies for its ability of operating at MHz and efficiencies beyond 90% (see Figure 2-19). Moreover, the low harmonic output, low sensitivity and low component stress has truly made this converter the most popular for many TET systems [81, 82].
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![Class-E tuned power oscillator with an inductive link](image)

**Figure 2-20:** A class-E tuned power oscillator with an inductive link.

The authors of [83] worked on the concept of a class-E self-resonant amplifier with a single switch that utilizes the parasitic inductance of an inductive coupler for resonance (see Figure 2-20). The amplifier uses an auxiliary winding to feedback to the switch and maintains the oscillations and to track the resonant frequency when the inductive coupler's characteristics varies. The circuit is self-resonant with a feedback loop to maintain the resonance at the resonant tank frequency. To regulate the output voltage an additional driving circuit is used to inject a current and alter the biasing of the switch gate.

### 2.7.4 AC-AC converter

Figure 2-21 demonstrates a direct AC-AC power converter proposed for IPT applications in [84]. As the name states, this topology converts the low frequency AC power to a high frequency AC power directly without any rectification to generate high frequency current. Therefore, the commutation and synchronization of the source voltage and the resonant loop branch need to be considered. To determine the switch operation, it is necessary to identify the polarity of the input voltage and the resonant current. The fundamental operation of the ac-ac converter is based on the concept of free oscillation and energy injection control as detailed in [84]. In principle, if there are no power losses in the resonant tank (when $R_{eq}=0$), the circuit can oscillate infinitely without the need of energy injection. However, in practice power loss always exists because of ESRs and the presence of any reflected load from the secondary side. When more energy is required, the resonant tank is connected to the power supply by the converting network. Such an operation based on the discrete quantum energy injection and free oscillation is very different from normal converters. In consequence, the current is sensitive to the load condition, polarity and phase of the input AC voltage.
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![Figure 2-21: Direct AC to AC converter.](image)

### 2.8 Review of power flow regulation methods

Controlling the power flow is essential for contactless slipring systems. Because of the large air gap and relative rotation between the two sides, there is a possibility of power flow variation. The aim of power regulation is to control the amount of required power using a reference, rather than delivering an arbitrary amount of power to the load. There is a wide range of power flow controllers that are proposed and implemented over time by the researchers as shown in Figure 2-22 [37].

![Figure 2-22: Block diagram of power flow controllers used in IPT systems.](image)
2.8.1 Primary side power flow control

To control the power flow from the primary side, normally a feedback signal such as output voltage of the secondary side using wireless communications channels is required. However, due to the limited bandwidth of resonant circuits and delays in wireless communication channel; primary controllers suffer from slow transient response as the output voltage can not be regulated in sub-millisecond time intervals. Moreover, since the wireless communication system is operating very close to the IPT system, any switching noise from the high power electronics devices can easily cause interference. To overcome these problems, new approach of prediction and construct the output voltage of the secondary by measuring the reflected impedance on the primary side is proposed in [85]. However, this technique can only be used in applications where there is one secondary coil with a fixed coupling as per its control requirement. Primary side of power control is generally achieved by either primary current control or operating frequency control as detailed below:

2.8.1.1 Current control

The magnetic field of an IPT system can be controlled indirectly by controlling the primary current, in turn, controlling the power delivered to the secondary side. Controlling the primary current can be achieved in different ways depending on the inverting bridge topology. For the current-fed push-pull topology, the simplest way is to control the input DC voltage to the inverter [86]. Input voltage control is usually implemented using an extra buck converter and adjusting its duty cycle. For a voltage-fed series tuned converter on the other hand, the more common method is to use phase control [87]. Primary current control is an efficient method of power control as the controller can reduce the current when little or no power is required by the load. The reduction in current reduces the I^2R losses as less current is flowing in the overall system which results in a higher overall operating efficiency.

2.8.1.2 Frequency control

The power delivered to the load can also be controlled by controlling the operating frequency of the system [4]. The primary frequency controllers are divided into fixed and variable frequency control schemes as discussed below:

2.8.1.2.1 Fixed frequency control

This is an open-loop control and the controller forces the operating frequency to a fixed value regardless of any feedback from the load and other system parameters. With a fixed-frequency control, the switching frequency remains constant and the driver adjusts the on-
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time of the switch. This approach is known as Pulse Width Modulation (PWM) and offers the advantage of easy paralleling and synchronising of several converter modules to achieve higher output power [88]. The overall design of the converter for a fixed frequency operation is easier and the losses can be accurately predicted. With a fixed primary frequency, the secondary resonant frequency is designed to be equal to the operating frequency in order to ensure maximum power transfer to the secondary. Fixed-frequency controllers thus have the advantage that a compensated secondary will always run at its correct tuned frequency. However, the power supplies need to be overrated since the displacement power factor (DPF) will not be unity due to varying operating conditions (system parameter variations). For example the reflected impedance from the secondary back to the primary will change from that expected and causes the primary inductance to be detuned. Therefore the power supply needs to be overrated to supply the extra VAR loads on the system.

2.8.1.2.2 Variable frequency control

This method may be easily implemented in a stand-alone converter, however it makes it difficult to synchronise more parallel modules. The controller allows the operating frequency to vary over a predetermined range depending on the primary resonant tank parameters. This is normally achieved by implementing switched inductors and capacitors (variable inductors and capacitors), to adjust the resonant frequency. However, the secondary tuning would also vary as the secondary side is normally tuned to the primary resonant frequency. By controlling the primary operating frequency, the inverter output voltage and current can be controlled to be in phase allowing the converter to operate close to unity power factor. However, one of the major drawbacks of variable frequency controllers is that it is hard to achieve high secondary quality factor Qs as primary operating frequency may vary. Another drawback of variable frequency controllers is the frequency stability issue, which is called the ‘Bifurcation phenomenon’ [48, 64]. In fact, variable frequency controllers can have more than one stable operating frequency at high Qs, and the controller may jump between these operating points. In such cases the primary operating frequency will shift considerably away from the secondary resonant frequency. These frequency shifts result in substantial secondary detuning and significantly reduce the power transferred to the secondary.

2.8.2 Secondary side power flow control

Generally, the secondary converter is used for power conditioning and control according to the load requirements. This includes a few steps to direct suitable power from the terminals of
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the secondary coil to the connected electric load as demonstrated in Figure 2-23. Although there are many types of secondary side power controllers proposed over time as depicted in Figure 2-22, in this section only buck and boost types of controllers which are mostly used for contactless slipring systems are presented.

![Figure 2-23: Block diagram of the secondary side of a contactless slipring system.](image)

### 2.8.2.1 Buck controller

Figure 2-24 shows the general circuit diagram of the buck type of secondary controller. This type of controller is used with a series tuned secondary inductance. A series tuned \( L_S \) behave like a constant voltage source and the power taken from the secondary depends on how much current is flowing to the load. The behaviour of this controller is exactly like a buck converter. The buck converter usually steps down a higher input voltage to a lower output voltage while increasing the input to output current ratio. In this way, the output voltage of the buck converter can be regulated to any reference lower than the input voltage. In practice the buck converter controls the duty cycle to regulate the output voltage to a desired value, and the current delivered by the secondary directly depends on the load after the buck converter. The output of the buck converter will behave similarly to an ideal voltage source.

![Figure 2-24: Series-tuned secondary with a buck type of power controller.](image)
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2.8.2.2 Boost controller

The boost topology on the other hand, is usually used with a parallel tuned secondary inductance as shown in Figure 2-25.

![Figure 2-25: Parallel-tuned secondary with a boost type of power controller.](image)

There are two different types of operation for the boost controller - fast switching and slow switching as discussed below:

Fast switching refers to the switching frequency of the controller to be relatively high and is not related to the system operating frequency. Under fast switching conditions, the controller fully acts as a boost converter, increasing the input to output voltage ratio and decreasing the input to output current ratio. To regulate the output voltage, the output of the secondary coil is considered as a voltage source, with one limitation that the output current of this voltage source is limited by the short circuit current of the secondary multiplied by the duty cycle. Likewise, slow switching refers to the switching frequency of the boost converter switch to be relatively low (typically 1-100Hz). Under slow switching, the controller acts very differently as compared to the fast switching operation. Since the output of the secondary is a current source, the switching occurs depending on whether the load needs the current. When the switch is open, the output current of the secondary charges $C_o$ and the output voltage will rise. And under the switch closed condition, the secondary will be decoupled from the primary as shorting a current source would dissipate no energy, while the output voltage drops as $C_o$ discharges through the load resistor. Generally hysteresis type of controller is required to regulate the output voltage within certain limits.

Unlike the fast switching operation, the slow switching operation completely collapses the resonant capacitor voltage when the switch is closed. This voltage across the resonant capacitor builds up when the switch is opened. The resonant energy will be completely transferred to the DC inductor and dissipated in the internal resistance in the switch and
rectifier diodes each time it is shorted. However the switching losses are reasonable as the frequency is low. For fast switching, the resonant capacitor voltage is not collapsed; however it is associated with greater switching losses in the boost converter stage as the switching frequency is much higher. When the circuit is not requiring full power, fast switching operation would always keep the resonant circuit VAR’s at a minimum while keeping the output voltage constant. However, the slow switching topology would always alternate the resonant circuit between full power and decoupled conditions. From this, the steady state losses in the resonant tank would be lower for a fast switching controller as lower resonant currents would lead to a squared timeless dissipation in $I^2R$ losses in the ESR of the secondary inductor and tuning capacitor, while the slow switching operation would only be linearly proportional.

2.9 Summary

An overview of the technologies involved in power transfer for rotary applications has been presented in this chapter. Their advantages and disadvantages are discussed in relation to IPT-based contactless slipring systems. The IPT-based contactless slipring, being the subject of this research, has been reviewed in details. In rotary applications, mechanical sliprings are widely used for delivering power to an electrical load on a rotary shaft. However, because of the physical contacts between the stationary brushes and rotating metal rings, frequent maintenance is required due to mechanical and electrical wear and tears. Capacitive power transfer is another option for rotary applications, but it is limited to very low power levels. In comparison, currently only the IPT-based contactless sliprings can transfer power in kW range to a rotating shaft across large air gaps constrained by mechanical clearance.

After a general review of power converter topologies used for IPT systems, primary and secondary power flow regulation methods have been presented. Resonant converters are the mostly used in contactless power transfer applications for their soft-switching characteristics which greatly reduces switching losses and EMI. The primary side power flow regulation is normally achieved either by controlling the frequency or the current flowing through the resonant inductor. To control the power flow from the primary side, normally a feedback signal such as output voltage of the secondary side using wireless communications channels is required. However, due to the limited bandwidth of resonant circuits and delays in wireless communication channel; primary controllers suffer from slow transient response. The two
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buck and boost secondary side power flow regulators are mostly used for contactless slipring systems.
Chapter 3: Single-phase contactless slipring systems based on pulsating magnetic field

3.1 Introduction

This chapter is about single-phase contactless slipring systems based on pulsating magnetic field. In principle, pulsating magnetic field is stationary in its location and its magnitude varies with time along its magnetic axis. A very familiar electrical machine based on this theory is the electrical transformer. The focus of this chapter is the analysis of the single-phase magnetic coupling structure of the sliprings, and the power supply for driving it is presented in details in Chapter 4, and a full system-based 3D-FEM analysis as well as practical implementations of the proposed single-phase systems is presented in Chapter 5.

After a brief introduction to the basic principle of operation of contactless slipring systems, existing single-phase magnetic structures are presented in this chapter. Then improved magnetic coupling structures with high magnetic coupling coefficients are presented, followed by accurate modeling of the physical structure of a single-phase contactless slipring system including the core losses. Then, the mainly used methods for measuring the magnetic coupling level in IPT systems are presented. In addition, a new method is proposed which can determine real time magnetic coupling coefficient for an IPT system under operating conditions. Finally a new method for improving the magnetic field coupling of the single-phase contactless slipring systems by varying the ratio of copper and ferrite usage is proposed to increase the coupling coefficient within the existing system dimensions with U, E and pot-core geometry configurations.

3.2 General diagram and basic principle of operation

Figure 3-1 shows a general diagram of a pulsating magnetic field-based contactless slipring system. If an alternating current flows through the primary coil, it produces a pulsating mmf wave, whose amplitude and direction depends on the instantaneous value of the current flowing through the coil. Figure 3-2 illustrates the mmf distribution in the space at various instants due to an alternating current flowing through the primary coil. In fact in such a magnetic field, the magnetic axis of the field is fixed in its location and the magnitude is varying with the time along its magnetic axis. This is the main difference of this type of
Chapter 3: Single-phase contactless slipring systems based on pulsating magnetic field

magnetic field with an axial travelling or rotating magnetic fields presented in Chapter 6 and Chapter 7. In the case of axial travelling/rotating magnetic field, the magnitude is constant and the magnetic axis of the resultant magnetic field moves in location.

![Diagram of a single-phase contactless slipring system](image1)

**Figure 3-1:** General diagram of a single-phase contactless slipring system.

![Mmf distribution in space of the primary coil](image2)

**Figure 3-2:** Mmf distribution in space of the primary coil for different instants.

When the secondary coil is located in the field generated by the primary current, an emf is induced at its terminals. The presence of the time varying field within the secondary coil induces currents which can be converted to power and supplied to the load. The induced voltage in the secondary coil then is connected to the secondary converter. The secondary converter realizes rectifying, regulating and power converting according to the load requirements. Note that the primary and the secondary coils shown in Figure 3-1 are for illustration purposes only, and in an actual system they can have various rotary configurations as detailed next.
3.3 Existing magnetic coupling structures

Depending on the application, the magnetic coupling structure used in contactless slipring systems has different geometry/size that defines its performance. The magnetic coupling structures can be categorized into two main groups namely: non-through-hole and through-hole structures.

3.3.1 Non-through-hole structures

Non-through-hole contactless slipring systems are designed as coaxial or adjacent layouts normally using pot-core geometry as shown in Figure 3-3[2, 3, 89]. As compared to through-hole layouts, inherent symmetry of pot-core geometry with the fixed air gap makes the magnetic flux linkage smoother between the primary and the secondary coils and results in constant electrical characteristics for the system [3, 21]. In the case of through-hole structures, because the number of cores used in the primary is not the same number of cores used in the secondary side, the flux linkage between the two sides is not smooth as in the pot-core types. However, pot-core structures are usually associated with more flux leakage because of their middle limb same as an E-shape of ferrite [90, 91]. Therefore, they are suitable for small air gaps where the distance between the side limbs and the middle limb is greater than the air gap between the two sides. Another drawback of pot-core geometry is that they can be designed to be placed at the end of the shaft as shown in Figure 3-3 which in some cases modifying the existing shaft may be required.

![Diagram of non-through-hole contactless sliprings](image)

Figure 3-3: Non-through-hole contactless sliprings (Top) Coaxial and (Bottom) Adjacent designs.

3.3.2 Through-hole structure with coaxial layout

Figure 3-4 shows a typical through-hole contactless slipring system with a coaxial configuration. The whole structure is made of several ferrite cores to provide a complete ring around the existing shaft of the system. Normally there will be no need to modify the existing

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shaft. This is the main advantage of a through-hole type that can be designed for any shaft’s diameter, while the pot-core types are limited to certain diameters. However, existence of shaft in a through-hole system is a concern. If there is no shaft or for a non-metal shaft around the system, there is no need for adding magnetic shielding to protect the shaft from the heat generated by stray magnetic fields [92]. On the other hand, when a metal shaft is a part of the system, a portion of the stray flux lines pass through the shaft. This creates power loss due to the eddy currents induced in the shaft. If the amount of the power lost in the shaft is considerable, it may heat up the shaft and creates more damage to the system gradually. To overcome this, usually a passive magnetic shielding is placed between the shaft and the magnetic coupling structure as shown in Figure 3-4 [93], [94]. One of the common materials used is aluminium. Although, aluminium material avoids the magnetic field passing through the shaft, it creates more loss for the system and usually it is one of the design trades-offs. A new method of magnetic shielding using flexible ferrite materials is proposed in this thesis (in Chapter 5) which is not associated with power losses as seen in aluminium [95]. The contactless slipring shown in Figure 3-4 has been developed by PowerbyProxi Ltd and currently installed as a trial for a wind turbine pitch control [26, 29].

Figure 3-4: Through-hole contactless slipring with coaxial layout.

3.4 Improved magnetic coupling structures

3.4.1 Face to Face structure

From Figure 3-4 it can be seen that in a through-hole contactless slipring with coaxial layout, the primary side encircles the secondary side fully, which results in a system with a fixed air gap and asymmetrical primary and secondary. In such a design, in addition to individual
design procedures for each side, any changes in the air gap requires a complete system redesign, which can be time consuming and costly. Moreover, if there is a barrier wall along the system shaft, the coaxial structure would not work. To overcome this, a new through-hole type contactless slipring with identical primary and secondary magnetic structures proposed as shown in Figure 3-5. As it can be seen the face to face configuration of this system allows for flexible air gaps, and can work even when a barrier wall is located between the primary and secondary sides [95]. One of the clear advantages of this structure is that it accommodates the possibility of linear and rotary movement for the secondary as compared to the existing coaxial contactless slipring. Moreover, its face to face feature provides the possibility of using such a design with a primary located outside and a secondary in the other side of a barrier wall. However, this is possible if the barrier wall is made of a non-magnetic and non-conductive material.

![Figure 3-5: Face to Face through-hole contactless slipring system.](image)

### 3.4.2 Double-Stator structure

Although the above coaxial and face to face contactless sliprings solve the direct contact problem associated with mechanical slipring assemblies. Because of the existence of the air gap between the two separated sides, they have a lower magnetic coupling coefficient compared to traditional transformers [25, 69]. As a result, it is a challenging task to achieve high power transfer capability from a single contactless system to meet the load requirements. This may lead to the need of installing multiple units to meet high power demands. Apart from the increased component counts, a system with multiple units needs extra space to avoid cross coupling between the adjacent coils, which would further complicate the design and
increase the system size. In this section an improved face to face system with a novel double-stator layout for contactless slipring systems to increase the power transfer capability of the CSS with a single power supply is proposed [96] as shown in Figure 3-6. It is still a through-hole configuration, but there are two stationary primary parts, and the secondary is placed between them. The two primary sides are identical and connected in series with aiding polarities to enhance the magnetic field generation and accordingly enhance the magnetic coupling coefficient. The two primary parts are fixed to the housing, while the rotating coil is fixed on the shaft by a non-magnetic and non-conductive material coil holder made of fiber glass, to prevent any electrical or magnetic influences on the performance of the system. There is no core used in the secondary part which is a great advantage as ferrite material may break during rotation. Therefore, only a primary non-rotating core is used in the stationary part of the magnetic structure. This helps to eliminate the magnetic force interaction between permeable ferrite surfaces because all the ferrite material remains fixed [92, 97]. With a fixed air gap, the secondary coil has a constant magnetic coupling with the primary for any angle of rotation. This leads to a constant mutual inductance between the two sides, which help to simplify the system design.

The two primary parts provide interleaved windings layout which cause a significant difference in leakage inductance reduction because of mmf distribution. As a result of the interleaving feature the leakage inductance is reduced and improves the magnetic coupling.
between the two sides [98, 99]. From the mmf distribution curves shown in Figure 3-7(a) it can be seen that unlike the typical design (see Figure 3-7(b)), the mmf distribution has been shifted to be a symmetrical curve on the axis. The maximum magnetizing force is reduced to half of the primary current which results in a significant reduction in the energy enclosed in the winding space for the double-stator design. Such a double stator structure can increase the power transfer from the same space as compared to a coaxial/face to face CSS as presented in Chapter 5.

![MMF distribution curves](image)

**Figure 3-7: mmf distribution: (a) Double-stator system (b) Typical system.**

### 3.4.3 Single air gap structures

In contactless slipring systems similar to other IPT systems, one way to improve the mutual coupling between the two sides is to design a system with a small air gap. Low flux leakages as well as good magnetic coupling as a result of the reduced air gap, has a great impact on increasing power transfer to the load side [62]. In this section two magnetic structures are proposed for contactless slipring systems with single air gap. The first design is a sandwiched layout and the other one is a non-sandwiched layout as shown in Figure 3-8 [92, 97, 100]. Only one air gap is cast-off for the ferrite core in order to improve the magnetic coupling between the coils, reduce the changes in the reluctance as well as reduce the magnetic flux leakages. Therefore, undesired electromagnetic emissions can considerably be reduced as a result of single air gap. The length of the air gap is correlated to the thickness of the coil holder. In other words, it can be as small as possible if the coil holder is thin and strong enough to hold the rotating coil, in turn, the associated losses and EMI would reduce.
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In these structures, there is no core used in the rotating part which simplifies the system design. This is one the advantages of the single air gap structures which avoid any breakage of ferrite material during rotation. Similar to the double-stator system, this helps to eliminate the magnetic force interaction between permeable ferrite surfaces because all of the ferrite material remains fixed. Ferrite core and stationary coils are fixed to the housing, while the rotating coil is fixed to the shaft by the means of the coil holder and rotate with the shaft. However, for the sandwiched structure the rotating coil is sandwiched between the two parts of the stationary coils. This is done by dividing the primary winding into two parts and...

Figure 3-8: Single air gap magnetic coupling structures: (a) sandwiched and (b) non-sandwiched.
connected in series, while the secondary coil is placed in between them as shown in Figure 3-8(a), while the non-sandwiched structure consist of a single primary coil as shown in Figure 3-8(b). Unlike the structures with double air gaps, due to the specific core geometry with a single air gap a good coupling factor is achievable. Moreover, the generated fringing flux and flux leakage is less than a contactless slipring system with two air gaps. To avoid the interaction with the fringing magnetic field, the coils are placed away from the air gap in the core window area. This results in reduction of the coils power losses as well as increases the voltage gain due to the less voltage drops across the coils [3].

3.5 Obtaining electrical model from geometry

3.5.1 Modelling without resistive components

Contactless slipring system poses unusual design restrains due to the air gap between the primary and the secondary sides. Having a reliable prediction and an accurate model of such a gapped structure is essential for design prejudgments before construction. T-equivalent circuit or mutual inductance model presented in Chapter 2 are used only for an already built system with identified inductances. Modelling a physical magnetic structure as an electrical network offers an improved analysis of the system performance, simulation, and derivation of mathematical equations to put a figure on each aspect of a magnetic component. This helps the designer to have a better understanding about the system and its associated limitations for any possible changes, in order to design the system so as to meet the application requirements. There are several methods such as testing /measurements and FEM analysis that can be used for design analysis. However, the measurement method can only be applied in practical systems and the finite element analysis is only suitable for preliminary calculations and simulations.

The proposed method here is obtaining the equivalent reluctance model by identifying different magnetic flux paths in the structure. Then the reluctance model is converted to an electric equivalent network using the duality rules between magnetic and electric quantities [57, 101, 102]. Figure 3-9 shows the flux paths of a typical contactless slipring system using UU arrangement. The primary current flows in the primary coil and sets up an mmf \(N_1I_1\) which is considered as a driving mmf. If a load is connected to the secondary coil, then the induced current in the secondary coil produces a back-mmf \(N_2I_2\) where its magnetic flux \(\phi_2\)
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opposes the primary flux $\varphi_1$ and thus subtract from it. The resultant of these two mmf’s is the working main flux $\varphi_m$ (mutual flux) is given by:

$$\varphi_m = \varphi_1 - \varphi_2$$

(3.1)

where:

$$\varphi_2 = \varphi_s - \varphi_m$$

(3.2)

Figure 3-9: Flux paths of the magnetic structure.

From Figure 3-9, the fringing fields around the air gaps will be ignored, except the effective gap area might be corrected to take the fringing field into account as presented in Chapter 2 (Section 2.6.1). Therefore, the total flux linked by each winding can be divided into two components: a mutual component $\varphi_m$ that is common to both coils and a leakage flux component ($\varphi_{lk1}$ and $\varphi_{lk2}$) that links only the coil itself. The reluctance of each region of the structure is calculated from its area, length and permeability ($\mathcal{R} = l/\mu A$), and inserted with its specific value into an appropriate location in the reluctance model as shown in Figure
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3-10. Note that the magnetic field sources (ampere-turns) of the coils are assigned to any discrete point where the flux is not divided. The series reluctances of the same core section with the same dimensions can be added together giving the total primary and secondary core reluctances as following:

\[ R_{cp} = R_{c1} + R_{c2} + R_{c3} \]  

(3.3)

\[ R_{cs} = R_{c4} + R_{c5} + R_{c6} \]  

(3.4)

\[ R_m = R_{c1} + R_{c2} + R_{c3} + R_{c4} + R_{c5} + R_{c6} + R_{g1} + R_{g2} \equiv R_{g1} + R_{g2} \]  

(3.5)

The air gap reluctances on the other hand, are much greater than the adjacent ferrite core sections. Thus the reluctance of the air gaps is governing the working flux and controls the magnetic flux of the core. Therefore, the core reluctances could be eliminated from the reluctance model as stated in (3.5). This leads to a simplified reluctance model as shown in Figure 3-11(a). A dual now needs to be created for the magnetic circuit of Figure 3-11(a) based on the duality relationships between the two networks. For instance, in a reluctance model, nodes are the meshes in the electrical model, open branch will be shorted, series will be parallel, mmf is a voltage source, and reluctance is permeance. Figure 3-11(b) shows a permeance network of the reluctance model after applying a network transformation between the two networks. Next, the permeance model of Figure 3-11(b) is converted to inductance values by multiplying them with the respective number of turns squared \( N_t^2 \) as a reference winding (Eqn.(3.6)). Thus, terminal \( V_1/N_1 \) and \( V_2/N_2 \) are multiplied by \( N_1 \) to become terminal voltage, and \( N_1 I_1 \) and \( N_2 I_2 \) are divided by \( N_1 \) to become terminal current.

\[ \begin{align*}
R &= \frac{Ni}{\phi} \\
L &= \frac{N\phi}{i}
\end{align*} \rightarrow L = \frac{N^2}{R} = N^2 P 
\]  

(3.6)

Based on the above relationship, the primary leakage and magnetizing inductances are expressed by:

\[ L_{lk1} = \frac{N_1^2}{R_{lk1}} = N_1^2 P_{lk1} \]  

(3.7)

\[ L_{m1} = \frac{N_1^2}{R_{g1} + R_{g2}} = \frac{N_1^2}{R_m} = N_1^2 P_m \]  

(3.8)

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Likewise, for the secondary side:

\[ L_{n2} = \frac{N_2^2}{R_{g2}} = N_2^2 P_{n2} \] (3.9)

\[ L_{m2} = \frac{N_2^2}{R_{g1} + R_{g2}} = \frac{N_2^2}{R_m} = N_2^2 P_m \] (3.10)

Taking the ratio of \( L_{m1} \) and \( L_{m2} \):

\[ L_{m2} = \frac{N_1^2}{N_2^2} L_{m1} = a^2 L_{m1} \] (3.11)

The mutual inductance on the other hand depends on both coils and can be expressed by:

\[ M = \frac{N_1 N_2}{R_{g1} + R_{g2}} = \frac{N_1 N_2}{R_m} = N_1 N_2 P_m \] (3.12)

From (3.11) and (3.12) the relationship between the mutual inductance and magnetizing inductances of the primary and secondary can be obtained as:

\[ L_{m1} = M \left( \frac{N_1}{N_2} \right) = M \cdot a \] (3.13)

\[ L_{m2} = M \left( \frac{N_2}{N_1} \right) = M \cdot a^{-1} \] (3.14)

where \( a = N_1/N_2 \), is the number of turn’s ratio.

Using the above equations leads to the inductance based electrical model as demonstrated in Figure 3-12(a). It can be seen that the series reluctances of the air gaps show up as parallel inductances and the leakage inductances are in series. These parallel branches are the mutual inductance between the primary and secondary coils. It can be observed that the model of Figure 3-12(a) is the electrical circuit for the transformer seen from the primary side. To obtain the \( \pi \)-equivalent model, an ideal transformer now needs to be added to allow different turn’s ratio. Thus, the final stage of transformation involves moving the secondary leakage inductance to the secondary side. This leads to Figure 3-12(b) which is in consistent with the well-known \( \pi \)-equivalent model of a transformer.
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Figure 3-11: (a) Simplified magnetic circuit and (b) its permeance network.

Figure 3-12: (a) Inductance model and (b) its π-equivalent model.

Up to now, modelling the physical structure is presented for a system with a single pair of primary and secondary cores. In the case of a whole CSS, there is a complete ring of ferrite
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cores used to provide the magnetic coupling structure as shown in Figure 3-13. This makes its magnetic circuit more complicated and it requires some reluctance relationships for modelling as presented next.

Complete rings of ferrite cores around the shaft

Figure 3-13: Complete geometry of magnetic coupling structures of typical through-hole contactless slipring systems: Coaxial (Right side) and Face to Face (Left side).

3.5.1.1 Reluctance calculations (Coaxial design)

Figure 3-14 shows the cross-section view of a coaxial design with the relevant dimensions. It is assumed that the primary and the secondary cores are identical with a circular symmetry as in Figure 3-13.

![Cross-section view of a coaxial design.](image)

Figure 3-14: Cross-section view of a coaxial design.

To calculate the mutual inductance, the air gap reluctances have to be obtained. As the air gap reluctances are axial in this design, their active cross-section area can be calculated for a cylindrical coordination as follows:
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\[ Area = \frac{Volume}{Length} = \iiint r \, dr \, d\phi \, dz \]

(3.15)

Therefore, for \( R_{g1} \) and \( R_{g2} \), the cross-section area for magnetic flux flow is:

\[ A_{g1} = A_{g2} = \int_{h_1}^{h_2} \int_{r_0}^{r_2} \int_{0}^{2\pi} rdr \, d\phi \, dz = \frac{(h_2 - h_1)(2\pi)(\frac{1}{2})(r_3^2 - r_2^2)}{(r_4 - r_2)} \]

(3.16)

\[ = \pi (h_2 - h_1)(r_3 + r_2) \]

However, the effective air gap area should be corrected to take the fringing field into account, thus the length of the air gap is added to the outer sides of the air gaps as:

\[ h'_2 \approx (h_2 + l_g) \approx [h_2 + (r_3 - r_2)] \]

(3.17)

\[ h'_1 \approx (h_1 + l_g) \approx [h_1 + (r_4 - r_1)] \]

(3.18)

The air gap reluctances then are expressed as:

\[ R_{g1} = R_{g2} = \frac{(r_5 - r_2)}{\mu_0 \pi \left(h'_2 - h_1\right)(r_3 + r_2)} \]

(3.19)

In the case of the leakage inductances, practically the major portion of the leakage flux start leaking back when it gets closer to the air gap after passing the first half of the core limb. This means, the second halves of the core limbs ("\( r_2 - r_1 \)" for the secondary and "\( r_4 - r_3 \)" for the primary in Figure 3-14) have to be considered for their cross-section area as follows:

\[ A_{l1} = \pi \left(r_4^2 - r_3^2\right) \]

(3.20)

\[ A_{l2} = \pi \left(r_2^2 - r_1^2\right) \]

(3.21)

And their relevant reluctances are:

\[ R_{l1} = \frac{(2h_1)}{\mu_0 (A_{l1})} \]

(3.22)

\[ R_{l2} = \frac{(2h_1)}{\mu_0 (A_{l2})} \]

(3.23)
3.5.1.2 Reluctance calculations (Face to Face design)

Similarly, for the design shown in Figure 3-15 for the air gap reluctances, first the areas have to be corrected. Thus, for \( R_{g1} \) and \( R_{g2} \):

\[
\begin{align*}
R_{g1}' &\approx R_{g1} + (2h_l) \\
R_{g2}' &\approx R_{g2} - (2h_l)
\end{align*}
\]

(3.24) (3.25)

And their relevant reluctances are:

\[
\begin{align*}
R_{g1} &= \frac{(2h_l)}{\mu_0 \pi (r_4^2 - r_5^2)} \\
R_{g2} &= \frac{(2h_l)}{\mu_0 \pi (r_2^2 - r_1^2)}
\end{align*}
\]

(3.26) (3.27)

Likewise, for the leakage flux lines, the second halves of the core limbs are considered for the flux flow. However, because they are radial in this design, the cross-section area for the flux flow is calculated from (3.15), and accordingly the reluctances can be obtained from the following relationships:

\[
R_{l1} = R_{l2} = \frac{(r_5 - r_2)}{\mu_0 \pi (h_2 - h_1)(r_3 + r_2)}
\]

(3.28)

Figure 3-15: Cross-section view of a face to face design.
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Table 3-1 summarises the derived equations for air gaps as well as leakage reluctances for both coaxial and face to face designs which can be used for inductance calculation before construction.

Table 3-1: Reluctance equations derived for coaxial and face to face designs.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_{g1}$</td>
<td>$\frac{(r_3 - r_2)}{\mu_0 \pi (h_2' - h_1)(r_3 + r_2)}$</td>
<td>$\frac{(2h_1)}{\mu_0 \pi (r_4'^2 - r_3^2)}$</td>
</tr>
<tr>
<td>$R_{g2}$</td>
<td>$\frac{(r_3 - r_2)}{\mu_0 \pi (h_2' - h_1)(r_3 + r_2)}$</td>
<td>$\frac{(2h_1)}{\mu_0 \pi (r_2^2 - r_1'^2)}$</td>
</tr>
<tr>
<td>$R_{lk1}$</td>
<td>$\frac{(2h_1)}{\mu_0 \pi (r_4'^2 - r_3^2)}$</td>
<td>$\frac{(r_3 - r_2)}{\mu_0 \pi (h_2 - h_1)(r_3 + r_2)}$</td>
</tr>
<tr>
<td>$R_{lk2}$</td>
<td>$\frac{(2h_1)}{\mu_0 \pi (r_2^2 - r_1'^2)}$</td>
<td>$\frac{(r_3 - r_2)}{\mu_0 \pi (h_2 - h_1)(r_3 + r_2)}$</td>
</tr>
</tbody>
</table>

3.5.2 Accurate modeling with resistive components

The coils losses in a magnetic structure of a contactless slipring system are classified into DC and AC losses. The coils resistances can be calculated using Cheng’s [71, 103] or Hurley’s [104] methods. In this section only a new method of estimating the core losses in a contactless slipring system is presented using T-equivalent circuit shown in Figure 3-16. This loss is consists of two components: 1) the hysteresis loss and 2) eddy current loss. The hysteresis loss is the energy loss when the magnetic material is going through a cycling state and the eddy current loss is caused when the magnetic flux lines pass through the core, inducing electrical currents in it.

![Figure 3-16: T-equivalent circuit.](image)

The frequency domain relationships of the above circuit can be expressed as:
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\[ V_1 = j\omega L_1 I_1 + j\omega M I_2 \]  \hspace{1cm} (3.29)

\[ V_2 = j\omega M I_1 + j\omega L_2 I_2 \]  \hspace{1cm} (3.30)

For the secondary side open-circuited (I_2=0), the open circuit voltage is:

\[ V_{oc} = j\omega M I_1 \]  \hspace{1cm} (3.31)

Replacing M by \((L_{m1} a^{-1})\) from (3.13) gives the open circuit voltage in terms of primary magnetizing inductance as:

\[ V_{oc} = j\omega (L_{m1} a^{-1}) I_1 \]  \hspace{1cm} (3.32)

The open circuit voltage of (3.32) is in fact the voltage across the magnetizing branch of the equivalent circuit shown in Figure 3-17. From this model, the primary voltage can be expressed by:

\[ V_i = I_1 (R_1 + j\omega L_{a1}) + V_{oc} = I_1 [R_1 + j\omega (L_{a1} + L_{m1} a^{-1})] \]  \hspace{1cm} (3.33)

Since the total primary current in an unloaded magnetic structure is the exciting current and establishes an alternating flux in the magnetic circuit is calculated as:

\[ I_1 = I_o = \frac{V_i \angle 0^\circ}{R_1 + j\omega (L_{a1} + L_{m1} a^{-1})} = I_1 \angle \varphi_{i1} \]  \hspace{1cm} (3.34)

where:

\[ |I_1| = \frac{V_i}{\sqrt{R_1^2 + \omega^2 (L_{a1} + L_{m1} a^{-1})^2}} \]  \hspace{1cm} (3.35)

\[ \varphi_{i1} = -\tan^{-1} \left( \frac{\omega(L_{a1} + L_{m1} a^{-1})}{R_1} \right) \]  \hspace{1cm} (3.36)

As the magnetic flux in the core lags 90° behind the primary voltage, the current drawn by the primary coil from the source to produce this flux is also lags the supply voltage by 90°. Thereby, the exciting current comprises of two components, one is lagging by 90° and the other in phase as depicted in Figure 3-18.
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![Figure 3-17: Equivalent circuit with magnetizing branch.](image)

![Figure 3-18: Phasor diagram of the circuit in Figure 3-17.](image)

The in-phase component supplies the power absorbed by hysteresis and eddy-currents losses in the core. Where, the remainder is the magnetizing current as given below:

\[
I_c = \text{Re}[I] = |I| \cos \varphi_{i1} \quad (3.37)
\]

\[
I_m = \text{Im}[I] = |I| \sin \varphi_{i1} \quad (3.38)
\]

According to the core loss and magnetizing components of the exciting current and their relevant impedances, it can be written that:

\[
\frac{I_m}{I_c} = \frac{R}{\omega L_{m1}} = \frac{|I| \sin \varphi_{i1}}{|I| \cos \varphi_{i1}} \quad (3.39)
\]

From the above relationship, the resistance representing the core losses then is:

\[
R_c = \omega L_{m1} \tan \varphi_{i1} \quad (3.40)
\]

From (3.40), (3.36) and replacing \(L_{l_k1} + L_{m1} = L_1 \quad L_{m1}. \alpha^{-1} = M \) giving the following relationship for \( R_c \):

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\[ R_c = \omega J_{m1} \left( \frac{\omega (L_{m1} + L_{m1} \cdot a^{-1})}{R_1} \right) = \omega M \frac{N_1}{N_2} \left( \frac{\omega L_a}{R_1} \right) \]  

(3.41)

From (3.41) it can be observed that \( R_c \) is directly related to the quality factor \( (\omega L_a/R_1) \) of the primary coil. Hence:

\[ R_c = \omega M \frac{N_1}{N_2} (Q_1) \]  

(3.42)

Equation (3.42) presents a resistance representing the core losses in a CSS irrelevant of current, voltage and phase angles and it is calculated based on the parameter values obtained from the physical dimensions for the given frequency. This is a strong and reliable tool for estimating the resistance representing the core loss with no need for any experimental or simulation studies.

### 3.6 Magnetic coupling structure measurements

The magnetic coupling coefficient is a very important parameter determining the performance of inductive power transfer systems [61, 105, 106]. The output power of an IPT (Inductive Power Transfer) system can be quantified as a function of the magnetic coupling coefficient \( k \) by [107]:

\[ P_{out} = V_1 I_1 k^2 Q_s \]  

(3.43)

where \((V_1I_1)\) is the Volt-Ampere of the primary side, and \(Q_s\) is the quality factor of the secondary circuit.

Currently the magnetic coupling coefficient is measured manually using LCR meters [58, 108-111]. However the method can only be used offline, so it does not work for live IPT systems. The actual magnetic coupling coefficient often varies in practical operation, particularly for loosely coupled IPT systems having large air gap variations [60, 112, 113]. It is important to monitor the real time magnetic coupling condition of an IPT system for its power flow control [114].

In this section, in addition to the existing methods, a new method is proposed in this thesis for determining the magnetic coupling coefficient of IPT systems under real time operating
conditions. The proposed method uses the voltage and current ratios which can easily be obtained from open circuit and short circuit online tests.

### 3.6.1 Self and leakage inductances method

This method is based on the measurements of self and leakage inductances, which can be expressed by:

\[
k = \sqrt{1 - \left(\frac{L_{lk1}}{L_1}\right)} \quad \text{or} \quad k = \sqrt{1 - \left(\frac{L_{lk2}}{L_2}\right)} \tag{3.44}
\]

where \(L_1\) and \(L_2\) are the self-inductances of the primary and the secondary coils measured at each coil when the other coil is open (see Figure 3-19(a)); and \(L_{lk1}\) and \(L_{lk2}\) are the leakage inductances measured at each coil when the other coil is shorted (see Figure 3-19(b)).

The mutual inductance then can be simply found for the known self-inductances \((L_1\) and \(L_2)\) and the coupling coefficient of (3.44) by:

\[
M = k \sqrt{L_1 L_2} \tag{3.45}
\]
3.6.2 Series-aiding and Series-opposing method

The second method is a series-aiding/series-opposing approach, which can be expressed by:

\[
k = \frac{L_{sr+} - L_{sr-}}{4\sqrt{L_1 L_2}} \quad (3.46)
\]

where \( L_{sr+} = L_1 + L_2 + 2M \) is measured when the total inductance of the primary and the secondary coils are connected in series with aiding polarities; and \( L_{sr-} = L_1 + L_2 - 2M \) is measured when the two coils are connected in series with opposing polarities as illustrated in Figure 3-20.

![Figure 3-20: Mutual inductance measurement using interchanged leads.](image)

Using the above relationship, the mutual inductance can be calculated as:

\[
M = \frac{1}{4}(L_{sr+} - L_{sr-}) \quad (3.47)
\]

Clearly, what is not mutually coupled is leakage; therefore the leakage inductance coefficient can be expressed as:

\[
\sigma = 1 - k \quad (3.48)
\]

Likewise, the primary and secondary leakage inductances can be calculated from the following relationships:

\[
L_{1l1} = \sigma \times L_1 \quad (3.49)
\]

\[
L_{1l2} = \sigma \times L_2 \quad (3.50)
\]
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3.6.3 Voltage ratio method

The third method is based on excitation of both the coils, which can be expressed by:

\[ k = \sqrt{\frac{V_{oc2}}{V_1} \cdot \frac{V_{oc1}}{V_2}} \]  

(3.51)

where \( V_{oc2} \) is the open circuit voltage measured at the secondary coil when the primary coil is excited by voltage \( V_1 \); and similarly \( V_{oc1} \) is the open circuit voltage measured at the primary coil when the secondary coil is excited by voltage \( V_2 \) as shown in Figure 3-21.

![Figure 3-21: Magnetic coupling coefficient measurement based on voltage ratio method.](image)

3.6.4 The proposed method

Unlike the existing methods which are based on off-line measurements, the new method determines the magnetic coupling coefficient of the coupled coils by online measurements of the voltage and current ratios, which can be expressed by:

\[ k = \sqrt{\frac{|V_{oc}|}{|V_1|} \cdot \frac{|I_{sc}|}{|I_1|}} \]  

(3.52)

where \( V_{oc} \) is the open circuit voltage of the secondary coil when the voltage of the primary coil is \( V_1 \); and \( I_{sc} \) is the short circuit current flowing through the secondary coil when the current of the primary coil is \( I_1 \).

It should be noted that in IPT systems, the open circuit voltage (\( V_{oc} \)) and the short circuit current (\( I_{sc} \)) are the two fundamental parameters used to determine the performance of the secondary power pickup and power transfer capability analysis [113, 115-117]. The results of these tests can then be advantageously used to determine the magnetic coupling coefficient of the system from (3.52).
3.6.4.1 Theoretical proof

The two coupled coils of an IPT system can be modelled with a T-equivalent circuit as shown in Figure 3-22.

![T-equivalent circuit of two magnetically coupled coils.](image)

From this model, the open circuit voltage (when \( R_L = \infty \)) and short circuit current (when \( R_L = 0 \)) can be expressed as:

\[
V_{oc} = V_1 \frac{j\omega M}{R_1 + j\omega[M + (L_1 - M)]}
\]

\[
I_{sc} = I_1 \frac{j\omega M}{R_2 + j\omega[M + (L_2 - M)]}
\]

From (3.53) and (3.54) a voltage gain \( G_v \) as a ratio of \( V_{oc} \) and \( V_1 \), and a current gain \( G_i \) as a ratio of \( I_{sc} \) and \( I_1 \), can be expressed as:

\[
|G_v| = \left| \frac{V_{oc}}{V_1} \right| = \frac{\omega M}{\sqrt{R_1^2 + (\omega L_1)^2}}
\]

\[
|G_i| = \left| \frac{I_{sc}}{I_1} \right| = \frac{\omega M}{\sqrt{R_2^2 + (\omega L_2)^2}}
\]

These two gains can be further expressed using the quality factors of the primary coil (\( Q_1 = \omega L_1/R_1 \)) and the secondary coil (\( Q_2 = \omega L_2/R_2 \)):

\[
|G_v| = \frac{M/L_1}{\sqrt{\frac{1}{Q_1^2} + 1}}
\]
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\[ |G_r| = \frac{M}{L_2} \sqrt{\frac{1}{Q_2^2} + 1} \]  \hspace{1cm} (3.58)

In a practical IPT system, coil quality factors are normally designed to very high (Q_1 and Q_2 \gg 1). Thus, (3.57) and (3.58) can be simplified as:

\[ |G_r| = \frac{M}{L_1} \quad \text{and} \quad |G_i| = \frac{M}{L_2} \] \hspace{1cm} (3.59)

From (3.59) it can be proven that the magnetic coupling coefficient k can be determined by (3.52), which can also be expressed by the voltage and current gains:

\[ k = \frac{M}{\sqrt{L_1 L_2}} = \sqrt{|G_r| \times |G_i|} = \frac{|V_{oc}|}{|V_i|} \cdot \frac{|I_{sc}|}{|I_i|} \] \hspace{1cm} (3.60)

3.6.4.2 Simulation and practical results

To verify the proposed method, a closely coupled setup similar to a traditional transformer, and a loosely coupled setup with an air gap of 5mm are built. Figure 3-23 shows their 3D finite elements models developed by JMAG package. The practical setups are built with the same geometries wound around the U-shape ferrite cores as the simulated model (see Figure 3-23).

In this research, the simulation and experiments were accomplished at 5 Watts output power from the power pickup as an example for verification. It should be noted that the actual layout of the magnetic structure and the power level of IPT systems can vary depending on specific application and load requirements. Table 3-2 shows the system parameters for simulation study and practical experiments. The magnetic coupling coefficient is obtained from the finite elements simulation for both setups. Then practical testing is performed using the traditional self and leakage inductances offline measurements method. The offline measurement is conducted by disconnecting the coils from the circuit and measuring the inductances of both side coils individually using an LCR meter and then calculating the magnetic coupling coefficient between them from (3.44). Finally each setup is driven by a high frequency power converter, and the magnetic coupling coefficients are determined using the proposed method. The primary side voltage and current are measured directly across the primary coil by using voltage/current probes and an oscilloscope (Model: DSO5054A...
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Agilent). As the operating frequency of circuit is only at kHz levels and the impedance of the measurement probe is very high, the effect of the probe connection to the circuit on the system tuning is negligible. On the secondary side, when the pickup circuit and the load are in operation, they are disconnected for a short time for getting the reading using the two added switches S1 and S2 as shown in Figure 3-24. Because practical power pickups have filters at the output with high time constants, the effect of short time disconnection from the pickup coil on the output voltage would be very small.

There is another measurement method that has been proposed in [118], which requires the values of the inductances as well as the capacitances of the system for the initial measurements, and then the contribution of the capacitance is subtracted from the calculated results to separate the contribution of the inductive components. The proposed method determines the magnetic coupling by using the voltage and current ratios (V_{oc}/V_1 and I_{sc}/I_1) without the inductances values, and it does not need to involve the capacitances of the system, making online measurement of the coupling coefficient possible. The results are listed in Table 3-3 which shows a very good agreement between the simulation, offline and the proposed online method. This signifies that accurate magnetic coupling coefficient can be obtained using the proposed method.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
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<tbody>
<tr>
<td>f (kHz)</td>
<td>50</td>
</tr>
<tr>
<td>N1 = N2</td>
<td>18</td>
</tr>
<tr>
<td>L1 = L2 (µH) (closely coupled setup)</td>
<td>260</td>
</tr>
<tr>
<td>L1 = L2 (µH) (loosely coupled setup)</td>
<td>16.1</td>
</tr>
<tr>
<td>Q1 = Q2 (closely coupled setup)</td>
<td>1614</td>
</tr>
<tr>
<td>Q1 = Q2 (loosely coupled setup)</td>
<td>101</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>5</td>
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<tr>
<td>Bs of the Mn-Zn ferrite material (T)</td>
<td>0.5</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Magnetic coupling structure</th>
<th>JMAG Simulation</th>
<th>Offline method</th>
<th>The proposed online method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Un-gapped structure</td>
<td>0.9513</td>
<td>0.9510</td>
<td>0.9479</td>
</tr>
<tr>
<td>Gapped structure</td>
<td>0.2253</td>
<td>0.2191</td>
<td>0.2141</td>
</tr>
</tbody>
</table>
Chapter 3: Single-phase contactless slipring systems based on pulsating magnetic field

Figure 3-23: 3D finite elements models (a) un-gapped setup and (b) gapped setup.

Figure 3-24: The proposed method for measuring the coupling coefficient $k$

In addition to the above verifications, the proposed method is used for the practical slipring systems in the lab and it is shown that it is a reliable and accurate method for coupling coefficient determination.
Chapter 3: Single-phase contactless slipring systems based on pulsating magnetic field

The proposed method is for determining the coupling coefficient for single phase systems, but it may also be extended to poly-phase. As it will be presented later in Chapter 6 and Chapter 7, in poly-phase contactless slipring systems, each primary phase is magnetically coupled to all of the secondary phases and contributing to the voltages induced across their terminals. In such a complex system, it is not an easy task to determine a system-based magnetic coupling coefficient using the existing methods such as LCR meter. The magnetic coupling efficient of (3.52) is a root of the total measured VA at the output side by the VA of the primary side. To extend this for a case with a single secondary coupled to a primary with three phases (a, b and c) average \( k \) can be determined as follows:

\[
 k_{ave} = \sqrt{\frac{S_u}{S_{in}}} = \sqrt{\frac{V_{oc} I_{sc}}{V_a I_a + V_b I_b + V_c I_c}} \tag{3.61}
\]

where \( V_{oc} \) and \( I_{sc} \) are the open-circuit voltage and short-circuit current measured at the secondary coil when all the primary phases are driven at the same time.

And when the secondary side also has power pickup coils, average \( k \) can be expressed as:

\[
 k_{ave} = \sqrt{\frac{S_{u, total}}{S_{in, total}}} = \sqrt{\frac{V_{oc1} I_{sc1} + V_{oc2} I_{sc2} + V_{oc3} I_{sc3}}{V_a I_a + V_b I_b + V_c I_c}} \tag{3.62}
\]

Note that the open-circuit voltages and short-circuit currents of (3.62) are obtained when 1) all the primary phases are driven simultaneously and 2) all the secondary coils open-circuited and short-circuited at the same time.

### 3.7 A new method for improving the magnetic field coupling

With all the advantages of contactless slipring systems, the industry often demands for more power without increasing the system size. Implementing a large system may not be possible because the physical space available on the shaft for wireless power transfer system is often limited in most rotary applications for robotic joints and wind power pitch control [119, 120]. This requires better magnetic coupling structure design to improve the coupling coefficient to achieve higher power density and efficiency. Researchers have tried to improve the magnetic coupling coefficient for IPT applications by powering multi primary coils coupled to a single secondary coil, or introducing new core structures such as I-shaped cores [107, 118, 121, 122] [18]-[21], but they result in a larger size which may not suit slipring applications with a constrained space around the shaft. Previous sections of this chapter have introduced
improved magnetic coupling structure designs regardless of considering the size constrains. This section is to further improve the magnetic field coupling by varying the ratio of copper and ferrite usage without increasing the system size. The proposed method is based on shaping the magnetic field distribution by reducing the leakage area between the core limbs to improve the overall magnetic coupling between the primary and secondary coils for the same size [123].

3.7.1 Theoretical analysis

The proposed method in this section is based on modifying the reluctance model of the magnetic structure and accordingly the inductances of different regions. In a contactless slipring system, a smaller air gap would help to achieve a better magnetic field coupling as long as mechanical clearance is available to avoid direct contact. When the air gap is large, the magnetic flux tends to leak within the core limbs instead of linking to the other side, which can result in low magnetic field coupling [90, 91, 124]. A detailed modelling of the reluctance circuit is presented in Section 3.5 of this chapter for typical UU coupling structures. In the case of reluctance models of pot-core and EE layouts, centre-leg flux is divided into two equal portions through the outer legs encircling the coils. Because of this symmetry, the model can be divided into two identical UU layouts. As presented in Section 3.5 the reluctance models finally translate to the inductance-based electrical model using the duality principle. The reluctances of the regions between the cores limbs ($R_{lk1}$ and $R_{lk2}$) are of key importance, the fields in these regions translate into leakage inductance. Relative permeability equals 1.0 in these non-magnetic regions and in the copper conductors.

The magnetic coupling coefficient $k$ is calculated based on the inductances from:

$$k = \sqrt{\frac{M}{L_1} \cdot \frac{M}{L_2}} = \sqrt{\frac{M}{(L_{\mu 1} + M)} \cdot \frac{M}{(L_{\mu 2} + M)}}$$  \hspace{1cm} (3.63)

The leakage and mutual inductances of (3.63) can be expressed in terms of their reluctances as:

$$L_{\mu 1} = \frac{N_1^2}{R_{\mu 1}} = \frac{N_1^2 \mu_0 A_{\mu 1}}{I_{\mu 1}}$$  \hspace{1cm} (3.64)

$$L_{\mu 2} = \frac{N_2^2}{R_{\mu 2}} = \frac{N_2^2 \mu_0 A_{\mu 2}}{I_{\mu 2}}$$  \hspace{1cm} (3.65)
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\[ M = \frac{N_1 N_2}{R_m} = \frac{N_1 N_2 \mu_0}{l_g} \mu A_g \]  \hspace{1cm} (3.66)

Substituting (3.64), (3.65) and (3.66) into (3.63) gives \( k \) based on the system geometry:

\[ k = \sqrt{\frac{1}{\left(\frac{R_m}{R_{lk1}} + 1\right)} \cdot \frac{1}{\left(\frac{R_m}{R_{lk2}} + 1\right)}} \]  \hspace{1cm} (3.67)

For a system with identical primary and secondary sides such as the systems presented in this section, (3.67) can be rewritten as:

\[ k = \frac{1}{\left(\frac{R_m}{R_{lk1}} + 1\right)} = \frac{1}{\left(\frac{R_m}{R_{lk2}} + 1\right)} \]  \hspace{1cm} (3.68)

The magnetic coupling coefficient of (3.68) can be improved by reducing the ratios \( R_m/R_{lk1} \) and \( R_m/R_{lk2} \). Increasing the reluctance of the leakage flux path (\( R_{lk1} \) and \( R_{lk2} \)) can be achieved by either increasing the length of the leakage flux path \( l_{lk1} \) and \( l_{lk2} \) (the distance between the core limbs) or reducing the leakage cross-section areas \( A_{lk1} \) and \( A_{lk2} \) (see Figure 3-25). As the focus of the proposed method is improving the magnetic coupling coefficient within the existing structure dimensions limit, the distance between the core limbs (\( l_{lk1} \) and \( l_{lk2} \)) is fixed and cannot be increased. Thus \( R_{lk1} \) and \( R_{lk2} \) are increased by reducing \( A_{lk1} \) and \( A_{lk2} \) due to reducing the length of the cores limbs as shown in Figure 3-26. There is an optimum length of the core limbs which gives the maximum increase in the magnetic field coupling between the primary and the secondary coils as studied in the next section.

![Cross-section area and the length of the leakage flux lines.](image)

Figure 3-25: Cross section area and the length of the leakage flux lines.
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The scenario for the mutual reluctance $R_m$ (mainly the air gap reluctance) however is different. As the length of the cores limbs are reduced, the length of the air gap increases which results in greater air gap reluctances and accordingly reduced mutual inductance. The mutual inductance is then improved by filling the provided space (due to the shortened core limbs) with more number of turns. This would provide two improvements: 1) increasing $N_1$ and $N_2$ will increase $M$ as stated in (3.66), and 2) provide two loops of coils on each side in proximity as shown in Figure 3-26 which enhance the flux linkage between the two sides.

![Figure 3-26: The proposed improved designs.](image)

As a result of the above stated design modifications, the ratios $R_m/R_{lk}$ in (3.67) are reduced and results in an increased $k$. This is verified by simulations and practical experiments as presented in the next sections for UU and EE configurations.

### 3.7.2 3D-FEM study and analysis on the proposed method

#### 3.7.2.1 UU arrangements

Figure 3-27 shows the developed 3D models of contactless sliprings using UU-core arrangement. The simulation data are shown in Table 3-4. The air gap of 5mm between the primary and secondary coils is considered which is reasonable in contactless slipring applications.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f$ (kHz)</td>
<td>50</td>
</tr>
<tr>
<td>$N_1=N_2$ (Design 1 and 2)</td>
<td>4</td>
</tr>
<tr>
<td>$N_1=N_2$ (Design 3)</td>
<td>8</td>
</tr>
<tr>
<td>Ferrite Mn-Zn with $B_s$ (T)</td>
<td>0.5</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>0.5~50</td>
</tr>
</tbody>
</table>

Table 3-4: Simulation data (UU arrangements)
Chapter 3: Single-phase contactless slipring systems based on pulsating magnetic field

The first stage of the study is conducted to find the optimum length of the cores limbs for a certain air gap (here is considered to be 5 mm) between the primary and the secondary coils. This is achieved by setting two extreme points for this analysis, two extreme cases namely a design which uses a core with no limbs (0%) and a design which uses a core with full length of limbs (100%). Then the length of the core limbs is varied between the two reference points and the results are shown in Figure 3-28. It should be noted that as the limbs are reduced, the available provided space is used for current flow by adding more number of turns on both sides. As it can be seen from Figure 3-28 the magnetic coupling coefficient changes non-linearly versus the length of the cores limbs. The maximum magnetic coupling coefficient reached to about 0.653 when the length of the limbs reduced to 50%. As compared to the design with full length of limbs (typical design) with $k$ of 0.58, the magnetic coupling coefficient is improved about 13% without increasing the overall size of the system.

**Figure 3-27**: 3D-FEM developed models of UU (left) typical and (right) improved design.

**Figure 3-28**: Magnetic coupling coefficient vs. length of the cores limbs for UU arrangement.
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After finding the best length of the cores limbs with maximum \( k \), further study is conducted for a full range of air gap from 0.5-50 mm and the results are shown in Figure 3-29. It can be seen from Figure 3-29(a) that for small air gaps (< 3.5 mm) the magnetic coupling coefficient for the typical design is greater than the improved design because most of the flux lines tend to link to the other side. As the air gap increases (> 3.5 mm), the coupling coefficient of the improved design increases and remain greater than the typical design for all the air gap range. Such a trajectory for \( k \) can also be observed from the ratio of the mutual reluctance by the leakage reluctance \( \mathcal{R}_m / \mathcal{R}_{lk} \) as shown in Figure 3-29(a) in blue. The smaller, the ratio \( \mathcal{R}_m / \mathcal{R}_{lk} \), the higher the magnetic coupling coefficient as stated in (3.67). For the improved design, this ratio is reduced by reducing the leakage cross-section area (shortening the cores limbs) and using the available area for current flow by accommodating more number of turns. This would increase the system’s magnetic field coupling as well as mutual inductance greatly as depicted in Figure 3-29(b). The maximum improvement of \( k \) is reached to about 23% at 10 mm air gap. For this air gap, the magnetic coupling coefficient of the typical and the improved design are 0.35 and 0.43 respectively. From Figure 3-30 it can be observed that although the flux lines create larger loops when the length of the cores limbs are reduced to 50% of length, but higher flux linkage is maintained between the two sides. This signifies that the presence of the cores limbs have small impact on the magnetic field coupling between the primary and the secondary coils for a gapped magnetic structure already with a large dominant air gap.
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(b) Mutual inductance

Figure 3-29: Simulation results of UU arrangements.

Figure 3-30: Results of magnetic flux lines for 10mm air gap.

3.7.2.2 E-E configuration

Similar study is conducted for the EE-core configuration. 3D-FEM models of EE arrangements are developed (shown in Figure 3-31) using the simulation data of Table 3-5. Note that in the EE-core configuration, the inner and the outer conductors are connected in series in the opposite directions, so they carry equal currents in different directions.

Table 3-5: Simulation data (EE configuration)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>f (kHz)</td>
<td>50</td>
</tr>
<tr>
<td>(N_1=N_2) (Typical design)</td>
<td>20</td>
</tr>
<tr>
<td>(N_1=N_2) (Proposed design)</td>
<td>28</td>
</tr>
<tr>
<td>Ferrite Mn-Zn with Bs (T)</td>
<td>0.5</td>
</tr>
</tbody>
</table>
Figure 3-31: 3D-FEM developed models of EE (left) typical and (right) improved design.

Figure 3-32 shows the curve of the magnetic coupling coefficient versus the length of the cores limbs for 5 mm of air gap. The maximum magnetic coupling coefficient reached to about 0.48 when the length of the limbs is reduced to 50%. As compared to the design with full length of limbs (typical design) with $k$ of 0.4, the magnetic coupling coefficient is improved about 20%. The graphs of Figure 3-33 show the simulation results for a range of air gap from 0.5 to 50 mm. From Figure 3-33(a), for the air gaps smaller than 3 mm, $k$ for the typical design is greater than $k$ for the improved design, and after 3 mm, $k$ of the improved design is greater for all the air gaps. The maximum of 45% improvement in $k$ is achieved for EE layout for 9 mm of air gap. This is similarly occurred for the mutual inductance between both sides as shown in Figure 3-33(b). As the core limbs are shortened, the cross-section area of the flux leakage reduces and accordingly the reluctances of the flux leakage path reduced. After reducing the leakage inductances due to short cores limbs, the mutual inductance then is improved by increasing the number of turns of both sides using the provided available space. Applying these two modifications together reduces $\mathcal{R}_m/\mathcal{R}_{lk}$ and enhances $k$ as seen earlier in Figure 3-33(a). Figure 3-34 shows how the improved design maintains better flux linkage as compared to the typical design.
Figure 3-32: Magnetic coupling coefficient vs. length of the cores limbs.

Figure 3-33: Simulation results of EE arrangements.
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3.7.3 Practical verification of the proposed method

Basic practical models of EE structure shown in Figure 3-35 are built and tested for verification using the data of Table 3-6. The distance between the core limbs of the used ferrite core is 6mm similar to the simulations. The graph of Figure 3-36(a) shows the measured magnetic coupling coefficient for the three structures. As it can be seen the magnetic coupling \( k_3 \) for D_3 (with short limbs and extra number of turns), is greater than the other two structures for all the air gaps. It can be seen that when the air gap becomes greater than the distance between the core limbs (= 6mm), the magnetic coupling coefficient starts increasing for D_2 (with shorted limbs only) as compared to the typical design (D_1). This improvement of \( k_2 \) is due to the reduced cross-section area of the leakage inductance. However its mutual inductance \( M_2 \) is not improved and remains almost same as \( M_1 \). The mutual inductance then is improved for D_3 after filling the available space with extra turns as shown in Figure 3-36(b).

From the practical results and the simulation study for EE structure, a good agreement can be seen between their graph trajectories. This signifies that the proposed method of improving the magnetic coupling coefficient in contactless sliprings is feasible and can greatly results in an increased power transfer capability within the existing physical system dimensions. As a result greatly increases the power density of the system. A contactless system with a higher magnetic field coupling coefficient will normally lead to a higher power efficiency [105]. This is because if the coupling is higher, a lower current in the primary coil will be needed to provide the same amount of output power to the secondary coil. As a result, Joule losses in the primary coil will be reduced leading to a higher power efficiency [89, 121, 125].
Chapter 3: Single-phase contactless slipring systems based on pulsating magnetic field

Table 3-6: Practical data (EE-core arrangements)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f$ (kHz)</td>
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</tr>
<tr>
<td>$N_1=N_2$ (Design 1 and 2)</td>
<td>10</td>
</tr>
<tr>
<td>$N_1=N_2$ (Design 3)</td>
<td>14</td>
</tr>
<tr>
<td>Ferrite Mn-Zn with $B_s$ (T)</td>
<td>0.5</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>2~16</td>
</tr>
</tbody>
</table>

Figure 3-35: Basic practical models for verification (EE-core arrangements).

(a) Magnetic coupling coefficient
3.8 Summary

In this chapter, improved single-phase magnetic coupling structures have been proposed with high magnetic coupling coefficient for contactless slipring applications. The contactless sliprings are categorized into non-through-hole and through-hole types. Non-through-hole types are designed to be placed at the end of the shaft normally using pot-core geometry, and through-hole types are designed as a ring around the existing shaft. The leakage inductance associated with pot-cores is higher than a through-hole structure because of their middle limb.

A method of obtaining an electrical model from modelling the physical structure of a contactless slipring system based on duality principle has been developed and verified. This is completed by full derivation of mathematical equations for numerical calculations. It has shown that a resistance representing the core loss can be obtained based on the modelling from the physical structure which further increase the accuracy of the obtained model.

A method for determining the magnetic coupling coefficient between two coupled coils for IPT applications under operating conditions has been proposed. It has found that the magnetic coupling coefficient can be obtained by online measurements of the open circuit voltage and short circuit current of an IPT system. The method is proven by T-equivalent circuit analysis, and both finite elements simulation and experimental results have demonstrated that the proposed method can accurately determine the coupling coefficient of both closely and
loosely coupled coils with high quality factors. The method can be used for online monitoring of the coupling condition and real time power flow controller design of IPT systems.

A novel method for improving the magnetic coupling coefficient of contactless slipring systems by varying the ratio of copper and ferrite usage has been presented. It is shown that the reluctance model of a contactless slipring system can be modified to improve the magnetic coupling coefficient within the system physical dimensions. It has been found from the FEM simulation results that the magnetic coupling coefficient has improved about 21.3% for UU and 45% for EE core arrangements. The proposed technique is verified by testing a practical model with about 33% improvement of the magnetic coupling coefficient.
Chapter 4: Improved autonomous resonant converter for single-phase contactless sliprings

4.1 Introduction

One of the effective methods to reduce the size of power converters is to operate at a high switching frequency. However, the switching losses can be a problem if traditional hard switched PWM control is used. Therefore, normally soft switching techniques such as Zero Voltage Switching (ZVS) or Zero Current Switching (ZCS) are preferred [126, 127]. Resonant converters with ZVS or ZCS are widely used for generating high frequency currents for inductive power transfer with reduced switching losses and EMI. Current-fed push-pull resonant converters with ZVS operation in particular, are a good choice for various applications such as fuel cells, Photovoltaic (PV), and wireless charging systems [36, 128-132]. However, with all these advantages, their operation needs accurate zero voltage detection and control, which is challenging at high operating frequencies. In addition, a special circuitry is needed to start up the converter [114, 129, 133]. Based on the oscillatory property of the parallel resonant circuit of the converter, a self-sustained current-fed push-pull converter is proposed in [76] which eliminates the necessity of external controllers required by conventional converters. The converter can start up automatically and operate at ZVS. However, in this topology because both the power and ZVS signals for the gate drives are from the resonant voltage, there is a trade-off between the power loss and gate waveform quality in designing its gate drive circuit. As a result, the operating frequency of the converter is constrained, and it is very difficult to increase the system frequency above 1 MHz. In order to increase the switching frequency with ZVS operation, this chapter proposes an improved autonomous current-fed push-pull converter which can increase the operating frequency to 10MHz level with full resonance and soft switching operation. The proposed converter is practically built and used to drive the single-phase contactless slipring systems presented in the previous chapter.

4.2 Existing ZVS current-fed push-pull converter

In an IPT-based contactless slipring, the converter and the magnetic coupling structure are considered to be a set of ‘converter-transformer’ in order to achieve soft switching for
minimizing switching and conduction losses [2]. One of the commonly used converter topology for contactless sliprings is current-fed push-pull parallel resonant converter as shown in Figure 4-1[54]. In this topology, the reactive current remains inside the resonant tank without flowing through the switching devices. Thus the current rating of the switching devices can be smaller and conduction loss can significantly be reduced for a given power level [43]. The sinusoidal waveforms of such a topology have low harmonic content which result in reduced EMI. Moreover, as compared to full-bridge topology, the current-fed push-pull converter has simple gate drive design without high/low side isolation problems, and it also doubles the output resultant resonant voltage [54].

The front-end DC inductors in a current-fed push-pull topology generate a quasi-current source at the input of the converter. This constant current with sine-wave property makes it to fully utilize the magnetic component of the contactless slipring system. The equivalent circuit model of the two front-end inductors can be different depending on their mutual magnetic coupling. It has been shown in [54, 76] that if $L_1 = L_2 = L$ are partially coupled (coupling coefficient $k < 1$), it is equivalent to a fully coupled case with a common external leakage inductance as shown in Figure 4-2. This equivalent leakage inductance $L_{lk}$ is expressed as:

$$L_{lk} = \frac{(L-M)}{2} = \frac{(1-k)L}{2}$$ (4.1)

where $M$ is the mutual inductance between $L_1$ and $L_2$ and $k$ is the magnetic coupling coefficient between $L_1$ and $L_2$. 

![Figure 4-1: General circuit of a typical current-fed push-pull converter.](image)
Chapter 4: Improved autonomous resonant converter for single-phase contactless sliprings

As the leakage inductance of (4.1) appears in series with the $L_{DC}$, it can be used advantageously as a part or the complete DC inductance in the current-fed push-pull converter. Therefore, the DC inductor may be reduced or even eliminated if the leakage is high enough. This finding not only makes the design of a phase splitting transformer much easier, but also can significantly reduce the system cost.

![Figure 4-2: Front-end splitting transformer circuit.](image)

With all the above advantages of current-fed push-pull topology, its operation (required gate drive signals) needs accurate zero crossing detection or other advanced frequency following techniques such as a VCO (Voltage Controlled Oscillator) or a PLL (Phase-Locked Loop). Moreover, additional special circuitry required to start up the converter which makes it a challenging task to implement zero voltage detection and control at high operating frequencies [114, 129, 133]. Two common ways of achieving ZVS for a parallel push-pull resonant converter have been proposed in [54] and as presented next.

### 4.2.1 Traditional ZVS approach

A traditional way of achieving ZVS is to detect zero voltage crossing points of the oscillating voltage of the resonant tank and relatively turn ‘on’ and ‘off” the switches at zero voltage instants. Figure 4-3 demonstrates a general block diagram of a traditional ZVS method. Due to high voltage across the resonant tank, normally a stepping down stage is required to pull down the voltage of the resonant tank to a suitable range for the control circuitry. A step-down transformer can be used for this purpose. However, the transformer is bulky and not preferred. Thus it is more suitable to use a set of voltage divider resistors to pull the voltage down to any desired range. This voltage then is fed back to a frequency follower or a zero voltage detector for zero point detection. After detecting the zero points, gate signals are generated accordingly at the output of the zero voltage detection block. The current of these generated gates signals is usually low and may not be sufficient to charge up the input...
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capacitors of the switches; thus the output of this block is fed to a gate drive circuit for switching according to the detected zero points.

![Diagram of zero voltage detection method]

**Figure 4-3: General diagram of zero voltage detection method.**

A main drawback of this method is the propagation delay due to each interface of the control loop. In addition to the turn ‘on’ delays of the switches, extra delays as a result of processing and feedback of the control loop result in hard switching operation and power losses. The components incorporated in the ZVS control loop may be chosen carefully with minimum propagation delays. However, this is very challenging for high frequency operation as a small delay in control circuitry becomes significant relative to the system operating frequency.

### 4.2.2 Autonomous ZVS approach

Based on the oscillatory property of the parallel resonant circuit of the converter, a self-sustained current-fed push-pull converter as shown in Figure 4-4 is proposed in [76] which eliminates the necessity of external controllers required by conventional converters. Moreover, the converter can start up automatically and operate with ZVS which greatly simplifies the configuration. In this topology, $R_1$ and $R_2$ are current limiting resistors and designed according to the current rating of the Zener diodes and the resonant voltage [134]. Two capacitors in parallel with $R_A$ and $R_B$ usually will be added to speed up the switching operation. The time constant of the provided RC circuit has to be designed relative to the system operating frequency (less than half the period) as given by:

$$\tau = RC < \frac{T}{2} = \frac{1}{2f}$$

(4.2)
Chapter 4: Improved autonomous resonant converter for single-phase contactless sliprings

Although no additional controller is needed in this topology, because both the power and ZVS signals for the gate drives are from the resonant voltage, there is a trade-off between the power loss and gate waveform quality in designing the limiting resistors \( R_1 \) and \( R_2 \). As a result, the operating frequency of the converter is constrained, and it is very difficult to increase the system frequency above 1MHz. Even though operating at high frequency puts strain on switches, it is beneficial for two reasons. Firstly, high frequency operation helps to increase the power transfer capability as open-circuit voltage induced in the secondary coil \( V_{oc} \) is governed by [65, 135]:

\[
V_{oc} = j\omega MI_1
\]  

(4.3)

where, \( M \) is the mutual inductance between the primary and the secondary inductances, \( \omega \) is the primary operating frequency and \( I_1 \) is the current flowing through the primary inductor. This is even more advantageous as \( M \) is very small in some of the loosely coupled applications; a high \( \omega \) would be essential to provide sufficient power transfer [1, 105, 136]. Secondly is that high frequency operation helps to sustain the primary oscillation. As presented in [43, 54, 137], for a current-fed push-pull topology there is a minimum condition for the quality factor of the resonant tank. The minimum Q to ensure start-up and steady-state ZVS operation of such series-load converters has been found to be 2.54 and 1.86 respectively. If the quality factor drops below the minimum requirement, the converter will not be able to sustain the oscillation and consequently system failure. In fact a system with high Q is more stable.
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4.3 Improved autonomous current-fed push-pull resonant converter

4.3.1 Operating principle

Figure 4-5 shows the improved autonomous current-fed resonant converter. It comprises a DC power supply in series with two large DC inductors as compared to the primary inductance acting as a quasi-current source. The converter uses two main switches with a common ground and a parallel-tuned resonant circuit which consists of a resonant inductor $L$, a resonant capacitor $C$, and an equivalent reflected load resistance $R_{eq}$.

![Diagram of improved autonomous current-fed push-pull converter](image)

Figure 4-5: Improved autonomous current-fed push-pull converter with full ZVS operation.

The two input inductors ($L_1$ and $L_2$) divide the DC current in half under the steady state conditions, so that the current flowing into the resonant tank is approximately a square waveform with half the magnitude of the DC current. The drain-source current of each switch then can be expressed by:

\[
 i_{d1}(t) = \begin{cases} 
 i_{dc} & 0 \leq \omega t < \pi \\
 0 & \pi \leq \omega t < 2\pi 
\end{cases} \quad (4.4)
\]

\[
 i_{d2}(t) = \begin{cases} 
 0 & 0 \leq \omega t < \pi \\
 i_{dc} & \pi \leq \omega t < 2\pi 
\end{cases} \quad (4.5)
\]

The total square waveform current flowing through the resonant tank $i_T$ is the difference of the two out-of-phase drain-source currents and can be expressed by:
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\[
   i_r = i_{d1} - i_{d2} = \begin{cases} 
   i_{dc} & 0 \leq \omega t < \pi \\
   -i_{dc} & \pi \leq \omega t < 2\pi 
   \end{cases} 
\]

(4.6)

The circuit of Figure 4-6 shows the basic principle of ZVS operation of the improved autonomous converter. Note that the gate drive circuit is identical for both switches, thus only one side is shown and analysed here. As it can be seen the power for the gates is taken directly from the DC source in series with two current limiting resistors \(R_1\) and \(R_2\). And the signals are taken from the resonant tank via two cross-connected diodes. Since the turning ‘on’ of the switches is driven by the DC source (the main or a separate source), the input capacitors of the switches can be charged up quickly; and the turning ‘off’ is achieved by shorting the main switches via \(D_1\) and \(D_2\), the discharging process is fast. As a result, the quality of the switching signal can be maintained at a higher frequency without high power losses associated. This is the main advantage of the proposed converter compared to the existing topology of Figure 4-4, where the turning ‘on’ and ‘off’ are achieved only by the resonant voltage. At steady state, when the voltage at one side of the tank, say \(v_B\) is high, the voltage at terminal-K of \(D_1\) is higher than the voltage at its terminal-A which is equal to the voltage at the gate. During this time, the diode is reverse biased and therefore, keeping the voltage at the gate of \(S_1\) high \((V_{g1})\). For the second half-cycle when \(v_B = 0\); the voltage at terminal-K of \(D_1\) also goes to zero and consequently the voltage at the gate of \(S_1\). Similar scenario occurs for the other side of the circuit with 180° phase shift.

![Figure 4-6: Basic gate drive circuit.](image)

It can be observed from the above principle that depending on the voltage level on both sides of the resonant tank (at terminal-K of \(D_1\) and \(D_2\)), the voltage at the gates goes high and low following the frequency of the resonant tank as depicted in Figure 4-7. However, for the low state of the gate signals, the voltages do not exactly go to zero because of the voltage drop

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across the diodes. Typically the value of this forward bias voltage \((V_f)\) is very low and about 0.2 \(\sim\) 0.7 Volts depending on the type of the diode. The two diodes \(D_1\) and \(D_2\) play an important role in achieving the ZVS operation as summarised in Table 4-1.

![Figure 4-7: Autonomous gate signals operation following the resonant tank.](image)

Table 4-1: Logic of the ZVS circuit

<table>
<thead>
<tr>
<th>Parameter</th>
<th>First half-cycle</th>
<th>Second half-cycle</th>
</tr>
</thead>
<tbody>
<tr>
<td>(V_B)</td>
<td>High</td>
<td>Low</td>
</tr>
<tr>
<td>(V_A)</td>
<td>Low</td>
<td>High</td>
</tr>
<tr>
<td>(D_1)</td>
<td>OFF</td>
<td>ON</td>
</tr>
<tr>
<td>(D_2)</td>
<td>ON</td>
<td>OFF</td>
</tr>
<tr>
<td>(V_{g1})</td>
<td>High</td>
<td>Low</td>
</tr>
<tr>
<td>(V_{g2})</td>
<td>Low</td>
<td>High</td>
</tr>
</tbody>
</table>

For the proposed converter, the current limiting resistors are needed to prevent the shorting of the DC source regardless of the operating frequency. These resistors in combination with the input capacitances of the switches provide an RC circuit which largely determines the turning ‘on’ speed. For a given switch with a certain input capacitance \(C_{iss}\) a smaller resistance results in a smaller time constant and faster charging speed of the input capacitor. But smaller resistances increase the circuit losses, particularly at a high \(V_{dc}\). This would results in total power losses due to both resistors for one cycle about \(V_{dc}^2/R\). This is mitigated by adding two speedup capacitors \((C_{s1} \text{ and } C_{s2})\) in parallel with \(D_1\) and \(D_2\) as shown in Figure 4-5. Due to these capacitors, some charging current will also be supplied by the resonant voltage to make the turning ‘on’ faster, so the resistances \(R_1\) and \(R_2\) can be designed to be higher to reduce the circuit losses.
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4.3.2 Start-up analysis

The switching operation presented in the previous section is based on the steady state conditions when the converter passed the transient stage of the start-up time. In fact at steady state, the zero crossing points exist and detected by D₁ and D₂ allowing the switching operation to be employed based on the frequency of the resonant tank. At start-up however, this is not the case as initially there is no energy in the tank circuit and no zero point to be detected. Therefore, an initial energy is essential for starting up the circuit because without any initial energy in the circuit, the resonant voltage would have no zero crossings points [138], so automatic turning ‘on’ and ‘off’ of the switches would not occur. As presented in [138] the start-up problem can be solved, if energy is injected into the resonant tank prior to turn-on by 1) providing an initial voltage for the resonant capacitor, 2) providing an initial current for the resonant inductor or 3) giving an initial current to the dc inductor. However, all three options need either extra charging circuitry or extra control consideration in order to provide the required initial energy.

One outstanding feature of the autonomous converter of Figure 4-5 is that an initial DC current is established automatically during the startup transient process without using external gate control. Initially the two switches S₁ and S₂ are ‘off’. After the DC source is turned on, both the switches tend to turn ‘on’ by V_{dc} and the gate resistors R₁ and R₂, so some initial current on the equivalent DC inductance from L₁ and L₂ would be established. However S₁ and S₂ can not stay ‘on’ all the time because the ‘on’ state of one switch will short-circuit the gate of the other one and try to turn the other side ‘off’. In a practical circuit, the two switches would not act at exactly the same speed due to parameter differences and disturbances; the side that turns on faster will win the competition to short circuit the gate voltage of the other side, and further strengthen its ‘on’ state. For instance, a lower voltage, say V_{A}, due to faster turning ‘on’ of S₁, will provide a lower voltage at the gate of S₂, thus S₂ will turn off resulting in a higher voltage drop V_{B}, which will further increase the voltage at terminal-K of D₁ assuring that S₁ remains “on” until the resonant voltage changes the polarity. Consequently, this positive feedback (negative resistance) leads to the bi-stable circuit oscillation and full ZVS operation as shown in Figure 4-8. In fact the cross-coupled differential MOSFET pair presents a negative resistance to the resonator due to positive feedback. This negative resistance compensates for the equivalent resistance of the resonator and enables sustained oscillation [139-141].
The mathematical model governing the initial DC current building up when both switches are ‘on’ can be approximately expressed as:

\[
L_{dc} \frac{di_{dc}(t)}{dt} = V_{dc}
\]  

(4.7)

where \( L_{dc} \) is the equivalent DC inductance from \( L = L_1 = L_2 \) (for un-coupled inductors) given by:

\[
L_{dc} = (L_1 \parallel L_2) = \frac{L}{2}
\]  

(4.8)

So for a short time (start-up time \( t_0 \)) \( i_{dc} \) increases linearly and it can be expressed by:

\[
i_{dc}(t) = \frac{V_{dc}}{L_{dc}} t_0
\]  

(4.9)

4.3.3 Steady state analysis

4.3.3.1 Gate drive circuit analysis

The gate drive circuit dynamics during half a period to turn on the switch can be modeled with two voltage sources as shown in Figure 4-9. From this model, the voltage at the gate \( v_g \) is contributed by both the sources \( V_{dc} \) and \( v_{ds} \), which can be found by applying the superposition theorem.
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If only $V_{dc}$ exists ($v_{ds} = 0$), the gate drive circuit is simplified to an RC circuit as shown in Figure 4-10.

From Figure 4-10, the gate voltage $v_{g1}$ contributed by the DC source is expressed by:

$$v_{g1}(t) = V_{dc} \left[ 1 - e^{-\frac{t}{R(C_{iss} + C_s)}} \right]$$  \hspace{1cm} (4.10)

And if only $v_{ds}$ is present, the gate drive circuit is as shown in Figure 4-11.
The governing equations of Figure 4-11 can be expressed by:

\[ i_s = i_R + i_g \]  \hspace{1cm} (4.11)

\[ v_R = v_{g2} \]  \hspace{1cm} (4.12)

\[ v_s = v_{ds} - v_{g2} \]  \hspace{1cm} (4.13)

Equation (4.11) can be rewritten as:

\[ C_s \frac{dv_s}{dt} = C_{iss} \frac{dv_{g2}}{dt} + \frac{v_{g2}}{R} \]  \hspace{1cm} (4.14)

Substituting (4.13) into (4.14) and simplifying, gives:

\[ \frac{dv_{g2}}{dt} + \frac{1}{R(C_s + C_{iss})} v_{g2} = \frac{C_s}{C_s + C_{iss}} \frac{dv_{ds}}{dt} \]  \hspace{1cm} (4.15)

In (4.15), \( v_{ds} \) (the half-wave voltage across the resonant tank) is expressed by:

\[ v_{ds}(t) = \begin{cases} \pi V_{dc} \sin \omega t & 0 \leq t < \pi \\ 0 & \pi \leq t < 2\pi \end{cases} \]  \hspace{1cm} (4.16)

And the derivation of (4.16) is:

\[ \frac{dv_{ds}}{dt} = \omega \pi V_{dc} \cos \omega t \]  \hspace{1cm} (4.17)

From (4.15) and (4.17) a governing equation can be written as:

\[ \frac{dv_{g2}}{dt} + \frac{1}{R(C_s + C_{iss})} v_{g2} = \frac{C_s}{C_s + C_{iss}} \omega \pi V_{dc} \cos \omega t \]  \hspace{1cm} (4.18)

A complete solution (natural + forced) of (4.18) can be expressed as:

\[ v_{g2}(t) = v_{g2,n} + v_{g2,f} \]  \hspace{1cm} (4.19)

where:

\[ v_{g2,n}(t) = Ke^{-t/\tau} \]  \hspace{1cm} (4.20)
where the time constant \( \tau \) is expressed by:

\[
\tau = R(C_s + C_{iss})
\]  
(4.21)

The constant K can be found from initial conditions. As the initial condition applies to \( v_{gs}(t) \) and not \( v_{gs}(t) \), K should be found after \( v_{gs}(t) \) is calculated.

The forced solution of (4.19) can be expressed by:

\[
v_{gs}(t) = A \cos \omega t + B \sin \omega t
\]  
(4.22)

Substituting (4.22) in (4.18), A and B can be found as:

\[
A = \frac{V_m C_s R \omega}{(1 + \tau^2 \omega^2)} = \frac{\pi V_{dc} C_s R \omega}{(1 + \tau^2 \omega^2)}
\]  
(4.23)

\[
B = \tau \omega A
\]  
(4.24)

Combining (4.19), (4.20), (4.22), (4.23) and (4.24), the solution of (4.19) becomes:

\[
v_{gs}(t) = A \cos \omega t + \tau \omega A \sin \omega t + Ke^{-\frac{t}{\tau}}
\]  
(4.25)

Considering the initial condition \( v_{gs}(0) = 0 \), K is found (\( = -A \)). Thus, (4.25) can be rewritten as:

\[
v_{gs}(t) = A(\cos \omega t + \tau \omega \sin \omega t) - Ae^{-\frac{t}{\tau}}
\]  
(4.26)

The total gate voltage then is \( v_g(t) = v_{g1}(t) + v_{g2}(t) \) and expressed by:

\[
v_g(t) = \left[ V_{dc}\left(1 - e^{-\frac{t}{\tau}}\right)\right]
+ \left[\frac{V_{dc}}{(1 + \tau^2 \omega^2)}\left(\cos \omega t + \tau \omega \sin \omega t\right) - \frac{\pi C_s R \omega}{(1 + \tau^2 \omega^2)}\right]e^{-\frac{t}{\tau}}
\]  
(4.27)

Practically the speedup capacitances and gate limiting resistors need to be designed according to the operating speed and the gate characteristics of the switches. In a practical design, the voltage at the gates should not exceed the maximum gate voltage of the switch. If this is not ensured, a Zener diode with a \( V_z < V_{gs} \) may be added in parallel with \( C_{iss} \) for gate protection. Another alternative is to use a separate DC source with \( V_{dc} < V_{gs} \) for the gate drive circuit.
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without any speedup capacitors used. This allows the main DC source to be increased as per the application requirement without damaging the gates of the switches. This is a clear advantage of the proposed inverter compared to the existing inverter shown in Figure 4-4, where the resonant voltage is the only driving source of the gates.

4.3.3.2 Operating ZVS frequency analysis

The switching operation of the improved autonomous converter is determined by the polarity of the resonant voltage across $C$. To ensure the proposed converter is a soft-switched with ZVS operation, one switch is turned on while the other one is switched off. Thus the current flows through the resonant tank and the two switches in two different cycles according to the states of the switches illustrated in Figure 4-12.

From Figure 4-12; at steady state, the resonant voltage $v_c$ can be described as:
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\[ v_c(t) = \sqrt{2} V_c \sin(\omega t) \]  

(4.28)

where \( V_c \) is rms value and \( \omega \) the zero voltage switching frequency, which is approximately determined by \( L \) and \( C \).

Using the simple volt-second rule in this circuit allows the AC output voltage which is \((v_B - v_A)\), to be obtained as the peak voltage across the primary inductor proportional to the input DC voltage. The rms and peak value of the output AC voltage illustrated in Figure 4-13 can be expressed as [76]:

\[ V_c = \frac{\pi V_{dc}}{\sqrt{2}} \quad \text{and} \quad V_c^\wedge = \pi V_{dc} \]  

(4.29)

The magnitude of the current flowing through the primary inductor then can be determined as:

\[ I_L = \frac{\pi V_{dc}}{\sqrt{(\omega L)^2 + R^2}} \approx \frac{\pi V_{dc}}{\omega_0 L} \frac{Q}{\sqrt{Q^2 + 1}} \]  

(4.30)

where \( Q = \omega_0 L / R \) is the quality factor defined, and \( \omega \) is the practical operating frequency which is approximately equal to the un-damped natural frequency \( \omega_0 = 1/\sqrt{LC} \).

![Figure 4-13: AC and DC voltage balance at steady-state conditions.](image)

The accurate ZVS operating frequency at steady state can be determined by analysing a half cycle. In each half cycle, one of the two separate inductors (\( L_1 \) or \( L_2 \)) is in series with the
resonant tank, and on average it carries half of the DC current as illustrated in Figure 4-14. The other inductor is parallel with the DC source, so in principle it does not affect the resonant tank. Since normally the inductances of DC inductors $L_1$ and $L_2$ are much larger than that of the resonant inductor $L$, the switching network injects an approximately constant DC current into the resonant tank in an half cycle. As a result, Figure 4-14 can be simplified as a second order equivalent circuit as shown in Figure 4-15.

$$I = I_{dc}/2$$

Figure 4-14: 3rd order Steady-state equivalent circuit.

$$I = I_{dc}/2$$

Figure 4-15: Simplified 2nd order equivalent circuit.

From Figure 4-14, a state-space equation can be written as:

$$\frac{d}{dt} \begin{bmatrix} i_L \\ v_c \end{bmatrix} = \begin{bmatrix} \frac{R}{L} & -\frac{1}{L} \\ \frac{1}{C} & 0 \end{bmatrix} \begin{bmatrix} i_L \\ v_c \end{bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{C} \end{bmatrix} I$$  \hspace{1cm} (4.31)

Eqn. (4.31) can be rewritten as a second order differential equation in terms of $v_c$:

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\[
LC \frac{d^2v_c}{dt^2} + RC \frac{dv_c}{dt} + v_c = IR
\]  
(4.32)

By taking the initial conditions of (4.33), the full solution of (4.32) can be obtained as in (4.34).

\[
v_c(0) = 0, \quad \frac{dv_c}{dt}(0) = \frac{I + i_L(0)}{C}
\]  
(4.33)

\[
v_c(t) = \frac{IR}{\sin \theta_v} e^{-\sqrt{t/T}} \sin(\omega_f t - \theta_v) + IR
\]  
(4.34)

At the end of each half ZVS switching period, \(v_c = 0\), (4.34) becomes:

\[
e^{-\sqrt{t/T}} \sin(\omega_f t - \theta_v) + \sin \theta_v = 0
\]  
(4.35)

where, \(\omega_f = 2\pi f_f\) is the free ringing angular frequency, \(T = 2L/R\) is the time constant, and \(\theta_v\) is an initial phase angle expressed by [142]:

\[
\theta_v = \tan^{-1}\left( \frac{\omega_f IRC}{i_L(0) + I(1 - RC/\sqrt{T})} \right) = \tan^{-1}\left( \frac{\sqrt{4Q^2 - 1}}{2Q(K_i + 1) - 1} \right)
\]  
(4.36)

where \(K_i = i_L(0)/I\).

Similar to the resonant voltage, a simplified second order differential equation can be written in terms of the resonant current:

\[
LC \frac{d^2i_L}{dt^2} + RC \frac{di_L}{dt} + i_L = \frac{i_{dc}}{2} = I
\]  
(4.37)

A complete solution of (4.37) is:

\[
i_L(t) = \frac{I + i_L(0)}{\sin \theta_i} e^{-\sqrt{t/T}} \sin(\omega_f t + \theta_i) - I
\]  
(4.38)

where:

\[
\theta_i = \tan^{-1}\left( \frac{2\omega_f L(i_L(0))}{R(I - i_L(0))} \right) = \tan^{-1}\left( \frac{(1 + K_i)\sqrt{4Q^2 - 1}}{1 - K_i} \right)
\]  
(4.39)
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At the end of each half cycle, the resonant current changes its polarity, the relationship $i_L(t_z) = -i_L(0)$ must hold. Thus, Eqn. (4.38) becomes:

$$e^{-t_z/T} \sin(\omega_f t_z - \theta_f) + \frac{K_i - 1}{K_i + 1} \sin \theta_i = 0$$

(4.40)

In principle $t_z$ and $K_i$ can be solved from (4.35) and (4.40) with given circuit parameters. However, because (4.35) and (4.40) are nonlinear equations, getting a closed form analytical solution is not practical, so a numerical method needs to be employed. A further analysis can show that $K_i$ is close to the circuit Q is presented in [47]. Starting from this initial value, an algorithm to iterate $K_i$ and $t_z$ can be developed to find their final values. With $t_z$ known, the actual ZVS operating frequency will be determined by:

$$f_{zvs} = \frac{2}{t_z}$$

(4.41)

4.3.4 Simulation study

Simulation study for the improved autonomous converter of Figure 4-5 is carried out using the circuit parameters shown in Table 4-2. The simulation package used is LTspice IV. The same components used in the experiments are selected from the simulator library to make a consistent comparison. Two N-channel MOSFETs (IRFP240) with 200V/20A/0.18Ω are used for the kHz level operation, and (IRF510) with 100V/5.6A/0.54Ω for the MHz level. Note at 10MHz no additional resonant capacitor is used.

Table 4-2: Simulation/Practical data of the improved autonomous converter

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L$ (µH) for 91 kHz</td>
<td>14</td>
</tr>
<tr>
<td>$L$ (µH) for 10 MHz</td>
<td>1</td>
</tr>
<tr>
<td>$C$ (nF) for 91 kHz</td>
<td>220</td>
</tr>
<tr>
<td>$R$ (Ω) for 91 kHz</td>
<td>1~5</td>
</tr>
<tr>
<td>$R$ (Ω) for 10 MHz</td>
<td>1~27</td>
</tr>
<tr>
<td>$L_1 = L_2$ (mH)</td>
<td>1</td>
</tr>
<tr>
<td>$V_{dc}$ (V)</td>
<td>12</td>
</tr>
<tr>
<td>$V_{gs}$ (V)</td>
<td>20</td>
</tr>
</tbody>
</table>

In addition to the simulation, in this section a general design procedure to show how the gate drive circuit parameters are designed is presented as follows:
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1. Determining the operating frequency of the system which can be analysed using Eqn.(4.41) in relation to circuit parameters. In this study, 90.4 kHz and 9.911 MHz are chosen.

2. Depending on the system frequency the gate drive circuit is designed either by including or excluding the speedup capacitors. For \( C_s = 0 \), the gate circuit will be simplified to a first order \( RC_{iss} \) circuit governed by the first term of (4.27). For a given \( C_{iss} \) and \( V_{dc} \), designing \( R_1 \) and \( R_2 \) is a trade-off between the gate signal quality and the power loss. For example if the resistance (\( R_1 \) and \( R_2 \)) are chosen to be 300 \( \Omega \), the gate drive waveform has a flawless rising and falling edge and the switching is occurring precisely with the resonant tank voltage as shown in Figure 4-16(a), but the total power loss on the resistor is 0.48 W. This power loss can be reduced to 0.12 W by increasing \( R \) to 1.2 k\( \Omega \). However, the rising edge lasts longer and the charging time slows down which it can get worse for even higher resistances as shown in Figure 4-16(b).

![Figure 4-16: Gate waveforms simulation results without speedup capacitors at 91 kHz](image)

(b) \( R_1 = R_2 = 1.2 \, k\Omega \) and \( C_{S1} = C_{S2} = 0 \)

3. Adding speedup capacitors can help to speed up the gate signals and reduce the gate losses by allowing for large resistances \( R_1 \) and \( R_2 \). However, if the capacitance is too
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small, it would have little effect; and if it is too large, the gate drive voltage would drop to zero before the zero voltage crossing of the resonant voltage, and the peak gate voltage may become too high as shown in Figure 4-17(a). For the known current limiting resistance and operating frequency from the previous steps, there is a critical $C_s$ corresponding to $v_g = 0$ at $\omega t = \pi$ (turning ‘off’ instant where $t = T/2$), which can be found from (4.27) by considering the contribution of both the DC source and the resonant voltage. For example, with a 1.2 kΩ current limiting resistance, the critical $C_s$ is found to be about 1.5nF. This $C_s$ leads to a gate waveform in phase with the resonant voltage as shown in Figure 4-17(b), which helps to reduce the gate power losses. Note that this $C_s$ results in peak gate voltage at $\omega t = \pi/2$ (where $t = T/4$) which should be $< V_{gs}$. In this example the peak gate voltage is about 15V which is below the maximum gate rating voltage ($\approx 20V$).

![Figure 4-17: Gate waveforms simulation results with speedup capacitors at 91 kHz.](image)

The upper half of Figure 4-18 shows the simulated waveforms of the gate signals during start-up. At first there is no energy stored in the tank (until $t_0 \approx 20 \mu s$) before the first zero point
exists, during this period the two gate signals are high keeping both switched ‘on’ to provide the initial DC current \(i_{dc}(0)\). Then due to the practical imbalance or noise in the circuit, one of the two diodes (e.g. D1) will turn on, causing one gate voltage (\(V_{g1}\)) to be low, so \(S_1\) will turn “off” while \(S_2\) remains “on”. Consequently, the resonant voltage magnitude starts to oscillate with zero voltage switching operation. The lower part of Figure 4-18 shows the input inrush current as well as the current and the voltage of the switches during start-up under the worst case (no load condition). As it can be seen the DC input current is about 12A which decreases to the steady state level of about 1.1A after 160\(\mu\)s; and there is also a resonant voltage surge up to 250V peak in the process, which stabilizes to about 40V. The start-up over current and voltage may be protected by reducing the input voltage, having a voltage arrestor, or inserting a current limiting resistor during the transient period.

![Simulation waveforms during start-up.](image)

In the case of 10 MHz, if no speedup capacitors are used, and the limiting resistances are 25\(\Omega\), the circuit maintains ZVS operation with good waveform quality as shown in Figure 4-19(a), although the power loss caused by the resistors is high (about 5.76W). This loss can be reduced to 0.72W by increasing the resistors to 200\(\Omega\) although the gate waveform will be distorted as can be seen from Figure 4-19(b). Similar to the kHz case, the critical speedup capacitances are calculated to be about 0.4 nF, which results in an improved gate signal as shown in Figure 4-19(c).
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From the above analysis it can be seen that the resonant tank oscillation continued due to zero points detection captured by D1 and D2 and switching of S1 and S2 occurring precisely at the zero point defined by the frequency of the resonant tank. The signal at the gate of the second switch (V_{g2}) following the resonant voltage can be seen (in red). Steady-state waveforms of the voltage across the resonant tank and the current flowing through the primary inductor are good sinusoidal waveforms at 91 kHz and 10 MHz as illustrated in Figure 4-20. It can be seen that there is no distortions exist over the switch voltages during circuit transitions and in the
final resonant voltage waveforms as shown in Figure 4-20. In consequence, the resultant resonant current in the primary inductance contains very low harmonic components and produces minimal EMI radiation.

Figure 4-20: Steady state simulation results of resonant voltage and current.

In order to study the impact of the current limiting resistors as well as the speedup capacitors on the circuit power efficiency, the designed cases are simulated for a full range of loads, and the results are shown in Figure 4-21. At 91 kHz, the maximum efficiencies are about 92% and 86% for $R = 300 \, \Omega$ and $1.2\, k\Omega$ respectively (corresponding to 29W and 31W output power). The efficiency is increased slightly to 93% after adding the speedup capacitors in the case of 1.2 kΩ. It is worth-mentioning that the speedup capacitors also help to increase the start-up capability at heavy loads. This can be seen from Figure 4-21(a) that the loading range reaches about 54W output power for $R = 1.2k\Omega$ and $C_s = 1.5 \, \text{nF}$. Note that the nominal output power of the designed setup is about 630 W at 48 V DC input at 91 kHz. At 10MHz, it has been found that adding speedup capacitors has a greater impact on the power efficiency as illustrated in Figure 4-21(b). The maximum efficiencies are about 39% and 50% for $R = 25$
Chapter 4: Improved autonomous resonant converter for single-phase contactless sliprings

and 200 Ω respectively. After adding speedup capacitors, the efficiency is improved greatly to about 80% for R = 200 Ω and C_s = 0.4 nF. The maximum output power the inverter can start up at 10 MHz is about 4W and 2.6W for R = 25 and 200 Ω without speedup capacitors, and with the speedup capacitors it has increased to 7.2W. Figure 4-22 demonstrates the simulation results of the THD of the resonant current for a full range of series resistive load variation. The corresponding operating resonant frequency of the system is also added to the same figure to show their relationship. As it can be seen even at the worst case maximum loading condition (R = 4Ω), the THD is below 4%.

Figure 4-21: Power efficiency vs. load resistance.
4.3.5 Experimental results

The proposed improved autonomous converter shown in Figure 4-5 is practically built and has been used to drive the single-phase contactless slipring systems presented in Chapter 3. In this section the converter is validated using the same switches considered for the simulation and parameters shown in Table 4-2. The first experiment is carried out at about 91.7 kHz, which is close to the simulated condition at 91 kHz. The second experiment is conducted for a higher operating frequency at about 10MHz. The primary resonant capacitor is eliminated at this level of high frequency, and the output capacitance of the switches is used to resonate with the primary inductance although it is possible to add an additional capacitor. Figure 4-23 and Figure 4-24 show the resonant voltage $v_A$ (a half wave) and the gate drive signal $V_{G2}$ taken from the practical experiments. It can be seen that similar to the simulation results, the gate drive signal of one of the switches $V_{G2}$ (in green) occurs precisely at the zero voltage as the resonant tank $V_A$ (in yellow). Because the switch turn ‘on’ delay (≈ 8 ns) and turn ‘off’ delay (≈ 12 ns) are significant at 10 MHz, the gate waveforms are distorted. But the final steady state resonant current and voltage waveforms at both 91.7 kHz and 10.05 MHz are good sinusoidal as shown in Figure 4-25 and perfectly accepted for high frequency IPT systems [92, 97, 132, 143].

The practical circuits (with speedup capacitors) are tested for their maximum load at 4Ω (for 91.7 kHz) and 27Ω (for 10MHz) in series with the resonant inductor. The efficiencies measured from the practical circuit are about 85% and 74% corresponding to 38W and 6.6W.
Chapter 4: Improved autonomous resonant converter for single-phase contactless sliprings

of output power at 91.7 kHz and 10.05 MHz respectively. These values are in good agreements with the simulated results shown in Figure 4-21 at 4Ω and 27Ω with efficiencies of 90% and 80%. Based on the ZVS operating frequency analysis method developed in section 4.3.3.2, the actual ZVS frequencies calculated are 90.4 kHz and 9.911 MHz with the given circuit parameters. These numbers are in a very good agreement with the practically measured frequencies of 91.7 kHz and 10.05 MHz. The slight difference may be from the constant DC current assumption in the theoretical analysis, which is in practice, fluctuates slightly according to the actual values of the DC inductances.

(a) $R_1 = R_2 = 300\ \Omega$ and $C_{s1} = C_{s2} = 0$

(b) $R_1 = R_2 = 1.2\ \text{k}\Omega$ and $C_{s1} = C_{s2} = 1.5\ \text{nF}$

Figure 4-23: Steady state practical results of ZVS gates operation at about 92 kHz.

(a) $R_1 = R_2 = 25\ \Omega$ and $C_{s1} = C_{s2} = 0$

(b) $R_1 = R_2 = 200\ \Omega$ and $C_{s1} = C_{s2} = 0.4\ \text{nF}$

Figure 4-24: Steady state practical results of ZVS gates operation at about 10 MHz.
Chapter 4: Improved autonomous resonant converter for single-phase contactless sliprings

Figure 4-25: Practical steady-state results of resonant voltage and current.

4.4 Summary

This chapter has proposed an improved autonomous current-fed push-pull resonant converter which can increase the operating frequency to 10MHz level with full resonance and soft switching operation. Innovation in the proposed autonomous power supply involves using the input DC voltage as the driving power in combination with the resonant voltage as the ZVS signals via two cross-connected diodes. Since the turning ‘on’ of the switches is driven by the DC source (the main or a separate source), the input capacitors of the switches can be charged up quickly; and the turning ‘off’ is achieved by shorting of the main switches via fast diodes, the discharging process is fast. As a result, the quality of the switching signal can be maintained at a higher frequency without high power losses associated. The basic design concept and design equations of the proposed autonomous topology have been presented. The proposed converter has been investigated using LTSpice simulations. Two practical example circuits have been analysed and tested to verify the validity of the proposed converter at 92 kHz and 10 MHz. Both simulation and practical results have demonstrated that the proposed inverter is able to generate high quality resonant currents from kHz to MHz level for a wide range of applications such as Inductive Power Transfer (IPT), induction heating, and other applications where high frequency currents or voltages are required.
Chapter 5: Simulation and experimental study of the single-phase contactless sliprings

5.1 Introduction

Following the proposal and basic analysis of the single-phase magnetic structures presented in Chapter 3, and the power converter for driving it in chapter 4, this chapter studies the proposed single-phase contactless sliprings by thorough simulation and experiments. In order to investigate the magnetic characteristics of the proposed systems, 3D Finite Elements Models (FEM) are developed and simulation study is carried out for system assessment. A set of practical experiments are conducted to evaluate the system performance. The new designs are compared with the existing single phase contactless sliprings throughout the simulation and practical verification. Pot-core types contactless slipring magnetic coupling structures have been researched on and implemented in the past [2, 89, 144]. In [3, 17] two pot-core type of rotating transformers with two adjacent and coaxial arrangements are presented for airborne radar power supply applications, which have achieved a power efficiency of about 90% with 1 mm air gap, 1 kW of power at 100 kHz. Another 3 kW laboratory model of a pot-core type rotating transformer tested for 10mm air gap with an overall efficiency of 93% has been presented in [19]. Normally the pot-core types of contactless sliprings are designed to sit at the end of the shaft, while other versions of contactless sliprings are through-hole types which are designed around the shaft [57, 63, 92, 97, 145].

5.2 FEM analysis and process

In this thesis the 3D-FEM magnetic simulation undertaken for all models using JMAG Studio 10.0™ package. This software was developed by a Japan Research Institute Limited based on FEM modelling. FEM is based on the idea of dividing a complicated object into simple blocks, meshes or small manageable pieces. This technique is widely applied in design analysis especially to handle complicated computer simulations and experiments. FEM modelling is advantageous as it helps to verify the theoretical calculations without any practical model being built. Moreover it helps to study the magnetic analysis and predict the electrical properties of the magnetic structure including inductance calculation, coupling coefficient as well as determine the efficiency and power transfer capability of the system. In
some cases because of resource constraints such as cost and availability of materials it would take a longer time to build a physical system than to build a FEM model on a computer. Thus FEM analysis can save time in the design process with a reduced cost. The benefits of using 3D-FEM simulation also include its high accuracy and ability to make modification to the physical structure relatively easy.

The procedure is comprised of few basic stages as listed below:

1. Start JMAG Studio 10.0™ and click the [New File] icon to open the [New] dialog box.
2. Select ‘3D Magnetic Field Analysis (frequency response analysis)’ for the analysis type.
3. Drawing the 2D model (creating shape of the magnetic structure to be analysed).
4. Defining material properties and setting boundary conditions for the magnetic analysis.
5. After extruding the 2D model, creating the 3D mesh model to make sure that the model developed can be analysed using the finite element method. Note that at this stage, the size of the elements can be chosen to be small for higher accuracy.
6. Creating the relevant circuit based on the practical model. This includes the current/voltage source value, frequency, etc.
7. This stage is simulating and post processing the results.

5.3 Non-through-hole contactless sliprings using pot-core

In this section a non-through-hole contactless slipring system using typical pot-core geometry is studied. The typical pot-core geometry then is modified based on the presented new method in Chapter 3 (Section 3.7) to achieve higher power transfer capability due to higher magnetic coupling coefficient.

5.3.1 FEM simulation results

A detailed geometry of a typical and an improved pot-core contactless sliprings is shown in Figure 5-1. As a good magnetic coupling is important in contactless sliprings systems, the length of the outer cylinder of a typical pot-core design is reduced to half and the available provided space is used for more number of turns to achieve higher magnetic field coupling. More about this design method is presented in Chapter 3 (Section 3.7). To study the magnetic analysis and predict the electrical properties of the systems, finite element 3D models are
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings developed as shown in Figure 5-2. The magnetic 3D models are developed to calculate the magnetic coupling coefficient, power efficiency as well as determine power transfer capability of both systems. The two systems are modelled and analysed using the data shown in Table 5-1.

![Figure 5-1: Physical geometry.](image)

**Table 5-1: System specifications (pot-core designs)**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Typical</th>
<th>Improved</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_p$ (A)</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>$f$ (kHz)</td>
<td>50</td>
<td>50</td>
</tr>
<tr>
<td>$N_p$</td>
<td>12</td>
<td>15</td>
</tr>
<tr>
<td>$N_s$</td>
<td>12</td>
<td>15</td>
</tr>
<tr>
<td>Ferrite material</td>
<td>Mn-Zn with $B_s = 0.5$ T</td>
<td></td>
</tr>
</tbody>
</table>

(a) Typical
Figure 5-2: 3D-FEM models of pot-core contactless sliprings.

Figure 5-3 shows the graph of magnetic coupling coefficient versus the air gap between the primary and the secondary coils. Due to the design modifications with shorter outer cylinder and adding more turns, higher magnetic coupling coefficient is achieved for the same overall size. It can be seen that with the increase of air gap length after 1 mm, the magnetic coupling coefficient of the improved system is increased for the rest of the air gap length. The maximum improvement is achieved about 20% at 5 mm air gap. For this air gap, the magnetic coupling coefficients are 0.252 and 0.312 for the typical and the improved systems respectively.

For the same range of air gap length, open-circuit and short-circuit tests are applied to calculate the uncompensated power as shown in Figure 5-4. At 5 mm air gap, the power almost doubled for the improved system due to 20% of improvement in the magnetic coupling coefficient. The uncompensated power for the typical system reached to about 1.36 VA (1.81 V and 0.756 A) and it is about 2.62 VA (2.8 V and 0.936 A) for the improved system. Figure 5-5 illustrates the simulated magnetic flux linkage for both magnetic structures. As it can be observed, although for the improved design, the magnetic flux creates larger loops, but they maintain better flux linkage between the two coils. Moreover, as a result of large loops of magnetic flux lines, some of the leakage flux line loops will be able to partially link with the other coil and enhances the magnetic field coupling. However, large loops of magnetic field lines may increase the EMI and can be a problem in some applications as shown in Figure 5-6.
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

Figure 5-3: Magnetic coupling coefficient.

Figure 5-4: Uncompensated power ($S_u$).

Figure 5-5: Magnetic flux linkage: Typical design (a) and Improved design (b).
The two systems have been studied for a full range of load at 5 mm air gap and the results are shown in Figure 5-7(a). As it can be observed the improved design is able to extend the maximum system loading to a higher level at a higher efficiency. The maximum power is reached to about 0.66 and 1.25 W for the typical and improved systems respectively without tuning. The power efficiency curves are shown in Figure 5-7(b) with the maximum of 66% and 77% for the typical and the improved structures respectively.

A series tuning is considered for the secondary inductance to boost the power up for the system. The assigned wire for simulation similar to the practically used wire is AWG14, which can handle maximum current about 5.9 A (at 30 °C). This would limit the Qs to about 6 (≈ 5.9/0.936). For Qs of 6, known frequency (50 kHz), and known secondary coil inductance (\(V_{oc}/\omega I_{sc}\)); the corresponding load would be 0.4 and 0.5 Ω for the typical and improved structures respectively. The results for the compensated conditions are demonstrated in Table 5-2. It can be seen that, the power delivered to the load by the typical pot-core system at Qs of 6 is about 6.8 W and it is doubled for the improved structure to about 13.4 W. A parallel tuning on the other hand is more popular in IPT applications due to its natural current limiting.
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings characteristic, so that the tuning naturally boosts the voltage to the load rather than the current (right side of Table 5-2). Consequently, the power profile will only be able to be met if sufficient current can be delivered to the load [113].

![Output power graph](image1)

(a) Output power

![Efficiency graph](image2)

(b) Power efficiency

Figure 5-7: Output power and efficiency without tuning for 5 mm air gap.

| Table 5-2: Simulation results with series and parallel-tuned secondary at Qs = 6 |
|---------------------------------|-----------------|-----------------|
| Parameter                      | Series-tuned secondary | Parallel-tuned secondary |
|                                | Typical  | Improved | Typical  | Improved |
| $I_{out}$ (A)                  | 4.12    | 5.15     | 0.7      | 0.86     |
| $V_{out}$ (V)                  | 1.65    | 2.6      | 9.92     | 15.51    |
| $P_{out}$ (W)                  | 6.8     | 13.3     | 6.86     | 13.4     |
| Efficiency (%)                 | 87.9    | 90.33    | 87.73    | 90.14    |
5.3.2 Experimental results

The pot-core designs of Figure 5-2 are practically built and tested in the laboratory with the same system data shown in Table 5-1. Figure 5-8 shows both designs layouts before and after windings. To minimize the copper loss, a woven Litz wire is used similar to the assigned wire for simulation study. The first experiment is completed for open circuit and short circuit tests for 5 mm air gap and the results are demonstrated in Table 5-3. Similar to the simulation study, both systems are tested at Qs of 6 for series and parallel-tuned secondary circuit as shown in Table 5-4. A very good agreement can be seen between the simulation and experimental results. This further confirms that a typical pot-core design can simply be modified within its physical dimensions and double the system power transfer capability.

Figure 5-8: Practical model of pot-core designs before and after windings.

Table 5-3: Experimental results for uncompensated power for 5 mm air gap ($S_o$)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Typical</th>
<th>Improved</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k$</td>
<td>0.27</td>
<td>0.32</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>0.8</td>
<td>1</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>1.87</td>
<td>2.8</td>
</tr>
<tr>
<td>$S_o$ (VA)</td>
<td>1.5</td>
<td>2.8</td>
</tr>
</tbody>
</table>

Table 5-4: Experimental results with series and parallel-tuned secondary at Qs = 6

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Series-tuned secondary</th>
<th>Parallel-tuned secondary</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Typical</td>
<td>Improved</td>
</tr>
<tr>
<td>$I_{out}$ (A)</td>
<td>4.72</td>
<td>5.37</td>
</tr>
<tr>
<td>$V_{out}$ (V)</td>
<td>1.8</td>
<td>2.7</td>
</tr>
<tr>
<td>$P_{out}$ (W)</td>
<td>8.5</td>
<td>14.5</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>86</td>
<td>89</td>
</tr>
</tbody>
</table>
5.4 Through-hole contactless sliprings with two air gaps

In this section three through-hole contactless sliprings which they utilise two air gaps in their magnetic structures are studied and compared against each other. One is the existing through-hole system with coaxial layout (Chapter 3-Section 3.3.2) and the other two are the improved through-hole magnetic structures with face to face and double-stator layout (Chapter 3-Sections 3.4.1 and 3.4.2).

5.4.1 Power transfer capability analysis

Figure 5-9 shows the developed 3D-FEM models of the coaxial, the face to face, and the double stator systems. Table 5-5 shows the data for simulation. The parameters of the 3D developed models are assigned similar values to the prototyped models, in order to have a consistent comparison. The air gap of 8.5 mm between the primary and secondary has been considered for all 3 systems. In a practical contactless slipring system design, a smaller air gap would help to achieve a better magnetic field coupling as long as there is no direct contact. However, the required air gap for practical mechanical clearance varies from 1mm to about 10mm in contactless slipring applications.

The first stage of FEM magnetic simulation is carried out without any shaft considered (similar to a non-metallic shaft) and the results are shown in Table 5-6. It is clear that the magnetic coupling is enhanced for the double-stator system compared to the coaxial and face to face systems. The uncompensated power as a product of $V_{oc}$ and $I_{sc}$ is about 157.55 VA for the double-stator system which is about 1.86 times the coaxial, and 2.73 times the face to face systems. This would be a great advantage as in a double-stator system the only extra added material to the face to face design is a secondary coil sandwiched in between the two existing parts. When a metal shaft is used for the system, some of the stray flux lines that pass through the shaft create power loss within the shaft due to induced eddy currents. Table 5-7 demonstrates the uncompensated power when a steel shaft is used for each system. As compared to the results of Table 5-6 (without a metal shaft), about 9.66% of the power is lost in the shaft for the face to face, 4.64% for the coaxial and 17.64% for the double-stator systems. As it can be seen, minimum amount of power loss is occurred for the coaxial system, while the double-stator system has the maximum power loss. This is because unlike the coaxial structure, one of the air gaps in the face to face and double-stator structures is adjacent to the shaft and prone to more fringing flux lines. This can be seen in Figure 5-9.
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

Table 5-5: Simulation data for the developed models

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double-Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>I_p (A)</td>
<td>3</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>f (kHz)</td>
<td>50</td>
<td>50</td>
<td>50</td>
</tr>
<tr>
<td>N_p</td>
<td>10</td>
<td>10</td>
<td>20 (2*10)</td>
</tr>
<tr>
<td>N_s</td>
<td>10</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>Primary cores</td>
<td>71</td>
<td>44</td>
<td>88 (2*44)</td>
</tr>
<tr>
<td>Secondary cores</td>
<td>46</td>
<td>44</td>
<td>0</td>
</tr>
<tr>
<td>Shaft length (mm)</td>
<td>65</td>
<td>80.5</td>
<td>99</td>
</tr>
<tr>
<td>Ferrite material</td>
<td>Mn-Zn with B_s = 0.5 T</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

(a) Coaxial

(b) Face to Face
Table 5-6: Simulation results (No shaft, no shielding)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double-Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>k</td>
<td>0.64</td>
<td>0.61</td>
<td>0.7</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>42.59</td>
<td>31.59</td>
<td>33.24</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>1.98</td>
<td>1.82</td>
<td>4.74</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>84.32</td>
<td>57.52</td>
<td>157.55</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>1.403e-002</td>
<td>1.959e-002</td>
<td>2.027e-002</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>2.473e+005</td>
<td>2.586e+005</td>
<td>2.599e+005</td>
</tr>
</tbody>
</table>

Table 5-7: Simulation results with steel shaft without shielding

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double-Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>k</td>
<td>0.63</td>
<td>0.59</td>
<td>0.66</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>41.11</td>
<td>29.38</td>
<td>27.97</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>1.95</td>
<td>1.76</td>
<td>4.64</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>80.28</td>
<td>51.91</td>
<td>129.8</td>
</tr>
<tr>
<td>Loss in the shaft</td>
<td>4.64%</td>
<td>9.66%</td>
<td>17.64%</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>1.396e-002</td>
<td>1.944e-002</td>
<td>1.966e-002</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>3.252e+005</td>
<td>4.972e+005</td>
<td>7.133e+005</td>
</tr>
</tbody>
</table>

Figure 5-10 shows the current density due to the induced eddy current in the shaft. The maximum current density is reached to about $2.1*10^5$, $3.88*10^5$ and $5.88*10^5$ (A/m$^2$) within the shaft for the coaxial, face to face and double-stator systems. By comparing these values with the maximum current density of each system form Table 5-7; they are considerable.
values and most likely creates substantial heat in the shaft. To overcome this, usually a passive magnetic shielding is placed between the shaft and the magnetic coupling structure [93], [94]. One of the common materials used is aluminium, although it creates power loss and usually a trade-off design. In this thesis, flexible ferrite material (FFSX) which is not associated with power loss is proposed and used for shielding as detailed in Section 5.4.2. The flexible ferrite material proposed for shielding is available in the market and practically used for experiments as presented later in Section 5.4.4 [146].

![Figure 5-10: Current density within the shaft (A/m²): (a) Coaxial, (b) Face to Face and (c) Double-Stator.](image)

### 5.4.2 Impact of fringing magnetic field and its shielding

#### 5.4.2.1 With aluminium shielding

In this section the three structures (coaxial, face to face and double-stator) are simulated with an aluminium sleeve placed around the shaft for magnetic shielding. Since there are two possibilities for the power loss (steel shaft and aluminium material) the simulation is performed for a case when both the shaft and the aluminium sleeve are present in the system. Then the shaft is removed and the simulation is repeated with the same data with aluminium. This will determine the power loss within the aluminium material and that lost in the shaft when both are present in the system. The simulation result is given in Table 5-8. Figure 5-11 shows the simulated current density when aluminium is used for magnetic shielding. The current density of the induced eddy currents in the shaft is greatly reduced to $3.3 \times 10^4$, $1.05 \times 10^4$ and $1.5 \times 10^4$ (A/m²) as compared to Figure 5-10 without any shielding. However the maximum current density is simulated about $1.372 \times 10^6$, $1.323 \times 10^6$ and $1.364 \times 10^6$ (A/m²) for
the coaxial, face to face and double-stator systems respectively and it is within the aluminium shield. As a result, the power decreased even more as compared to a system with a steel shaft only. By comparing the results of Table 5-8 (with aluminium shielding) and the results of Table 5-7 (with a steel shaft only); it can be observed that about 5.39% (10.03-4.64) more power loss occurred for the coaxial system, this is about 13.89% and 19.62% for the face to face and double-stator systems respectively. From Figure 5-11 and Figure 5-12(a) it can be seen that aluminium material avoids the magnetic field from passing through the shaft. However this creates more loss for the system. Therefore, a good compromise has to be taken in consideration, if aluminium is used for magnetic shielding. Another alternative is to place a slit in the aluminium sleeve to avoid the sleeve acting as a shorted electrical loop [18].

Table 5-8: Simulation results with a steel shaft and an aluminum sleeve

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double-Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>k</td>
<td>0.62</td>
<td>0.56</td>
<td>0.62</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>39.37</td>
<td>26.13</td>
<td>22.36</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>1.92</td>
<td>1.68</td>
<td>4.42</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>75.82</td>
<td>43.89</td>
<td>98.8</td>
</tr>
<tr>
<td>% Total Loss</td>
<td>10.04</td>
<td>23.55</td>
<td>37.26</td>
</tr>
<tr>
<td>% AL</td>
<td>10.03</td>
<td>23.55</td>
<td>37.26</td>
</tr>
<tr>
<td>% Shaft</td>
<td>0.01</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>1.388e-002</td>
<td>1.873e-002</td>
<td>1.828e-002</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>1.372e+006</td>
<td>1.323e+006</td>
<td>1.364e+006</td>
</tr>
</tbody>
</table>

(a) Coaxial
5.4.2.2 With ferrite shielding

Here ferrite material is used as a magnetic shielding in the place of aluminium. In addition to keeping the shaft safe from stray magnetic fields, ferrite has the advantage of enhancing the magnetic coupling coefficient too. Though ferrite is usually considered to be fragile and cannot be moulded to any shape; but with the technology developments nowadays, ferrite is usually found in a variety of shapes, softness, flexibility, etc. Therefore, for simulations a sleeve of ferrite is placed as a magnetic shielding similar to the practical case shown in Figure 5-19. Table 5-9 depicts the simulation results for this case. A complete system is simulated with the presence of both shaft and ferrite, and then with shaft removed. It can be seen from Table 5-9 that by using ferrite as a magnetic shield, instead of creating power loss it boosts the power to a higher level by enhancing the magnetic coupling.
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>k</td>
<td>0.66</td>
<td>0.71</td>
<td>0.84</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>47.15</td>
<td>50.1</td>
<td>76.08</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>2.04</td>
<td>2.14</td>
<td>5.36</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>96.32</td>
<td>107.16</td>
<td>407.63</td>
</tr>
<tr>
<td>Power Increase</td>
<td>19.96%</td>
<td>109.97%</td>
<td>214.23%</td>
</tr>
<tr>
<td>Loss in the shaft</td>
<td>1.76%</td>
<td>1.18%</td>
<td>1.47%</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>1.433e-002</td>
<td>2.294e-002</td>
<td>2.800e-002</td>
</tr>
<tr>
<td>Current density (A/m²)</td>
<td>5.870e+005</td>
<td>5.811e+005</td>
<td>1.063e+006</td>
</tr>
</tbody>
</table>

Referring to Table 5-7 and Table 5-9; the power increased from 84.32 VA to 96.32 VA which is about 19.96% for the coaxial system and rose about 109.97% and 214.23% for the face to face and the double-stator systems correspondingly. Nevertheless, the power increase for the coaxial system is very low as compared to other two systems. This can be realized from their structure layout. In fact, the ferrite sleeve on the shaft provides a magnetic path with low reluctance for the flux lines and closes one of the air gaps for the two face to face and double stator designs, whereas, for the coaxial design, it adds extra ferrite material to the existing secondary ferrite as shown in Figure 5-12(b). In order to determine the amount of power lost in the shaft or how immune the shaft is from the heat generated by the fringing and stray magnetic fields; the results of the power lost within the shaft when ferrite is used for magnetic shielding is added in the 7th row of Table 5-9. The power loss within the shaft when ferrite is used as a magnetic shield is still negligible and it is valuable relative to the power increase made by the ferrite. As with the case of aluminium, no power loss occurs in the shaft but all of it is lost in the aluminium material. This has increased the total power loss increases for the system. For the ferrite case, in addition to shielding purpose, the system power transfer capability increases significantly. Figure 5-13 illustrates the current density within the shaft when ferrite material is used for magnetic shielding. The current density of the induced eddy currents in the shaft is reduced to 3.42*10⁵, 3.35*10⁵ and 6.43*10⁵ (A/m²) as compared to Figure 5-10 without any shielding. These values are about half of the maximum current density for each system which is about 5.78*10⁵, 5.81*10⁵ and 1.06*10⁶ (A/m²) for the coaxial, face to face and double-stator systems respectively. As it can be seen using a ferrite shield totally covers the shaft from the fringing magnetic fields due to the air gap adjacent to the shaft. However there are some eddy currents induced at the ends of the shaft and they can be mitigated by extending the ferrite shield.
Figure 5-12: Simulated flux lines for magnetic shielding with (a) aluminum and (b) ferrite materials.
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5.4.3 Effect of rotation

Since in a contactless slipring system, one side is stationary and the other may rotate, it is important to design a system that is not affected by the rotation. This is because the rotation speed is not controlled or determined by the slipring system and it may be slow as in wind turbine applications or fast as in excitation systems of the electrical machines. For the coaxial and face to face systems, when the secondary core rotates, there will be magnetic force interaction between the two cores surfaces. Whereas, for the double-stator system, there is no magnetic force interaction between permeable ferrite surfaces since all of the ferrite material remains fixed. Moreover, the flux linkage remains smooth without any abrupt changes as
there is no change in the reluctance. Such an inherent symmetry with a fixed air gap, results in fixed mutual inductance for the system. This is because with the mutual inductance variations, the open-circuit voltage and short-circuit current of the system vary too. Controlling $V_{oc}$ and $I_{sc}$ is highly related to the mutual inductance between the two sides [70]. In the case of the coaxial and face to face systems, the magnetic coupling between the primary and the secondary side may not remain fixed and vary with the relative position of the secondary. Therefore, a special control needs to be designed for the output power of such a system. And in the case of the double-stator system, as the magnetic coupling is constant between the two sides a simpler control strategy can fulfill the required power conditioning for the load. A series of FEM simulation work is completed for this section, in order to investigate power flow variations to the load. The simulation is performed similar to the practical testing data for a constant primary current of 3A and a constant frequency of 50 kHz. The power efficiency is calculated based on the defined input and output power ($P_{in}$ and $P_{out}$) as shown in Figure 5-14.

![Diagram](image)

**Figure 5-14: The input and output power for efficiency calculation.**

The graphs of Figure 5-15, shows the simulated output power and efficiency for the three systems versus the secondary circuit quality factor $Q_s$. Since the secondary inductance $L_S$ and the operating frequency are known and constant, the quality factor then is defined by the load. Therefore, the resistive load is varied to cover all the possible practically achievable $Q_s$ for all systems. It can be seen from Figure 5-15 (a) that maximum power about 206 W is transferred to the load for the double stator system without tuning. This is about 4.6 times the power transferred by the coaxial system with 45 W, and it is about 3.4 times the power transferred by the face to face system with 60 W of output power. These maximum transferred powers have consistent agreement with theoretical principles as stated about ($V_{oc}.I_{sc}/2$). All three systems maintain almost same power efficiency with slight differences.
The maximum power efficiencies are 98.8%, 99% and 99.3% for the coaxial, face to face and the double-stator systems respectively.

The unregulated power (shown in Figure 5-15 (a)) fed to the load are not constant and depends on the load value, \( V_o \) and \( I_o \) are varying. For the double-stator system, the mutual inductance is constant which makes it easier to control the power fed to the load. If the secondary inductance is series-tuned (fully tuning), the equivalent reactance and its compensation capacitance cancel each other at the resonant frequency, making the output voltage independent of the load and equal to the secondary open circuit voltage. Therefore, for the double-stator system with constant primary current, constant frequency and a series-
tuned secondary; the voltage across the load can be designed to a certain level simply based on the operating frequency and primary current. In order to investigate this feature further, the same loading approach is performed for a fully series-tuned secondary case and the results are shown in Figure 5-15 (b). As it can be seen the output power of the double-stator system is increasing linearly with the load, while in the case of coaxial and face to face systems the output power is increasing non-linearly with the load. This linear and non-linear relationship is concealed in the profile of the output voltage across the load as shown in Figure 5-16. It can be observed that for the double-stator system, the output voltage across the load is almost constant with a reasonable decrease for the case with series tuning. The voltage across a load of 25 Ω (Qs = 0.57) was about 77 Volts and decreases only 7 V (to 70 Volts) across 0.5 Ω for (Qs = 28.73). This slight decrease occurs by the voltage dropped due to the ESR of the secondary coil which is increasing with the current increase. The load voltage is reduced from 54 to 42 V with about 12 V decrease for the face to face system and it is reduced from 43 to 26 V with about 17 V decrease for the coaxial system. For a practically achievable Qs = 9.6 [63], the power across the load is reached to about 3.7 kW at an efficiency of about 96% for the double-stator system with fully series-tuning. This is about 7.8 times the output power of the coaxial system with 470 W at the same Qs. And as compared to the face to face design, it is about 3.9 times the output power delivered by the face to face system with 940 W at the same Qs.

![Figure 5-16: Results of the output voltage (fully series-tuning) for the case with ferrite shielding.](image)

Another concern due to the rotation for the coaxial and face to face structures is extra eddy current power loss in the ferrite materials. For some of the secondary positions relative to the primary, there will be different main flux paths between two sides; and since eddy currents...
are induced in the core perpendicular to the main flux lines; when the ferrite material is not fixed, some of the flux lines complete their path differently as shown in Figure 5-17. As a result, there will be extra eddy current induced on the top surface of the cores to oppose the flux lines that fringe from the top to the other side. This aspect does not affect the proposed system as there is no rotating ferrite used for this design. This results in lower core losses and smoother flux linkage between the primary and the secondary sides.

![Figure 5-17: Extra eddy current loss in typical designs.](image)

### 5.4.4 Experimental results

In order to evaluate the coaxial, face to face and double-stator systems thoroughly, prototypes are constructed (see Figure 5-18) and tested for evaluation. A set of experiments and measurements are performed and a comparison with the simulated results is conducted. The simulation data are the practically tested data shown in Table 5-5. After the practical models are built, their self-inductances, mutual inductances and magnetic coupling coefficient are measured using LCR measuring methods for both cases with and without ferrite shielding as tabulated in Table 5-10. The measured magnetic coupling coefficients (k) are very close to the simulated results shown in Table 5-7 (without shielding) and Table 5-9 (with ferrite shielding). This further confirms the accuracy of the simulation study. The quality factor of the primary and the secondary coils are also measured and added to Table 5-10. It can be seen that the quality factors of the coils are high indicating low coil ESR’s. This further confirms that neglecting the coils resistances from the equivalent circuit in contactless slipring systems does not cause significant errors. Figure 5-19 shows a type of flexible ferrite material (FFSX) which can easily be rolled around the shaft performing the required magnetic shielding [146]. The temperature at the shaft point (behind the ferrite shielding) is tested by placing a metal
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

object in the place of the shaft and monitoring its temperature. The temperature of the metal object observed was constant for an hour of observed testing.

Table 5-10: Measured parameters of practical prototypes

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_p$ ($\mu$H)</td>
<td>67.52</td>
<td>56.98</td>
<td>120.27</td>
</tr>
<tr>
<td>$L_s$ ($\mu$H)</td>
<td>81.7</td>
<td>56</td>
<td>25.43</td>
</tr>
<tr>
<td>Q primary coil</td>
<td>350</td>
<td>403</td>
<td>454</td>
</tr>
<tr>
<td>Q secondary coil</td>
<td>410</td>
<td>396</td>
<td>235</td>
</tr>
<tr>
<td>M ($\mu$H)</td>
<td>46.28</td>
<td>34.56</td>
<td>36.28</td>
</tr>
<tr>
<td>k</td>
<td>0.623</td>
<td>0.606</td>
<td>0.656</td>
</tr>
</tbody>
</table>

Table 5-10: Measured parameters of practical prototypes (continued)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_p$ ($\mu$H)</td>
<td>71.88</td>
<td>78.7</td>
<td>208.02</td>
</tr>
<tr>
<td>$L_s$ ($\mu$H)</td>
<td>86.97</td>
<td>77.34</td>
<td>47.37</td>
</tr>
<tr>
<td>Q primary coil</td>
<td>372</td>
<td>556</td>
<td>538</td>
</tr>
<tr>
<td>Q secondary coil</td>
<td>436</td>
<td>545</td>
<td>335</td>
</tr>
<tr>
<td>M ($\mu$H)</td>
<td>51.4</td>
<td>54.6</td>
<td>80.45</td>
</tr>
<tr>
<td>k</td>
<td>0.65</td>
<td>0.7</td>
<td>0.81</td>
</tr>
</tbody>
</table>

Table 5-10: Measured parameters of practical prototypes (continued)

Figure 5-18: Practical models: coaxial (top), face to face (middle) and double-stator (bottom).
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The three prototypes are driven by a current-fed push-pull converter with ZVS operation at 50 kHz and constant primary current of 3 A using the power converter that presented in Chapter 4. The open-circuit and short-circuit tests are carried out for all three models for the cases without shielding and with ferrite shielding and the results are shown in Table 5-11. It can be seen from the experimental results that the double-stator system is capable of transferring power about 4.65 times the coaxial and about 3.65 times the face to face systems. Further, this uncompensated power can be boosted to Qs times to achieve higher output power levels. Similar to the simulation study, for 50 kHz, 3A constant current flowing through the primary coil and fully series-tuning the secondary inductance; the practical systems are tested for different resistive loads to achieve various Qs as shown in Figure 5-20. Note that this test is completed for the case with ferrite shielding for all three systems. For the double-stator system, the maximum efficiency has reached to about 98.8% for a 15 Ω load corresponding to Qs of about 0.95. The measured minimum efficiency is about 95% with highest Qs (≈ 5.5). This is due to the high current in the series-tuned secondary pickup circuit. The coaxial and face to face systems demonstrated almost similar increase in output power versus the secondary circuit quality factor Qs. The maximum efficiencies measured about 98.5% for both coaxial and face to face systems at Qs = 1. The minimum efficiencies are slightly lower about 97% for the coaxial and face to face systems at Qs about 5. The graphs of Figure 5-20 (b) show a reasonably constant voltage that equals the V\text{oc} of the system across the load, while the load varied from 25Ω and reduced to 2.5Ω to achieve a practical range of Qs for the systems. This signifies that by adjusting the primary parameters of the system, the power fed to the load can be regulated to a desired level. It can be seen from the results of Table 5-9 and Table 5-11 as well as Figure 5-16 and Figure 5-20 that there is a very good agreement between the simulation and the practical results.

Figure 5-19: Ferrite sheet around the shaft for shielding.
Table 5-11: Experimental results

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Coaxial</th>
<th>Face to Face</th>
<th>Double Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Without shielding</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>43.6</td>
<td>32.5</td>
<td>35</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>1.7</td>
<td>1.8</td>
<td>4.3</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>74.12</td>
<td>58.5</td>
<td>150.5</td>
</tr>
<tr>
<td></td>
<td>With ferrite shielding</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>48.5</td>
<td>51.5</td>
<td>76</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>1.75</td>
<td>2.1</td>
<td>5.2</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>84.87</td>
<td>108.15</td>
<td>395.2</td>
</tr>
</tbody>
</table>

Figure 5-20: Practical results without any secondary power regulator (fully series-tuned secondary).
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5.5 Through-hole contactless sliprings with single air gap

Figure 5-21 shows a detailed geometry of the single air gaps designs. As it can be seen they are still through-hole structures, but one of the air gaps is eliminated in these designs. The single air gap designs are studied based on 3D-FEM simulations as shown in Figure 5-22 using the simulation data shown in Table 5-12.

![Diagram of single air gap designs](image)

(a) Sandwiched  (b) Non-sandwiched

Figure 5-21: Single air gap designs (geometry details).

![3D-FEM models of single air gap designs](image)

(a) Sandwiched  (b) Non-sandwiched

Figure 5-22: 3D-FEM models of the single air gap designs.
Table 5-12: Simulation data for the developed single air gap models

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Non-sandwiched</th>
<th>Sandwiched</th>
</tr>
</thead>
<tbody>
<tr>
<td>I_p (A)</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>f (kHz)</td>
<td>50</td>
<td>50</td>
</tr>
<tr>
<td>N_p</td>
<td>5</td>
<td>10</td>
</tr>
<tr>
<td>N_s</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>Number of cores</td>
<td>46</td>
<td>46</td>
</tr>
<tr>
<td>Shaft length (mm)</td>
<td>65</td>
<td>65</td>
</tr>
<tr>
<td>Ferrite material</td>
<td>Mn-Zn with B_s = 0.5 T</td>
<td></td>
</tr>
</tbody>
</table>

Similar to the double air gap systems, the first step of simulation is completed for open circuit and short circuit tests to study the power transfer capability of both structures. Table 5-13 demonstrates the results of uncompensated power when there is no shaft in the system, and Table 5-14 shows the results when a steel shaft (without shielding) is added to the system. As a result of unique geometry with one air gap, the magnetic field coupling between the primary and the secondary coils is enhanced. The magnetic coupling coefficients are about 0.86 and 0.93 for the non-sandwiched and sandwiched designs respectively. The maximum uncompensated power reached about 60 VA for the non-sandwiched and about 263 VA for the sandwiched systems for the case without shaft. When a steel shaft is added to the system, the power of the non-sandwiched design reduced slightly (about 3%) to about 58 VA. And it is reduced (about 2%) to about 258 VA for the sandwiched design. As it can be seen, because the air gap is away from the shaft in these designs, very low power is lost within the shaft. Both structures are studied when the shaft is shielded using an aluminium sleeve as well as ferrite and the results are shown in Table 5-15 and Table 5-16. When aluminium is used for shielding in single air gap designs, in addition to the power loss in the shaft, extra power loss about 3.8% and 2.4% occurred (in the aluminium) for the non-sandwiched and sandwiched systems respectively. This can be seen from Figure 5-23 that the system maximum current density reached to about 1.1*10⁶ and 2.4*10⁶ for the non-sandwiched and sandwiched respectively and it is within the aluminium sleeve. And when the shaft is shielded by flexible ferrite, the power is increased, while the shaft is immune from the stray magnetic field. The uncompensated power increased about 10.3% for the non-sandwiched and about 6% for the sandwiched designs. In Figure 5-24, the induced eddy currents in one end of the shaft, can be simply mitigated by extending the ferrite shield over the shaft.
### Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

#### Table 5-13: Single air gap designs results (without shaft)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Non-sandwiched</th>
<th>Sandwiched</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k$</td>
<td>0.869</td>
<td>0.931</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>20.8</td>
<td>45</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>2.9</td>
<td>5.85</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>60.15</td>
<td>263.23</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>2.09E-02</td>
<td>3.34E-02</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>2.38E+05</td>
<td>2.96E+05</td>
</tr>
</tbody>
</table>

#### Table 5-14: Single air designs results (with a steel shaft without shielding)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Non-sandwiched</th>
<th>Sandwiched</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k$</td>
<td>0.856</td>
<td>0.93</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>20.23</td>
<td>44.09</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>2.88</td>
<td>5.85</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>58.45</td>
<td>258.1</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>2.06E-02</td>
<td>3.30E-02</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>2.80E+05</td>
<td>6.10E+05</td>
</tr>
<tr>
<td>Power loss in the shaft</td>
<td>3%</td>
<td>2%</td>
</tr>
</tbody>
</table>

#### Table 5-15: Single air designs results (with a steel shaft aluminium shielding)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Non-sandwiched</th>
<th>Sandwiched</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k$</td>
<td>0.851</td>
<td>0.93</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>19.51</td>
<td>43.05</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>2.88</td>
<td>5.85</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>56.31</td>
<td>251.85</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>2.02E-02</td>
<td>3.25E-02</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>1.14E+06</td>
<td>2.53E+06</td>
</tr>
<tr>
<td>Power loss in the shaft</td>
<td>3%</td>
<td>2%</td>
</tr>
<tr>
<td>Power loss in the AL</td>
<td>3.8%</td>
<td>2.4%</td>
</tr>
</tbody>
</table>

#### Table 5-16: Single air designs results (with a steel shaft and ferrite shielding)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Non-sandwiched</th>
<th>Sandwiched</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k$</td>
<td>0.856</td>
<td>0.93</td>
</tr>
<tr>
<td>$V_{oc}$ (V)</td>
<td>22.88</td>
<td>47.55</td>
</tr>
<tr>
<td>$I_{sc}$ (A)</td>
<td>2.9</td>
<td>5.86</td>
</tr>
<tr>
<td>$S_u$ (VA)</td>
<td>66.37</td>
<td>278.78</td>
</tr>
<tr>
<td>$B_m$ (T)</td>
<td>2.19E-02</td>
<td>3.30E-02</td>
</tr>
<tr>
<td>Current density (A/m$^2$)</td>
<td>4.6E+05</td>
<td>6.10E+05</td>
</tr>
<tr>
<td>Power increase</td>
<td>10.36%</td>
<td>6%</td>
</tr>
</tbody>
</table>
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

Figure 5-23: Current density when the shaft (left side) covered by aluminium (right side).

(a) Non-sandwiched

(b) Sandwiched

(a) Non-sandwiched
Both systems are studied for a full range of load under uncompensated and compensated secondary inductance. Figure 5-25 (a) shows the simulated output power for the un-tuned secondary and without power flow regulation. The maximum power transferred to the load is about 140 W for the sandwiched system, which is about 4.2 times the power transferred by the non-sandwiched system with 33 W. The power efficiency of the sandwiched design is slightly higher than the non-sandwiched structure as demonstrated in Figure 5-25 (b). The maximum power efficiency of the sandwiched structure reached to about 99.5% and it is about 99% for the non-sandwiched structure.

In the case of compensation conditions, a series capacitor is added to the power pickup circuit to fully tune the secondary inductance. Figure 5-26 depicts the results for both systems when series tuning is considered for the secondary circuit for a full practical range of Qs. As it can be seen from Figure 5-26 (a), the output power is increasing with precise linear relationship with the load, while there is not power flow regulation is used. In fact similar to the double-stator system, in the single air gap designs, there is no change in the reluctance of the magnetic flux path as the ferrite material is fixed. This results in constant mutual inductance between the primary and the secondary coils and fixed electrical properties. The output voltage across the load is accurately constant with negligible difference can be seen from Figure 5-26(b) for both systems.

Figure 5-24: Current density when the shaft (left side) covered by aluminium (right side).
Figure 5-25: Results without compensation and power flow regulation.
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings

5.6 Effect of number of ferrites on the magnetic field coupling

The previous study is based on the maximum number of ferrites that can be used in the magnetic structure. In a practical design, the number of ferrites may be reduced if its effect on the magnetic field coupling is not significant. This section investigates such an effect in through-hole type contactless slipring systems based on 3D-FEM simulation.

Excessive use of ferrites in a contactless slipring system not only increases the cost but also the weight of the system, so it would be ideal if the use of the ferrites can be reduced. In order to set reference points for this analysis, two extreme cases corresponding to a design without
Chapter 5: Simulation and experimental verification of the single-phase contactless sliprings using any ferrites (0%), and full use of ferrites (100%) are considered. The results obtained from the study of these two extreme cases are summarized in Table 5-17, which shows the maximum and minimum coupling coefficient values depending on the usage of ferrite in the system. Then the number cores are varied between the two reference points and the corresponding coupling coefficients are shown in Figure 5-27. As it can be seen from Table 5-17 and Figure 5-27, the double-stator system has the minimum difference between the two cases with Δk of 0.045, while the single air gap non-sandwiched system has the maximum difference in k with Δk of 0.234. This signifies that in the case of systems like the double-stator, the use of the ferrites may be greatly reduced with little effect on its overall magnetic field coupling.

Table 5-17: No ferrite vs. full ferrite results (through-hole systems)

<table>
<thead>
<tr>
<th>Coupling structure</th>
<th>‘k’ without ferrites</th>
<th>‘k’ with max number of ferrites</th>
<th>Δk</th>
</tr>
</thead>
<tbody>
<tr>
<td>Double Stator</td>
<td>0.648</td>
<td>0.693</td>
<td>0.045</td>
</tr>
<tr>
<td>Face to Face</td>
<td>0.517</td>
<td>0.617</td>
<td>0.1</td>
</tr>
<tr>
<td>Coaxial</td>
<td>0.535</td>
<td>0.639</td>
<td>0.104</td>
</tr>
<tr>
<td>Sandwiched</td>
<td>0.743</td>
<td>0.931</td>
<td>0.188</td>
</tr>
<tr>
<td>Non-sandwiched</td>
<td>0.625</td>
<td>0.859</td>
<td>0.234</td>
</tr>
</tbody>
</table>

Figure 5-27: Results of magnetic coupling coefficient in terms of usage of ferrite (%).
5.7 Summary

This chapter has presented system-based study of the existing and improved single-phase contactless slipring systems based on finite elements methods and experimental results. In the case of contactless slipring systems with pot-core geometry for 5mm air gap, the power almost doubled for the improved system due to the 20% enhancement in its magnetic coupling coefficient for the same size. The simulation and experimental results have demonstrated that about 14.5 W and 8.5 W of power delivered to the load for the improved and typical pot-core systems at Qs of 6 respectively. The power efficiency of the improved system has reached about 89% higher than the efficiency of the typical system with 86%.

In the case of the through-hole systems, it has been demonstrated that the magnetic coupling coefficient of the double-stator system has increased to 0.81 (after the secondary coil is sandwiched between the primary coils) compared to 0.65 for a coaxial and face to face contactless sliprings. It has been found that the double-stator system can transfer about 4 times more power than the coaxial and face to face contactless slipring system. The measured uncompensated power transferred to the load for the double-stator system has reached to about 395 VA compared to 108 VA for the coaxial and face to face systems. The single air gap through-hole systems have demonstrated a very good magnetic field coupling due to their unique geometry with one air gap. The magnetic coupling coefficients are achieved about 0.93 and 0.87 for the sandwiched and non-sandwiched structures respectively. Moreover, in the single air gap structures, the presence of the shaft has small effect on the power transfer capability as the air gap is designed away from the shaft. This leads to a system with no magnetic shield required for the shaft. Due to the constant magnetic coupling between the primary and the secondary coils, in the double-stator as well as the single air gap designs, it has been observed that if the secondary circuit is fully series-tuned, the output voltage is approximately constant.
Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

6.1 Introduction

The previous chapters are all about single-phase contactless slipring systems based on pulsating magnetic field coupling. This chapter proposes a completely new contactless slipring system based on axial travelling magnetic field generated by poly-phase power converters.

The chapter begins by describing the system and the related terminologies. The axial moving magnetic field produced by poly-phase power converters is then analysed. The performance and basic principle of operation of the proposed system are presented in details with analytical expressions. Then a poly-phase converter is proposed to drive the system by generating a resultant axial travelling magnetic field. This is followed by FEM simulation and experimental studies for system assessment. Finally, a performance comparison between the travelling magnetic field-based and pulsating magnetic field-based contactless sliprings systems is conducted.

6.2 The proposed system and its operating principle

In contrast to single-phase systems, the effective magnetic field of this system is a resultant magnetic field out of three-phases which electrically are shifted from each other in order to provide field motion. This type of magnetic field is moving with a constant magnitude along the air gap between the primary and the secondary sides. Figure 6-1 depicts a general view of the axial travelling magnetic field-based contactless slipring system. It comprised of a primary and a secondary parts with the inclusion of air gap to give the freedom of rotation between the two parts. Unlike the coil distribution of rotating magnetic field case (Chapter 7) for this system the coils are distributed along the axis of rotation, instead of being around the shaft. The primary has open symmetrical slots to provide high maximum peak of flux. The operation of this system is based on the inductance profile of the magnetic coupling structure which is related to the system geometry and physical dimensions. As it can be seen, there is complete overlap between the primary and the secondary sides for any angle of rotation.
signifying the inductance has a constant value and is maximum. As there is no change in the inductance, zero force is generated with an excitation current in the winding. This is an important feature as a contactless slipring should not exert any force on the shaft.

![Diagram of Poly-phase contactless slipring based on axial travelling magnetic field](image)

The system is a three phase system with a total of six primary and six secondary distributed coils. Such distributed windings make a better use of the core, copper and also improve the mmf waveform. Each phase of the system is comprised of 2 coils which are wound in the opposite direction and connected in series (180° out of phase) as shown in Figure 6-2 (a). These coils are then housed in the slots along the air gap to form phase winding. For instance, \((a \text{ and } a')\) forming phase-A, and similar to that \((b \text{ and } b')\) and \((c \text{ and } c')\) forming the other two
Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

phases B and C. The final coil arrangement is then provides 60 degrees phase shift between the two adjacent coils in order to produce the required travelling magnetic field along the axis of rotation.

Because of the uniform symmetry for any angle of rotation and a fixed air gap, the relative rotation does not affect the performance of the system; thus the electrical properties of the system remain constant. Once again because of the specific geometry, this system has a compact design and employs effective use of space. For instance, the coils are completely cylindrical so there is no overhang as such that the primary winding totally encircles the secondary. This ensures optimum use of the copper and stored energy. Moreover, due to the cylindrical geometry, the flux is concentrated towards the center of the system. As the flux flows radially from the primary to the secondary, the air gap flux density increases. This increase in the air gap flux density towards the center is proportional to the ratio of the primary surface area to the secondary surface area. Since the lengths of both the primary and secondary are the same, this increase in flux density is then proportional to the ratio of the respective circumferences and hence radii. As the primary encircles the secondary, this ratio is always greater than 1; signifying the increase of the flux density from the primary to the secondary. In principle, this linear coil distribution of the proposed system, provides a vector of magnetic field pointed radially in the r-direction (towards the center), and it moves along the z-direction (along the shaft), instead of rotating around the shaft as in the case of rotating magnetic field-based system [116]. This is one of the main features of such a design, in order to cancel any unwanted rotational associated torque with the rotating magnetic field case. This travelling magnetic field will induce voltage in the primary as well as the secondary windings. If the back-end of coil 2 is connected to back-end of coil 5, then, the total emf induced in the winding of phase-A is:

\[ E_A = E_2 + E_5 \]  \hspace{1cm} (6.1)

Similarly, for the phases B and C:

\[ E_B = E_6 + E_3 \]  \hspace{1cm} (6.2)

\[ E_C = E_4 + E_1 \]  \hspace{1cm} (6.3)
Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

Figure 6-2 (b) illustrates the vector diagram of the induced emf for the three-phase primary winding. It can be seen that coil 2 is connected to the back end of coil 5 to make the total emf of phase-A. This is done in a similar fashion for the other phases.

Based upon the fundamental laws governing electromagnetic phenomena at any point in space as a function of time for a system with cylindrical nature, can be expressed by Maxwell's equations as [111, 147-149]:

\[ \text{Curl } \vec{H} = \nabla \times \vec{H} = \vec{j} \]  \hspace{1cm} (6.4)

\[ \nabla \cdot \vec{B} = 0 \]  \hspace{1cm} (6.5)

\[ \nabla \times \vec{E} = -\frac{d\vec{B}}{dt} \]  \hspace{1cm} (6.6)

The development of the field equations using cylindrical coordinates (r, θ, and z) is based on the following simplifying assumptions [148]:

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Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

- The flux density $B$ has only $r$ and $z$ components.
- The current density $J$ has only a $\theta$ component.

The flux density and the magnetic field intensity are related through the relationship:

$$\vec{B} = \mu \vec{H} \quad (6.7)$$

The magnetic vector potential $\vec{A}$ is defined by:

$$\text{Curl } \vec{A} = \nabla \times \vec{A} = \vec{B} \quad (6.8)$$

where $\vec{A}$ is in the same direction as the current density.

Substituting (6.7) and (6.8) into (6.4) yields:

$$\nabla \times \nabla \times \vec{A} = \mu \vec{J} \quad (6.9)$$

With both the magnetic vector potential and the current density everywhere in the $\theta$-direction, the magnetic flux density has two components, $B_r$ and $B_z$. Using the definitions in cylindrical coordinates and with some simplifications, (6.8) becomes[111]:

$$\vec{B} = \nabla \times \vec{A} = -\frac{dA_\theta}{dz} \vec{a}_r + \left( \frac{dA_\theta}{dr} + \frac{A_\theta}{r} \right) \vec{a}_z \quad (6.10)$$

In (6.10), the transverse component of $B$ is cancelled by symmetry. Thus along the rotation axis of the system there is only a $z$-component of magnetic field travelling with a linear velocity $v$ as presented next.

**6.2.1 Synchronous velocity**

In a rotating field-based system such as induction machines, the magnetic field is rotating around the air gap and it is a function of theta ($\theta = \omega t$). Whereas, in the case of axial travelling magnetic field-based system, the magnetic field is travelling along the air gap and it is a function of distance $z$ from the suitably chosen origin as shown in Figure 6-3. The linear velocity of the travelling wave in the $z$-direction then is:

$$\text{Velocity} = \frac{\text{Displacement}}{\text{Time}} = \frac{dz}{dt} \quad (6.11)$$
Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

It can be seen from Figure 6-3 that if the mmf wave travels a complete pitch pole, the displacement \( z \) is equal to \( \tau \), at the same time \( \theta \) completed half a cycle (\( \pi \)) [53]. Note that pitch pole is the distance between two neighboring poles on the axial length of the primary given by:

\[
\tau = \frac{\text{Primary length}}{\text{Number of poles}} = \frac{l_p}{p} \tag{6.12}
\]

Therefore, it can be stated that \( z \) is proportional to the pitch pole \( \tau \) and \( \theta \) is proportional to half a period of the wave (\( = \pi \)). Therefore, it can be written:

\[
\frac{z}{\theta} = \frac{\tau}{\pi} \quad \text{or} \quad z = \frac{\theta}{\pi} \tag{6.13}
\]

As \( \theta = 2\pi ft \), the synchronous velocity of the travelling wave can be expressed by:

\[
v_s = \frac{dz}{dt} = \frac{dz}{d\theta} \times \frac{d\theta}{dt} = \frac{\tau}{\pi} \omega = 2\tau f \quad \frac{m}{s} \tag{6.14}
\]

Note that the synchronous velocity of (6.14) is not depending on the number of poles and only depend on the pitch pole.

\[\text{Figure 6-3: Pitch pole and the chosen origin.}\]

6.2.2 Axial travelling magnetic field principle

When a balanced three-phase supply is connected to the primary coils of the system, a resultant travelling flux density wave is created and travels along the length of the primary (along z-axis) as shown in Figure 6-4.
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Figure 6-4: Resultant travelling flux density wave along the air gap.

The balanced three-phase instantaneous currents flowing through the three-phase primary windings can be expressed as:

\[ i_a = I_m \cos \omega t \]
\[ i_b = I_m \cos (\omega t - 120^\circ) \]
\[ i_c = I_m \cos (\omega t + 120^\circ) \]

(6.15)

When these currents flow through the respective phase windings (phase-A), each produces a sinusoidally distributed mmf wave in space, pulsating along its magnetic axis and having a peak located along the axis. Each mmf wave can be represented by a space vector along the axis of its phase with magnitude proportional to the instantaneous value of the current. The resultant mmf wave is the net effect of the three component mmf waves at any point in the air gap, defined by a distance \( z \). The origin of \( z \) can be chosen to be the axis of phase-A, as shown in Figure 6-3. At any instant of time, all three phases contribute to the air gap mmf along the path defined by \( z \). The mmf along the \( z \)-axis then can be expressed by:

\[ F(z) = F_a(z) + F_b(z) + F_c(z) \]

(6.16)

These contributions along \( z \) can be expressed by the following relationships:

\[ F_a(z) = N_i \cos \left( \frac{z\pi}{\tau} \right) \]
\[ F_b(z) = N_i \cos \left( \frac{z\pi}{\tau} - 120^\circ \right) \]
\[ F_c(z) = N_i \cos \left( \frac{z\pi}{\tau} + 120^\circ \right) \]

(6.17)

where \( N \) is the effective number of turns in each phase.
The resultant mmf at point $z$ then:

$$F(z) = N_i \cos \left( \frac{z \pi}{\tau} \right) + N_i \cos \left( \frac{z \pi}{\tau} - 120^\circ \right) + N_i \cos \left( \frac{z \pi}{\tau} + 120^\circ \right)$$  \hspace{1cm} (6.18)

The currents $i_a$, $i_b$, and $i_c$ are functions of time and are defined by (6.15). Thus:

$$F(z,t) = N_i m \cos \omega t \cos \left( \frac{z \pi}{\tau} \right)$$

$$+ N_i m \cos \left( \omega t - 120^\circ \right) \cos \left( \frac{z \pi}{\tau} - 120^\circ \right)$$

$$+ N_i m \cos \left( \omega t + 120^\circ \right) \cos \left( \frac{z \pi}{\tau} + 120^\circ \right)$$  \hspace{1cm} (6.19)

$$= \frac{3}{2} N_i m \cos \left( \omega t - \frac{z \pi}{\tau} \right)$$

The resultant flux density wave can be expressed by:

$$B(z,t) = \frac{3}{2} B_{max} \cos \left( \omega t - \frac{z \pi}{\tau} \right)$$  \hspace{1cm} (6.20)

![Figure 6-5: Motion of the resultant flux density.](image)

The expression of (6.20) represents the resultant flux density wave in the air gap. This represents an mmf travelling at the constant linear velocity $v_s = 2f/\tau$. At any instant of time, say $(t_1=0)$, the wave is distributed sinusoidally in the air gap with the positive peak along $Z_1=v_s t_1$. At a later instant, say $(t_2=T/6)$, the positive peak of the sinusoidally distributed wave is along $Z_2=v_s t_2$; that is, the wave moved by $\Delta Z=v_s (t_2-t_1)$ along the air gap as shown in Figure 6-5.
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6.3 FEM motion analysis of the travelling magnetic field

In this section, graphical analysis based on FEM magnetic simulation for the travelling magnetic field is presented. This is important as understanding the motion of the generated magnetic field will give a better perspective to design the secondary power pickups to capture maximum power. Figure 6-6 demonstrates the motion of the travelling field along the air gap for different points on z-axis (axis of rotation). Consider the state at t = 0 (Figure 6-6(a)), the moment when the current of phase-B is at its maximum value $I_m$ in the negative direction. The mmf of phase-B then has its maximum value $F_B = F_{\text{max}}$ along its magnetic axis. In fact, the magnetic field at this moment is concentrated between coils b and b'. At this moment, currents $i_A$ and $i_C$ are both $I_m/2$ in the positive direction. The corresponding mmf's of phases A and C are both of magnitude $F_{\text{max}}/2$ along their magnetic axes. The resultant mmf, obtained by adding the individual contributions of the three phases, is a vector of magnitude $F = 3/2F_{\text{max}}$ centered on the axis of phase-B. It represents a sinusoidal space wave with its peak centered on the axis of phase-B with amplitude 3/2 times that of the phase-B contribution alone. Note that the moment $t=0$ is chosen based on the primary coils arrangement which starts with b' to c from left to right in Figure 6-1. At a later time $\omega t = \pi/3$ (after 60 electrical degrees), the currents in phases B and C are negative half maximum, and that in phase-A is a positive maximum. Therefore, the magnetic field at this moment is concentrated between coils a and a'. The resultant has the same amplitude as at t=0, but it has now moved to the right 60 electrical degrees along z-axis as shown in Figure 6-6(b). Similarly, at $\omega t = 2\pi/3$ when the current of phase-C is a negative maximum and the currents of phase-A and B are positive half maximum; the same resultant mmf distribution is again obtained, but it has moved to the right 60 electrical degrees farther and is now aligned with the magnetic axis of phase-C [Figure 6-6(c)]. As time passes, the resultant mmf wave retains its sinusoidal form and amplitude but moves progressively along the air gap, the net result can be seen to be an mmf wave of constant amplitude travelling at a uniform linear velocity. In one full cycle the resultant mmf must be back in the position of [Figure 6-6(a)]. The mmf wave therefore makes one revolution per electrical cycle in a two-pole system as illustrated [Figure 6-6(d, e, f)].

The method based on the proposed axial travelling magnetic field can also be used for other wireless power transfer applications. For example, it can be used for 2D charging pads for portable consumer electronics like mobile phones. The linear motion can provide a smooth magnetic field coupling across the whole area of the charging pad without dead zones, which helps to simplify the pad design as presented in [117].
Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

(a) $t=0$ ($F_B=F_{max}$, $F_A=F_{max}/2$)

(b) $t=\pi/3$ ($F_A=F_{max}$, $F_B=F_{max}/2$)

(c) $t=2\pi/3$ ($F_C=F_{max}$, $F_A=F_B=F_{max}/2$)

(d) $t=\pi$ ($F_B=F_{max}$, $F_A=F_C=F_{max}/2$)

(e) $t=4\pi/3$ ($F_A=F_{max}$, $F_B=F_C=F_{max}/2$)
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6.4 Secondary power pickup structures

The primary magnetic field of the proposed system is a 2-pole travelling magnetic field; the magnetic field intensity H in the air gap at any point Z along the axis of rotation under one pole is same in magnitude as that at (Z+π) under the opposite pole, but the fields are in the opposite direction. This signifies that, the voltage induced in the secondary coils due to one pole is the same in magnitude and opposite of that from the other pole. Based upon this principle of the travelling magnetic field, the secondary layout can be either a single-phase or poly-phase (similar to the primary side) as shown in Figure 6-7. However, the whole winding has to be designed to have the same number of poles as the primary magnetic field side to capture maximum possible power. To form a 2-poles single-phase power pickup, each three of the secondary coils under one pole are connected in series with the same polarity to form one of the poles. Similarly the other 3 coils which fall under the other pole, form the other pole of the pickup as shown in Figure 6-7(a). The poly-phase power pickup layout on the other hand is similar to the primary side layout as shown in Figure 6-7(b). Due to poly-phase feature with complete overlapping with the primary coils, the power captured by a poly-phase pickup would be higher than the power captured by a single-phase power pickup. These two power pickup layouts are compared with each other throughout FEM analysis and experiments later in Section 6.8.
6.5 Poly-phase mutual inductance equivalent circuit and analysis

In this section an equivalent circuit that can be used to study and predict the performance of the proposed poly-phase system is presented. Normally single-phase contactless power transfer systems are molded based on mutual inductance equivalent circuit for several reasons which make it more convenient to analyze the system as detailed in Chapter 2 (Section 2.5.5.2). Although modeling the proposed poly-phase system is more complicated than a single-phase system, it still can be presented based on the mutual inductance model as shown in Figure 6-8. As it can be seen each primary phase is magnetically coupled to all the secondary phases and contributing to the voltages induced across their terminals. This makes it difficult to analyze the impact of the secondary side onto the primary and predicts the power transfer capability of a poly-phase system [150, 151]. In this section, a new mutual factor based on the pole mutual flux is proposed in this thesis which converts the poly-phase system into a single-phase basis and simplify the analysis.

Normally in a single-phase mutual inductance model, open-circuit voltage in terms of the primary current, mutual inductance and a frequency is $V_{oc}=j\omega MI$ [152]. In case of a poly-phase system on the other hand, the open circuit voltage in each secondary phase is a contribution of all the primary currents. For instance, the open-circuit voltage induced in phase-S1 can be expressed by:

$$V_{oc1} = \omega(M_{A1}\overrightarrow{I_A} + M_{B1}\overrightarrow{I_B} + M_{C1}\overrightarrow{I_C}) \quad (6.21)$$

where $M_{A1}$ is the mutual inductance between phase-A and phase-S1, $M_{B1}$ and $M_{C1}$ are the mutual inductances between phase-S1 and phases-B and C of the primary side, $\overrightarrow{I_A}$, $\overrightarrow{I_B}$ and $\overrightarrow{I_C}$ are the primary currents and they are vector quantities.
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Figure 6-8: Poly-phase mutual inductance model.

As presented in section 6.3 for balanced primary currents, at every 60° electrical degrees the current in one of the phases is maximum and it is half magnitude in the other two phases. For instance at $t=\pi/3$, the current of phase-A is maximum $I_m$, and it is half magnitude ($0.5 I_m$) for phase-B and C. Thus the induced open-circuit voltage across the terminals of phase-$S_1$ contributed by all primary currents given in (6.15) is:

$$V_{oc1} = \omega M_{A1} I_m + \left( \frac{1}{2} \right) \left[ \omega M_{B1} I_m + \omega M_{C1} I_m \right]$$

(6.22)

where $I_m$ is the maximum magnitude of the primary currents and it is the same for all phases.

Likewise, in a similar approach the open-circuit voltages induced in the other two secondary phases are:

$$V_{oc2} = \omega I_m \left[ M_{B2} + \left( \frac{1}{2} \right) \left( M_{A2} + M_{C2} \right) \right]$$

(6.23)

$$V_{oc3} = \omega I_m \left[ M_{C3} + \left( \frac{1}{2} \right) \left( M_{A3} + M_{B3} \right) \right]$$

(6.24)

where $M_{B2}$ is the mutual inductance between phase-B of primary and phase-$S_2$ of the secondary, and $M_{C3}$ is the mutual inductance between phase-C of primary and phase-$S_3$ of the secondary.
Following the same procedure for \((R_L = 0)\), the short circuit currents can be expressed as [152]:

\[
I_{sc1} = \frac{I_m}{L_{s1}} [M_{A1} + \left(\frac{1}{2}\right)(M_{B1} + M_{C1})]
\]

\[
I_{sc2} = \frac{I_m}{L_{s2}} [M_{B2} + \left(\frac{1}{2}\right)(M_{A2} + M_{C2})] \quad (6.25)
\]

\[
I_{sc3} = \frac{I_m}{L_{s3}} [M_{C3} + \left(\frac{1}{2}\right)(M_{A3} + M_{B3})]
\]

In the above relationships of open-circuit voltages and short-circuit currents, the total term in the brackets is a net mutual inductance contributed by one pole of the primary magnetic field. Thus a term ‘\(M_P\)’ is introduced as a mutual inductance per pole for each secondary coil coupled to a poly-phase primary side as in (6.26). Using such a resultant mutual inductance simplifies the calculation and analysis of a poly-phase system to a single-phase basis.

\[
M_{P,1} = \left[ M_{A1} + \left(\frac{1}{2}\right)(M_{B1} + M_{C1}) \right]
\]

\[
M_{P,2} = \left[ M_{B2} + \left(\frac{1}{2}\right)(M_{A2} + M_{C2}) \right] \quad (6.26)
\]

\[
M_{P,3} = \left[ M_{C3} + \left(\frac{1}{2}\right)(M_{A3} + M_{B3}) \right]
\]

Thus the open-circuit voltages and short-circuit currents can be rewritten based on the ‘per pole mutual inductances as follows:

\[
V_{oc1} = \omega I_m M_{P,1} \quad V_{oc2} = \omega I_m M_{P,2} \quad V_{oc3} = \omega I_m M_{P,3} \quad (6.27)
\]

\[
I_{sc1} = \frac{I_m}{L_{s1}} \left( M_{P,1} \right) \quad I_{sc2} = \frac{I_m}{L_{s2}} \left( M_{P,2} \right) \quad I_{sc3} = \frac{I_m}{L_{s3}} \left( M_{P,3} \right) \quad (6.28)
\]

The mutual inductances of the above relationships are a function of the geometry of the magnetic structure between the two sides and they can be found by simulation, measurements
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or modelling the physical structure [17, 57]. The power transferred to the secondary side can then be calculated based on the Volt-Ampere of the secondary side as:

\[
S_{u_{\text{total}}} = V_{oc1}I_{w1} + V_{oc2}I_{w2} + V_{oc3}I_{w3} = \omega I_m^2 \left[ \frac{M_{p.s1}^2}{L_{s1}} + \frac{M_{p.s2}^2}{L_{s2}} + \frac{M_{p.s3}^2}{L_{s3}} \right]
\]  

(6.29)

It has been shown that the maximum potential power transfer capability of an uncompensated secondary inductance is about (≈ 0.5*Su) for a single phase IPT system [60]. If the required power to be transferred to the load exceeds this amount, a secondary tuning is required [153]. Therefore, the compensated power of the proposed three-phase system transferred to the load for a certain secondary circuit quality factor Q_s given in (6.31) is:

\[
P_z = Q_s \cdot S_{u_{\text{total}}} = Q_s \cdot \omega I_m^2 \left[ \frac{M_{p.s1}^2}{L_{s1}} + \frac{M_{p.s2}^2}{L_{s2}} + \frac{M_{p.s3}^2}{L_{s3}} \right]
\]

(6.30)

\[
Q_s = \begin{cases} 
\text{Seriestuning} : \frac{\omega_j L_s}{R_L} \\
\text{Parallel tuning} : \frac{R_L}{\omega_j L_s}
\end{cases}
\]

(6.31)

It can be observed from the above analysis that a poly-phase system with various mutual inductances can be simply analyzed similar to a single-phase system. However, the difference is that the mutual inductance here is per pole mutual inductance and it is per-phase in single-phase IPT systems.

6.6 Impedance balance by using cancelation transformer

Because two of the phases (B and C) of the proposed poly-phase system are affected by the air gaps at the two ends of the magnetic structure (see Figure 6-9), its primary side impedances are imbalanced. If the imbalance is considerable, it causes power to flow between the phases, and as a result, significant disturbance to the operation and even failure of the power supply may occur[154]. There are a few solutions such as phase rotation, flux compensation, etc. being investigated to minimize the imbalance [155], but they are mainly proposed for large power transfer systems, which do not suit slipring applications due to the size constraints. In this thesis, a new method of flux balancing the primary impedances using a cancellation transformer is proposed and practically evaluated.
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Figure 6-9: Asummetry of the magnetic structure.

In principle, when a set of three-phase balanced voltages/currents applied to a set of primary coils, their individual fluxes can be expressed as:

\[
\begin{align*}
\phi_a &= \phi_m \cos \omega t \\
\phi_b &= \phi_m \cos (\omega t - 120) \\
\phi_c &= \phi_m \cos (\omega t + 120)
\end{align*}
\]  

For a balanced system without any mutual interaction between the phases, the condition \((\phi_a + \phi_b + \phi_c = 0)\) must hold. In the case of the proposed poly-phase system, the above condition is not valid naturally because of the asymmetry of its magnetic structure. Hence, in order to provide the above balancing condition, the fluxes of the primary phases have to be magnetically coupled additionally as illustrated in Figure 6-10 using cancellation transformer in between the magnetic coupling structure and the power supply. Similar to the main structure, in the cancellation transformer each two coils are 180 degrees out of phase forming one of the primary phases. Based on the connections made, the magnetic flux of each one of the primary coils (not the phases), say \(\phi_a\) (coil-a of phase-A) links coil-c of phase-C in one direction and also links coil-c' of the same phase in the opposite direction. This is valid for all the other coils. This additional magnetic flux coupling due to the cancellation transformer is in the opposite direction to that caused by the main structure. As these two magnetic coupling will be summed up; by properly choosing the number of turns in the cancellation transformer, the net mutual magnetic coupling between the phases can be minimized to almost zero. Based upon the designed geometry for the proposed system, the mutual inductances between the three phases are practically measured before and after cancellation as demonstrated in Table 6-1. From the measured results, it can be observed that using the cancellation transformer method can minimize the interaction between the phases to almost zero. However, the phase
self-inductances will increase slightly as seen after the cancellation which still is not an issue for the system and they can be cancelled using a tuning capacitor if required.

![Magnetic coupling structure](image)

**Figure 6-10: Proposed impedance balancing method.**

**Table 6-1: Measured mutual inductances between the primary phases.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Before</th>
<th>After</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_A$ ($\mu$H)</td>
<td>28.7</td>
<td>32.3</td>
</tr>
<tr>
<td>$L_B$ ($\mu$H)</td>
<td>26.5</td>
<td>32.8</td>
</tr>
<tr>
<td>$L_C$ ($\mu$H)</td>
<td>26.6</td>
<td>32.7</td>
</tr>
<tr>
<td>$M_{AB}$ ($\mu$H)</td>
<td>8.6</td>
<td>$\approx 0.5$</td>
</tr>
<tr>
<td>$M_{AC}$ ($\mu$H)</td>
<td>8.9</td>
<td>$\approx 0.5$</td>
</tr>
<tr>
<td>$M_{BC}$ ($\mu$H)</td>
<td>3.4</td>
<td>$\approx 0.5$</td>
</tr>
</tbody>
</table>

### 6.7 Poly-phase current-fed semi-autonomous converter

A poly-phase current-fed converter with ZVS is proposed in this section to drive the poly-phase travelling magnetic field-based contactless slipring.

#### 6.7.1 Basic principle of operation

The proposed converter uses the improved autonomous topology presented in Chapter 4 as the driving phase (as with the lowest resonant frequency) to allow the switching of other phases to occur at zero voltages. Figure 6-11 shows the proposed poly-phase converter. It consists of
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three single-phase current-fed push-pull converters with a common DC power supply. The DC voltage source for each phase is in series with two large DC inductors as compared to the resonant inductor acting as a quasi-current source. In total the converter is configured as six main switches (S₁ ~ S₆) with common ground for three-phase power inversion. Each phase comprises a parallel tuned resonant circuit which consists of a primary inductor (L₁, L₂ and L₃), a tuning capacitor (C₁, C₂ and C₃), and equivalent reflected load from the secondary onto each primary phase R. The first phase operates autonomously with ZVS, and it is the driving phase of the whole poly phase converter. The other phases follow the driving phase, and their switching frequency is determined by the resonant frequency of the driving phase. To achieve ZVS for the following phases (phases-2 and 3), their resonant frequency should be kept slightly lower than their switching frequency (lower than the resonant frequency of phase-1). In a practical circuit, this can be achieved by designing the resonant tanks of phases-2 and 3 with a slightly higher resonant frequency than the resonant frequency of phase-1.

Figure 6-11: Poly-phase semi-autonomous converter.
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In Figure 6-11, the signals of $S_3$ of the second phase and for $S_5$ of the third phase are taken from the gate of $S_1$ and delayed by 120° and 240° phase displacement. The other switches of phase-2 and 3 ($S_4$ and $S_6$) are operating autonomously similar to phase-1. As it can be observed the major part of the gate drive circuit is achieved by the converter and only two timers are generated by an additional circuit as required. Based on the presented principle of the improved autonomous converter in Chapter 4 (Section 4.3.3) a ZVS operation is determined by the polarity of the resonant voltage across $C_1$ autonomously without any need for external controller and start-up circuitry. The poly-phase converter of Figure 6-11 also is able to start-up automatically owing to the driving phase (phase-1). A detailed start-up analysis is presented in Chapter 4 (Section 4.3.2). Once the driving phase started, signals will be fed to the controller to calculate the frequency of the driving-phase and accordingly turn ‘on’ the other phases with 120° and 240° phase delays as presented in the next section.

6.7.2 Dynamic phase-delay approach

In the case of poly-phase contactless slipring systems with moving magnetic field, it is required to provide 120° phase delay between the phases. In typical poly-phase power supplies with fixed switching frequency, this is normally achieved as a fixed delay. Fixed delay approach normally results in hard switching operation, EMI and power losses [156, 157]. For the proposed converter, a dynamic delay approach is realized by using FPGA-mode of an NI-C-RIO (National Instruments Compact-RIO) controller. NI C-RIO incorporates a real-time processor with operating frequency about 200 MHz, reconfigurable FPGA and swappable I/O modules [158]. The software integrated with this controller for programming is LabVIEW. This provides a graphical frame work and includes a wide range of built-in functions such as PID control, zero crossing detection etc. Using C-RIO in combination with LabVIEW and taking the advantage of user-programmable FPGAs, highly optimized reconfigurable control and acquisition systems can be realized.

Figure 6-12 shows the generated graphical control coding of the dynamic delay approach. The advantage of this approach is that when the frequency of phase-1 varies, the phase delays fed to the other phases adjust accordingly. As it can be seen a gate signal taken from phase-1(from $S_1$) is fed to the first block as an input (Mod6/DIO0). The period of phase-1 then is calculated and accordingly two phase delays are generated based on the calculated period (Delay Ph-2 and Ph-3). These two delays are added to the switching instant of $S_3$ and $S_5$ for other phases as shown in the second and third blocks of the graphical code. This can be seen from the second
block that the same signal from $S_1$ (Mod6/DIO0) is delayed by (Delay Ph-2) and then fed to (Mod3/DIO4) to turn ‘on’ $S_3$ of phase-2. A similar procedure is applied for the third phase as illustrated in the last block providing switching signal (Mod3/DIO5) for $S_5$.

Figure 6-12: Coding sequence for dynamic delay approach.

6.7.3 Simulation and experimental results on the poly-phase semi-autonomous converter

Simulation study is carried out using the circuit parameters of $L_1 = L_2 = L_3 = 33$ µH, $C_1 = C_2 = C_3 = 470$ nF, $L_{dc1} \sim L_{dc6} = 1$ mH and $V_{dc} = 12$ V. The simulation package used is LTspice IV. Two N-channel MOSFETs (IRFP240) with 200V/20A/0.18Ω are used for operation which are similar to the practically used switches. Figure 6-13 shows the gate waveform of the three phases (half-wave) accurately delayed with 120°, while precisely following the resonant voltage and operating with ZVS. It can be seen that the resonant tanks oscillation continued due to zero points detection captured by the clamping diodes and switching occurring at the zero point defined by the operating frequency of phase-1. Steady state simulation results of the resonant currents and voltages are good sinusoidal waveforms at 40 kHz as shown in Figure 6-14.
Figure 6-13: Simulation results of the gate signal operation.

Figure 6-14: Steady state simulation results for equal resonant inductors $L_1=L_2=L_3=33 \, \mu H$.

The above simulation is completed for equal resonant frequencies for all the phases. However, this is not the case in practice and due to parameters differences it is difficult to design the
resonant frequency same for all phases. Different resonant frequencies for the phases may lead to hard switching for the phases that follow the driving phase. This will take place if the following phases (phase-2 and 3) are switched at a frequency higher than their resonant frequency. For instance for resonant capacitors \( C_1 = C_2 = C_3 = 470 \, nF \) and resonant inductances \( L_1 = 33 \, \mu H \) and \( L_2 = L_3 = 35 \, \mu H \), the resonant frequency of the driving phase (phase-1) is 40 kHz \( (T = 25 \, \mu s) \) and the resonant frequency of other phases is about 39 kHz \( (T=25.64 \, \mu s) \). At these conditions, based on the calculated period of phase-1 \( (T=25 \, \mu s) \) and the generated phase delays for following phases, their switching instants occur earlier than zero point (where their resonant voltages are still not zero) as shown in Figure 6-15. To solve this problem and achieve ZVS for all phases, the driving phase should be the phase with the lowest resonant frequency (highest resonant inductance for similar resonant capacitors). Although the two following phases are not in full resonance, but their switching operations occur at an instant with zero volt. In fact at this condition, the switching frequency of the two following phases (which is the resonant frequency of phase-1) is set to be lower than their resonant tanks frequency. This results in a small blank time with zero volts which provides a safe region for switching occurrence as seen in Figure 6-15(b) for \( (L_1 = 33 \, \mu H \) and \( L_2 = L_3 = 31 \, \mu H) \).

![Graph](image)

(a) \( L_1 = 33 \, \mu H \) and \( L_2 = L_3 = 35 \, \mu H \)

(b) \( L_1 = 33 \, \mu H \) and \( L_2 = L_3 = 31 \, \mu H \)

Figure 6-15: Resonant voltages for unequal resonant inductors.
The proposed converter is practically built and tested using the same parameters as that of the simulation study for comparison. Figure 6-16 (a & b) show the practical resonant voltages and resonant currents of the three tanks. It can be seen that similar to the simulation study the captured waveforms are very good sinusoidal with accurate 120° phase delays. It can be observed from the gate waveforms shown in Figure 6-16(c) that all the phases are switching at 40.3 kHz following the resonant frequency of phase-1. Figure 6-17 shows the zero voltage switching operation for individual phases. It can be seen that precise ZVS is practically achieved for all phases at 40.3 kHz making the proposed converter a good option for poly-phase IPT systems. As the switching frequency of all phases determined by the autonomous driving phase, based on the ZVS operating frequency analysis method developed in Chapter 4 (Section 4.3.3.2), the actual ZVS frequency of the driving phase (phase-1) calculated is about 40 kHz with the given circuit parameters, which is in a very good agreement with the practically measured frequencies of 40.3 kHz.

Figure 6-16: Practical waveforms of the resonant tanks.
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Figure 6-17: ZVS operation of the gate signals at 40.3 kHz.
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6.8 Simulation and experimental results of the poly-phase system

A 3D-FEM model using JMAG package is developed to study the magnetic characteristics and power transfer capability of the system as shown in Figure 6-18. A practical set-up is also constructed for evaluation as shown in Figure 6-19. The prototype system is driven by the semi-autonomous poly-phase current-fed converter presented in the previous section. The simulation/practical data is shown in Table 6-2. The power transfer capability of the proposed system as a product of open-circuit voltages and short-circuit currents is demonstrated in Table 6-3 for a poly-phase power pickup and in Table 6-4 for a single-phase power pickup respectively. Typical waveforms of the open-circuit voltages and short-circuit currents are taken from practical testing with the poly-phase power pickup and shown in Figure 6-20, demonstrating 120° phase shifts among three phases. The single power pickup uses the same amount of copper and ferrite materials as the poly-phase power pickup, but captures only half of the power captured by the poly-phase power pickup. Good agreement with negligible error between the simulation and the practical results can be noticed, signifying the feasibility of the proposed system and its potential of power transfer. The maximum total unregulated power achieved is about 523 VA without any secondary tuning. As stated in [60] maximum real power about (0.5*Su) can be transferred to the load without any compensation at the secondary side. This statement can be observed from the graph of the load test shown in Figure 6-21(a); the maximum real power transferred to a resistive load is reached to about 255.6W (≈ 0.5*523 VA) at 12.5Ω. At this output power the efficiency is simulated about 97%. The simulation results of power losses within the magnetic coupling structure are shown in Figure 6-21(b). As it can be seen the total loss due to primary coils (Pw1) is constant at about 3W for all the load values because of the constant primary currents. The loss of the secondary coils (Pw2), on the other hand is variable because of the variable secondary current corresponding to the load variations. The power loss within the ferrite is maximum at light loads about 5W and stabilizes gradually at 3 W, as the load increases.

The system is investigated under compensation (series and parallel) conditions at the secondary side. The maximum possible Qs for the system is normally limited in practice for some reasons. In the case of the practical model of the proposed system, the secondary used wire gauge is AWG12 which can handle maximum current about 10.8 A @40°C; thus for the series-tuning case, the short-circuit current of the system shown in Table 6-3 can be increased to about 10A. This limits Qs to about 2.44. Since the operating frequency and secondary inductances are known for the circuit, the quality factor of is determined by the load.
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Therefore, the system tested for both series and parallel-tuned secondary side at Qs of 2.44 and the results are shown in Table 6-5. It can be seen that the proposed system can deliver power in kW range (about 1.2 kW) at about 97% power efficiency for contactless slipring applications.

Figure 6-18: 3D-FEM model of the proposed travelling magnetic field-based system.

Figure 6-19: Practical setup of a travelling magnetic field-based system.
Table 6-2: Simulation and practical parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_p$ (A)</td>
<td>5</td>
</tr>
<tr>
<td>$N_p$ per phase/ESR (Ω)</td>
<td>8 / 0.08</td>
</tr>
<tr>
<td>$N_s$ per phase/ESR (Ω)</td>
<td>8 / 0.068</td>
</tr>
<tr>
<td>f (kHz)</td>
<td>50</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>3</td>
</tr>
<tr>
<td>Ferrite</td>
<td>Mn-Zn ($\mu=2500$, $B_s=0.5$ T)</td>
</tr>
<tr>
<td>Total primary length (mm)</td>
<td>90</td>
</tr>
<tr>
<td>Outer diameter (mm)</td>
<td>42</td>
</tr>
</tbody>
</table>

Table 6-3: Simulation and practical results with poly-phase power pickups (Uncompensated conditions)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Simulation</th>
<th>Experiment</th>
<th>Error (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{oc1}$ (V)</td>
<td>40</td>
<td>43.1</td>
<td>7</td>
</tr>
<tr>
<td>$V_{oc2}$ (V)</td>
<td>38</td>
<td>41.3</td>
<td>8.6</td>
</tr>
<tr>
<td>$V_{oc3}$ (V)</td>
<td>38</td>
<td>41.3</td>
<td>8.6</td>
</tr>
<tr>
<td>$I_{sc1}$ (A)</td>
<td>4.47</td>
<td>4.17</td>
<td>7</td>
</tr>
<tr>
<td>$I_{sc2}$ (A)</td>
<td>4.5</td>
<td>4.3</td>
<td>4.6</td>
</tr>
<tr>
<td>$I_{sc3}$ (A)</td>
<td>4.5</td>
<td>4.31</td>
<td>4.6</td>
</tr>
<tr>
<td>Total Su (VA)</td>
<td>523</td>
<td>534.9</td>
<td>2.2</td>
</tr>
</tbody>
</table>

Table 6-4: Simulation and practical results with single-phase power pickups (Uncompensated conditions)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Simulation</th>
<th>Experiment</th>
<th>Error (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{oc1}$ (V)</td>
<td>85.86</td>
<td>87</td>
<td>1.3</td>
</tr>
<tr>
<td>$I_{sc1}$ (A)</td>
<td>3.24</td>
<td>3.5</td>
<td>8</td>
</tr>
<tr>
<td>Total Su (VA)</td>
<td>287.52</td>
<td>304.5</td>
<td>5.9</td>
</tr>
</tbody>
</table>

(a) Open-circuit voltages

173
Figure 6-20: Experimental results with poly-phase power pickup: Output waveforms.

(a) Output power and efficiency

(b) Power losses: all primary coils (Pw1), all secondary coils (Pw2) and core loss (Pc)

Figure 6-21: Simulation results (Load test for uncompensated case).
Chapter 6: Poly-phase contactless slipring based on axial travelling magnetic field

Table 6-5: Simulation/Experimental results with poly-phase power pickup for series and parallel tuned secondary at Qs=2.44

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Series</th>
<th>Parallel</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total output power (W)</td>
<td>1205.083</td>
<td>1205.836</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>97.647</td>
<td>97.32</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Experiment</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Total output power (W)</td>
<td>1260</td>
<td>1262</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>95.6</td>
<td>95</td>
</tr>
</tbody>
</table>

6.9 Comparison and discussion

Basically the travelling magnetic field-based system uses 6 single-phase structures. In this section, the power transfer capability of the poly-phase system is compared against a single-phase structure. The single-phase of Figure 6-22 which is (1/6) of the total amount of used copper and ferrite materials is studied based on FEM study as well as practical experiments. Table 6-6 shows the simulation and practical results of the $V_{oc}$ and $I_{sc}$ for the single-phase system. Similar to the proposed poly-phase system, a load test is simulated and the output power is multiplied by 6 to balance both systems based on the used materials. The graph of Figure 6-23(a) demonstrates the unregulated output power without tuning for both cases, the power efficiencies and the total power losses are illustrated in Figure 6-23(b) and (c).

The maximum power transferred to the load is about 68.4W at 90% efficiency for 6 individual single-phase systems and about 255.6 W at 97% efficiency for the proposed integrated poly-phase system. This is about 3.7 times more than a set of 6 individual single-phase systems and about 22 times more than one single-phase structure with about 11.4W maximum output power.

Table 6-6: Simulation and experimental results of a single-phase coaxial system

<table>
<thead>
<tr>
<th>Approach</th>
<th>$V_{oc}$ (V)</th>
<th>$I_{sc}$ (A)</th>
<th>$S_u$ (VA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Simulation</td>
<td>6.6</td>
<td>3.75</td>
<td>24.75</td>
</tr>
<tr>
<td>Experiment</td>
<td>7</td>
<td>3.9</td>
<td>27.3</td>
</tr>
</tbody>
</table>
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Figure 6-22: Single phase pulsating magnetic field-based system.

(a) Un-regulated output power

(b) Efficiency
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The fundamental difference between the single-phase system and the proposed system is in the induction fashion. For the single-phase system, the induced voltage in the secondary coil occurs due to the time-varying magnetic field. Whereas, for the proposed system the induction is due to a motional mmf with a constant magnitude contributed by multi-phases. A magnetic field with constant magnitude does not go into the cycling of the hysteresis loop as it is the case for a pulsating magnetic field. This results in a very low core loss in the system and accordingly increases the power transfer capability and efficiency. Moreover, because of higher magnetic flux density (1.5 times the time-variable type of field), the primary current can be reduced, and consequently the power loss within the coils ($I^2R$) would reduce.

It should be noted that although the linear velocity of the moving field can be high depending on the pitch hole length (see Eqn.(6.14)), its actual induction speed determined by the magnetic field variation frequency across the coils and ferrites stays the same as the pulsating field. This is because for the pulsating field, it is clear that the induction frequency is determined by the operating frequency of the current; and for the proposed travelling field the magnetic variation speed across the coil and ferrite is also determined by the operating frequency based on rotating magnetic field theory (extended to linear movement in this case). As a result, the loss mechanisms of the two systems are fundamentally the same, although the directions of flux variation in relation to the ferrite cross sections are different. In a practical design, the differences in ferrite shapes may cause slightly different eddy current and hysteresis losses, but there is no major impact on power losses and excitation power. Figure 6-23(c) demonstrates that the total power losses of the two systems are very close under...
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Heavy loading conditions although the difference increases with the reduction of the load resistance. Because the proposed system based on a moving field increases the overall power transfer capability, a higher output power makes its power efficiency higher when the total power loss does not change much.

6.10 Summary

This chapter has presented a poly-phase travelling magnetic field-based contactless slipring system. Single-phase and poly-phase power pickups are investigated and it has shown that the poly-phase power pickup is able to capture twice power as that captured by a single-phase power pickup.

To drive the poly-phase system, a poly-phase current-fed push-pull resonant converter with ZVS operation led by one of the phases has been proposed in this chapter. It has been shown that designing one autonomous phase with the lowest resonant frequency as the driving phase and allowing other phases to follow, full ZVS for all phases can be achieved although the following phases may not be in full resonance. Accurate dynamic 120° phase displacements with ZVS operation are verified by experimenting at about 40 kHz. The simulation and experimental results are in a very good agreement, and they have demonstrated that the proposed converter is able to generate poly-phase high quality currents for poly-phase inductive power transfer systems.

It has been found that the proposed poly-phase system can transfer about 3.7 times more than a set of 6 individual single-phase coaxial systems and about 22 times more than a one single-phase structure at much higher efficiency. Maximum power transferred to the load is about 68.4W at 90% efficiency for 6 individual single-phase systems and about 255.6 W at 97% efficiency for the proposed poly-phase system. The main fundamental difference between a single-phase system and a poly-phase system is in the induction occurrence. For the single-phase system, the induced voltage in the secondary coil occurs due to the time-varying magnetic field, while for the poly-phase system occurs by a motional mmf as a resultant of multi-phases. The resultant magnetic field of the travelling magnetic field-based system is a special travelling type with a constant magnitude (1.5 times the time-variable type of field) which does not go into the cycling of the hysteresis loop as it is the case for a pulsating magnetic field. This results in a low core loss in the system and accordingly increased power.
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efficiency. Moreover, because of this higher magnetic flux density, the primary current then can be reduced, and consequently the power loss within the coils (I^2R) would reduce.

The practical results have verified that the inter-phase mutual inductances can be reduced to almost zero by applying a cancellation transformer method.
Chapter 7: Poly-phase contactless slipring system based on rotating magnetic field

7.1 Introduction

Further to the poly-phase contactless slipring based on travelling magnetic field presented in Chapter 6, this chapter proposes another new poly-phase contactless slipring system based on rotating magnetic field around the shaft of the system. The purpose of the rotating magnetic field in this system is to deliver power to the secondary power pickup coils rather than generating a mechanical driving torque on the shaft. A new poly-phase power converter with full-autonomous operation of all the phases is developed to drive the poly-phase system.

The chapter begins by describing the basic principle of operation of the rotating magnetic field-based system. This is followed by a detailed study of a mutually coupled equivalent electric circuit and analytical expressions are derived for system performance analysis. FEM simulation studies are conducted for system assessment and verifications. A system prototype is constructed for experimental measurements and evaluation. Finally, a comparison between the rotating magnetic field and typical pulsating field is conducted based on FEM and experimental study.

7.2 The proposed system and its operating principle

Figure 7-1(a) shows the block diagram of a Rotating Magnetic Field-based (RMF-based) system and its general circuit is shown in Figure 7-1(b). The system comprises of a primary power supply, a set of three-phase stationary distributed coils sharing a common ferrite core, a set of three-phase rotatable distributed secondary coils sharing a common ferrite core and a secondary converter. The primary power supply comprises three identical current-fed push-pull resonant converters sharing a common DC source and 120˚ electrically shifted in phase as shown in Figure 7-1(b). A detailed presentation about the employed converter is conducted in Section 7.9 of this chapter. Due to magnetic coupling between the primary and the secondary coils, voltage is induced in the secondary coils. This voltage then is rectified and regulated according to the load requirements by the secondary power converter.
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If an alternating current flows through phase-A, it produces a pulsating mmf wave along its magnetic axis and having a peak located along the axis. The amplitude and direction of this wave depends on the instantaneous value of the current flowing through the coil as illustrated in Figure 7-2. Each phase coil produces a similar sinusoidally distributed mmf wave, but displaced by 120 electrical degrees from each other. Therefore, the resultant mmf wave at any point in the air gap, defined by an angle θ, is the net effect of the three component mmf waves. The origin of θ can be chosen to be the axis of phase-A, as shown in Figure 7-2.

Figure 7-2: Instantaneous values of mmf distribution in phase-A.
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The balanced three-phase instantaneous currents flowing through the three-phase primary coils can be expressed as:

\[ i_a = I_m \cos \omega t \]
\[ i_b = I_m \cos (\omega t - 120^\circ) \]
\[ i_c = I_m \cos (\omega t + 120^\circ) \]  \hspace{1cm} (7.1)

At any instant of time, all three phases contribute to the air gap mmf along the path defined by the angle \( \theta \). Therefore, the resultant mmf along \( \theta \) is:

\[ F(\theta) = F_a(\theta) + F_b(\theta) + F_c(\theta) \]  \hspace{1cm} (7.2)

The contributions from the individual phases along \( \theta \) then can be given as:

\[ F_a(\theta) = N_i_a \cos \theta \]
\[ F_b(\theta) = N_i_b \cos (\theta - 120^\circ) \]  \hspace{1cm} (7.3)
\[ F_c(\theta) = N_i_c \cos (\theta + 120^\circ) \]

where \( N \) is the effective number of turns per phase and \( i_a, i_b, i_c \) are the phase currents.

Substituting the above relationships of (7.3) in (7.2) give the resultant mmf at angle \( \theta \) as follows:

\[ F(\theta) = N_i_a \cos \theta + N_i_b \cos (\theta - 120^\circ) + N_i_c \cos (\theta + 120^\circ) \]  \hspace{1cm} (7.4)

Combining (7.1) and (7.4), gives the resultant mmf wave as a function of time and location as:

\[ F(\theta, t) = N I_m \cos \omega t \cos \theta \]
\[ + N I_m \cos (\omega t - 120^\circ) \cos (\theta - 120^\circ) \]
\[ + N I_m \cos (\omega t + 120^\circ) \cos (\theta + 120^\circ) \]
\[ = \frac{3}{2} N I_m \cos (\omega t - \theta) \]  \hspace{1cm} (7.5)

Likewise a resultant flux density of the above mmf can be expressed as:

\[ B(\theta, t) = \frac{3}{2} B_m \cos (\omega t - \theta) \]  \hspace{1cm} (7.6)

The generated magnetic field of (7.5) rotates at a much greater angular velocity (several kHz) than the speed of the shaft; thus the rotation of the shaft does not affect the system.
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performance. Taking into account the principle of operation, if the primary magnetic field rotates in the opposite direction of the shaft, the speed of the shaft adds up to the speed of the magnetic field. As a result, greater voltages induced in the secondary coils as stated below:

\[ e_z = N_2 \times \left[ B \times (v_f + v_{sh}) \right] \times l \]  

(7.7)

where \( N_2 \) is the total secondary coil number of turns, \( B \) is the magnitude of the rotating magnetic field, \( v_f \) is the velocity of the rotating magnetic field, \( v_{sh} \) is the velocity of the shaft and \( l \) is the length of secondary coil turns.

7.3 Primary magnetic structure layouts

7.3.1 Design 1 (with magnetic axes in parallel with the planes of coils)

Figure 7-3 shows the overall geometry of the primary side of design-1 of the rotating magnetic field-based contactless slipring system (RMF-D1). The system is a two-pole structure which the phases are physically and electrically 120° out of phase. Each phase consists of 2 coils with 180° phase shift. For instance, two coils (a and a’) forming phase-A, and similar to that (b and b’) and (c and c’) forming the other two phases B and C. The magnetic axes of the primary phases (A-A’, B-B’ and C-C’) are in parallel with the planes of the coils of each phase as shown in Figure 7-3 [145].

Figure 7-3: Design-1 of RMF-based system (RMF-D1).

It is important to understand the motion theory of the field and how this field travels in the air gap, to gain better standpoints for designing the secondary power pickups. Figure 7-4 demonstrates the motion of the rotating magnetic field around the shaft in the air gap.
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Consider the state at \( t=0 \), the moment when the current of phase-A is at its maximum value \( I_m \) in the positive direction. The mmf of phase-A then has its maximum value \( F_A = F_{\text{max}} \) along the magnetic axis of phase-A as shown in Figure 7-4(a). At this moment, currents \( i_b \) and \( i_c \) are both \( I_m/2 \) in the negative direction. The corresponding mmf's of phases-B and C are both of magnitude \( F_{\text{max}}/2 \) along their magnetic axes respectively. The resultant, obtained by adding the individual contributions of the three phases, is a vector of magnitude \( F = 3/2 F_{\text{max}} \) centered on the axis of phase-A. It represents a sinusoidal space wave with its positive peak centered on the axis of phase-A and having amplitude 3/2 times that of the phase-A contribution alone.

After 60 electrical degrees at \( \omega t = \pi/3 \) (Figure 7-4(b)), the currents in phases A and B are positive half maximum, and that in phase-C is a negative maximum. Therefore, the magnetic field at this moment is concentrated between two coils \( c' \) and \( c \) of phase-C. The resultant has the same amplitude as at \( t=0 \), but it has now moved to counter clockwise 60 electrical degrees around the air gap. Similarly, at \( \omega t = 2\pi/3 \) when the current of phase-B current is a positive maximum and phase-A and phase-C currents are a negative half maximum (Figure 7-4(c)), the same resultant mmf distribution is again obtained, but it has moved counter clockwise 60 electrical degrees further and is now aligned with the magnetic axis of phase-B. As time passes, the resultant mmf wave retains its sinusoidal form and amplitude but moves progressively in the air gap between the primary and secondary sides; the net result can be seen to be an mmf wave of constant amplitude rotating at a uniform angular velocity. In one cycle the resultant mmf must be back in the position of Figure 7-4(a). The mmf wave therefore makes one revolution per electrical cycle as illustrated in Figure 7-4(d, e, f). Note that for easy comparison, the geometry of the simulated model shown in Figure 7-4 are based on the practical model built for evaluation, so they are in hexagon rather than a circle.

(a) \( t=0 \) and \( \theta = 0 \) (\( F_a = F_{\text{max}} \), \( F_b = F_{\text{c}} = F_{\text{max}}/2 \))
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(b) $t = \pi/3$ and $\theta = 60 \degree (F_c = F_{\text{max}}$, $F_b = F_a = F_{\text{max}}/2)$

(c) $t = 2\pi/3$ and $\theta = 120 \degree (F_b = F_{\text{max}}$, $F_a = F_c = F_{\text{max}}/2)$

(d) $t = \pi$ and $\theta = 180 \degree (F_a = F_{\text{max}}$, $F_b = F_c = F_{\text{max}}/2)$
7.3.2 Design 2 (with magnetic axes perpendicular with the planes of coils)

In this design, the primary coils are wound around a toroidal shape of ferrite to enhance the magnetic coupling by reducing some of the field cancellation associated with the previous design. Figure 7-5 shows a general magnetic structure of the RMF-D2 system. As it can be seen the magnetic axes of the primary phases (A-A’, B-B’ and C-C’) are perpendicular to the planes of the coils of each phase, while in the previous design the magnetic axes of the phases are in parallel with the coils of each phase. This would results in some flux cancellations in the corners between each pair of adjacent coils as illustrated earlier in Figure 7-3. This flux cancellation is eliminated in RMF-D2 structure by winding the primary coils in toroidal fashion. In a toroidal structure, the magnetic field is normally confined to the inside and almost zero in the outside the toroid which results in a self-shielded system[159].
resultant magnetic field in this design (RMF_D2) moves in the same fashion as in the previous design (RMF_D1) as illustrated in Figure 7-6.

Figure 7-5: Design-2 of the RMF-based system (RMF-D2).

(a) \( t=0 \) and \( \theta =0 \) (\( F_a=F_{\text{max}}, F_b=F_c=F_{\text{max}}/2 \))

(b) \( t=\pi/3 \) and \( \theta =60 \) (\( F_a=F_{\text{max}}, F_b=F_c=F_{\text{max}}/2 \))
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(c) $t=2\pi/3$ and $\theta = 120$ ($F_b = F_{\text{max}}, F_a = F_c = F_{\text{max}}/2$)

(d) $t=\pi$ and $\theta = 180$ ($F_a = F_{\text{max}}, F_b = F_c = F_{\text{max}}/2$)

(e) $t=4\pi/3$ and $\theta = 240$ ($F_c = F_{\text{max}}, F_b = F_a = F_{\text{max}}/2$)
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(f) \( t=5\pi/3 \) and \( \theta = 300 \) (\( F_b=F_{\text{max}}, F_a=F_c=F_{\text{max}}/2 \))

Figure 7-6: Motion of the magnetic field around the air gap (RMF_D2).

7.4 Secondary power pickup layouts

Since the primary magnetic field of the proposed system is a 2-pole rotating magnetic field; the magnetic field intensity \( H \) in the air gap at any angle \( \theta \) under one pole is the same in magnitude as that at angle \( (\theta+\pi) \) under the opposite pole, but the fields are in the opposite direction. This signifies that, the voltage induced in the secondary coils due to one pole is same in magnitude and opposite of that from the other pole (\( e_1 \) and \( e_2 \)). Therefore, unlike the primary, the secondary side does not need to be three-phase configuration; but, it has to be designed based on the number of poles of the primary magnetic field. This is illustrated in Figure 7-7 (a) that the two secondary coils are 180 degrees out of phase and physically positioned over the full pitch of the primary. Each coil of the secondary positioned under one pole of the magnetic field and other side of the coil is under the other pole and connected in series in the opposite direction to provide the total voltage induced in the secondary coil.

The total voltage induced in the secondary power pickup of Figure 7-7 (a) is a summation of \( e_1 \) and \( e_2 \). This total induced voltage can be diverted to one side by shorting one of the poles as shown in Figure 7-7 (b) for the second design of secondary power pickup. In fact by shorting one of the poles, an mmf is generated in the shorted pole and opposes the primary mmf passing through that pole. As a result, all the primary flux lines will be diverted to the other side doubling the induced voltages in the other pole. Although, the total induced voltage of the unipolar design is equal to the 2-poles design, but it is simpler in design and uses half the copper material which it may be considerable as the size of the system increases.
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(a) Typical 2-poles design  (b) Unipolar design

Figure 7-7: Single-phase power pickups.

The other power pickup design is a balanced 3-phase layout as shown in Figure 7-8. The advantages of this design is that it captures most of the power due to all the primary phases and also helps to reduce the output filter volume after rectification as it is a balanced three-phase system. Thus, similar to the primary arrangement, 6 secondary coils wound around a toroidal ferrite are used to form three-phase secondary winding with 120° physical displacement. The two coils (1, 2) form phase-S₁, (3, 4) phase-S₂ and (5, 6) form phase-S₃ of the secondary side. This arrangement assures that for each secondary phase there will always be one of the coils that falls under one pole, while the other coils falls under the other pole of the generated magnetic field forming balanced 3-phase outputs. It should be noted that toroidal shape of the secondary ferrite results in zero applied torque on the secondary side due to the primary magnetic field as presented later in Section 7.8 of this chapter [145].

Figure 7-8: Poly-phase power pickups design.
7.5 Resultant per pole flux and the induced voltages

The rotating magnetic field of (7.6) will induce voltages in the secondary coils which can be obtained by using Faraday’s laws of induction. The rotating magnetic field is a vector in the air gap and varies sinusoidally with mechanical angle \( \theta \). Since the direction of \( B \) is always radially outward, the flux density distribution in the air gap can be expressed as:

\[
B(\theta) = B_{\text{peak}} \cos(\theta)
\]  

(7.8)

The air gap flux per pole for the two-poles proposed system is expressed as [53]:

\[
\phi_p = \int_{0}^{\pi/2} B(\theta) l \, d\theta = B_{\text{max}} l r \left[ \sin(\frac{\pi}{2}) - \sin(-\frac{\pi}{2}) \right] = 2 B_{\text{max}} l r
\]  

(7.9)

where \( r \) is the average radius at the air gap and \( l \) is the axial length of the primary side.

As the rotating magnetic field moves (or the magnetic poles rotate) the flux linkage with the secondary coils varies accordingly. For instance, the flux linkage for phase-S1 (coils 1 & 2 of the 3-phase secondary layout) will be maximum and equal \((= N_{s1} \phi_p)\) at \( \theta=0 \) and zero at \( \theta=90 \). This indicates that the flux linkage of \( \phi_p \) with phase-S1 will vary as the Cosine of the angle \( \theta = \omega t \) and expressed by:

\[
\lambda_{s1}(\omega t) = N_{s1} \phi_p \cos \omega t
\]  

(7.10)

Therefore, the voltage induced in phase-S1 obtained from Faraday’s law is:

\[
e_{s1} = \frac{d\lambda_{s1}}{dt} = -(N_{s1} \frac{d\phi_p}{dt} \cos \omega t + \omega N_{s1} \phi_p \sin \omega t)
\]  

(7.11)

The first term of (7.11) is the transformer induced voltage equation and it is present only if the amplitude of the air gap flux wave changes with time. The second term is the speed voltage generated by the relative motion of the air gap flux wave and the secondary coils. Since, in the normal steady-state operation, the amplitude of the air gap flux wave is constant; the first term is zero and the generated voltage is basically the speed voltage:

\[
e_{s1} = -\omega N_{s1} \phi_p \sin \omega t = E_{\text{max}} \sin \omega t
\]  

(7.12)

Likewise for the other two phases of the secondary, the induced voltages are 120° shifted in phase as given by:
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\[ e_{s2} = E_{\text{max}} \sin(\omega t - 120^\circ) \]  
\[ e_{s3} = E_{\text{max}} \sin(\omega t + 120^\circ) \]

The rms value of the induced voltage per secondary phase then can be conveyed as:

\[ E_{\text{rms}} = \frac{\omega N_s \phi_p}{\sqrt{2}} = \frac{2\pi f}{\sqrt{2}} N_s \phi_p = 4.44 f N_s \phi_p \]

The above equation has the same form as that for the induced voltage in transformers. However, \( \phi_p \) represents the resultant flux per pole of the system which is 1.5 times the flux due to one phase alone.

### 7.6 Poly-phase mutual inductance equivalent circuit

For the rotating magnetic field-based system, a similar poly-phase model (see Figure 7-9) presented in Chapter 6 (Section 6.5) can be used to study and predict the performance of the system.

![Figure 7-9: Poly-phase mutual inductance model.](image)

However, because of the symmetry of the RMF-based systems, the introduced ‘per pole mutual inductances’ in Chapter 6 (Section 6.5) are equal here and can be expressed by:
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\[
M_p = \left[ M_{A1} + \left( \frac{1}{2} \right) (M_{B1} + M_{C1}) \right]
\]
\[
\approx \left[ M_{B2} + \left( \frac{1}{2} \right) (M_{A2} + M_{C2}) \right]
\]
\[
\approx \left[ M_{C3} + \left( \frac{1}{2} \right) (M_{A3} + M_{B3}) \right]
\]

(7.16)

Due to uniform symmetry of the RMF-based system, the inductances of the secondary phases are practically equal (\(L_{S1}=L_{S2}=L_{S3}=L_S\)), thus the open-circuit voltages and short-circuit currents can be rewritten based on \(M_p\) and \(L_S\) as:

\[
V_{oc1} \approx V_{oc2} \approx V_{oc3} = \omega I_m M_p = V_{oc}
\]

(7.17)

\[
I_{sc1} \approx I_{sc2} \approx I_{sc3} = I_m \left( \frac{M_p}{L_s} \right) = I_{sc}
\]

(7.18)

The power transferred to the secondary side can then be calculated based on the Volt-Ampere of the secondary side from:

\[
S_{u,A} = V_{oc1} I_{sc1} = \omega \frac{M_p^2}{L_s} I_m^2 = S_{u,B} = S_{u,C}
\]

(7.19)

The compensated power for a certain \(Q_s\) then can be expressed by:

\[
P_2 \approx 3 S_{u,A} = 3 Q_s \omega \frac{M_p^2}{L_s} I_m^2
\]

(7.20)

where:

\[
Q_s = \begin{cases} 
\text{Series tuning} : & \frac{\omega_0 L_s}{R_L} \\
\text{Parallel tuning} : & \frac{R_L}{\omega_0 L_s}
\end{cases}
\]

(7.21)

A higher \(Q_s\) would help increase the output power although practical \(Q_s\) needs to be less than 10 [60]. It should be noted that for the parallel tuned case, the pickup coil performs as a current source, and for a series tuned condition it acts as a voltage source. The tuned output of the pickup coil is then rectified and regulated to a constant output voltage to drive the load. Based on the tuning topology, different buck or boost power regulation methods may be used on the secondary side. In practice, a parallel tuned power pick-up with a boost controller is
commonly used in such applications due to its voltage boosting capability with a natural current limiting characteristic. However, the maximum current that a parallel tuned power pickup can supply is the coil short circuit current \([61, 113]\).

7.7 Reflected impedance analysis

In this section the reflected impedances from the secondary side onto primary phases presented based on the introduced mutual inductance per pole. The reflected impedance \((Z_r)\) from the secondary to each primary phase is determined indirectly from the voltages that are induced in the primary phases and the primary currents as [152]:

\[
Z_{r1} = \frac{V_{r1}}{I_m} \quad Z_{r2} = \frac{V_{r2}}{I_m} \quad Z_{r3} = \frac{V_{r3}}{I_m}
\]  
(7.22)

The reflected voltages of (7.22) on the other hand are determined by the current flowing within the secondary coil and the mutual inductance between the two sides as given by:

\[
V_{r1} = -\left[\omega I_{sc1}(M_P)\right] \\
V_{r2} = -\left[\omega I_{sc2}(M_P)\right] \\
V_{r3} = -\left[\omega I_{sc3}(M_P)\right]
\]  
(7.23)

Combining (7.23) and (7.22) gives the reflected impedances as:

\[
Z_{r1} = Z_{r2} = Z_{r3} = \frac{-\left[\omega I_{sc}(M_P)\right]}{I_m}
\]  
(7.24)

The short circuit current \(I_{sc}\) of each secondary phase is obtained as a result of open circuit voltage divided by the secondary impedance of that phase as:

\[
I_{sc} = \frac{V_{sc}}{Z_s} = \frac{\omega I_{m} M_p}{Z_s}
\]  
(7.25)

Substituting \(I_{sc}\) from (7.25) into (7.24) and simplifying, gives the reflected impedances onto each primary phase as:

\[
Z_{r1} \approx Z_{r2} \approx Z_{r3} = \frac{\omega^2 M_p^2}{Z_s}
\]  
(7.26)
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Since the secondary network depends on the selected compensation topology, the reflected resistance and reactance at the resonant frequency onto each phase of the primary side are different as given in Chapter 2 (Section 2.5.5.2) [63, 64]. The power transfer from the primary to the secondary is simply the reflected resistance (real part of the reflected impedance) multiplied by the squared primary current. The total power transferred to the secondary side as a result of three phases then can be given by:

\[
P_2 = 3 \times \left[ \text{Re}(Z_r) \right] I_m^2
\]

(7.27)

### 7.8 Applied torque cancellation

Using the rotating magnetic field for contactless slipring systems is a very advantageous way of transferring power wirelessly through the air. However, the associated rotational torque may be a concern in such a system, because the contactless slipring is supposed to transfer electrical power across, rather than drive the shaft to rotate mechanically as in the case of an induction motor. A method of cancelling such torque is proposed and investigated in this section. As presented earlier in Section 7.4 the secondary coils are located on the opposite poles of the primary magnetic field and connected in series to form a single secondary winding. Thus, one turn of each coil is considered to investigate the applied torque on each pole as shown in Figure 7-10. The net torque on the secondary side then is analyzed as a result of the torque applied on both secondary coils. It is assumed that the two loops of Figure 7-10 are at some arbitrary angle \( \theta \) with respect to the rotating primary magnetic field \( B_P \) and current \( i \) flowing through them in the shown direction. Because of the toroidal shape of the secondary core, the rotating primary magnetic field enters the secondary toroid from one side and completes its path through the toroid in a circular fashion to exit from the other side as demonstrated in Figure 7-10. In fact because of the toroidal shape, the primary field is uniformly passing through the secondary core as well as the secondary turns. Therefore, each one of the secondary turns can be considered as a current carrying loop placed in a uniform external magnetic field. The direction of the magnetic force applied to these loops is given by the right hand rule and it is pointed towards the center of the loop and given by:

\[
F = i (B_p \times dl) = i r B_p \int_0^{2\pi} \sin \theta d\theta = 0
\]

(7.28)

It can be seen that the net force applied to the secondary side due to the primary magnetic field itself is zero. However, the secondary winding generates its own magnetic field \( B_S \) in a
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way to oppose the primary magnetic field $B_P$. The magnetic field of each loop (circular shape here) can be expressed as:

$$B_i = B_{loop} = \frac{\mu i}{G} = \frac{i \mu_0 \mu_r}{2r}$$ (7.29)

where $\mu_0$ is the permeability of free space and $\mu_r$ is the permeability of the secondary ferrite and $r$ is the radius of each turn of the secondary coil. The interaction between these two magnetic fields produces a torque in the system as given by:

$$\tau_{ind} = k \left( B_p \times B_s \right) = k B_p B_s \sin \delta$$ (7.30)

In (7.30), $k$ is a factor depending on the construction of the secondary side as given by:

$$k = \frac{AG}{\mu}$$ (7.31)

where $A$ is the cross section area of the loop ($= \pi r^2$ for a circular case of this research) and $G$ is a factor that depends on the geometry of the loop. For a circular loop $G=2r$ and for a rectangular loop it is depending on the exact length-to-width ratio of the loop [10]. Since both the magnitude and direction of the applied torque in the system is expressed in (7.30) as cross product and it is directly related to the Sine of the angle between the two magnetic fields; it can be seen from Figure 7-10 that because of the toroidal shape of the secondary layout, the angle between the two magnetic fields is practically zero which means the applied torque is almost zero in the system. This is verified by simulation in Section 7.13.2 of this chapter and also on the practical setup during experiments.

**Figure 7-10:** Secondary winding layout and the angle between $B_p$ and $B_s$. 

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7.9 Full-autonomous poly-phase power converter

Although ZVS operation is achieved by the poly-phase converter presented in Chapter 6 (Section 6.7), it still requires extra control circuitry in order to generate phase delays between the phases. In this section an autonomous poly-phase current-fed push-pull resonant converter with full resonance and ZVS operation is proposed. The proposed converter uses three identical autonomous single-phase converters with symmetrical magnetic coupling to provide equal phase shift for generating a rotating magnetic field. The proposed converter eliminates the detection and control requirements of conventional converters, and can generate accurate phase delays to form a balanced poly phase system with full ZVS operation.

7.9.1 Basic structure

Figure 7-11 (a) shows the general diagram of the proposed poly-phase system. It consists of three identical single-phase current-fed push-pull resonant converters with a common DC power supply. The DC voltage source for each phase is in series with two large DC inductors as compared to the resonant inductor acting as a quasi-current source. The two input inductors of each phase divide the DC current in half under steady state conditions, so that the current flowing into the resonant tank is approximately a square waveform with half the magnitude of the DC current of each phase. In total the converter is configured as six main switches (S1 ~ S6) with common ground for three-phase power conversion. Each phase comprises of a parallel tuned resonant circuit which consists of a primary inductor (L1, L2 and L3) and a tuning capacitor (C1, C2 and C3). The primary coils are magnetically coupled to the secondary power pickups with the inductances Ls1, Ls2 and Ls3, so voltages will be induced in the secondary coils. A secondary power converter is then used to realize circuit tuning, rectification and power regulation according to the load requirements. In Figure 7-11 (a), the magnetic coupling between the primary resonant inductors (k12, k13 and k23) is for the balanced autonomous operation of the converter among the phases, and the magnetic coupling with the secondary side kps is for transferring power to the load. At steady-state, when the secondary pickup coil is fully tuned, the reactive impedance reflected back to the power supply would be very small and can be compensated when designing the primary inductor. Therefore, for the steady-state conditions of the power supply, it is sufficient to simplify the reflected impedance of the secondary pickups as a pure resistive load Req, as shown in Figure 7-11 (b) [160]. Note that as the phases are identical, a complete circuit of the first phase only is shown in Figure 7-11 (b) for analysis.
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In Figure 7-11 (a) it is assumed that the magnetic structure is symmetrical with equal resonant inductors ($L_1=L_2=L_3=L$) and equal mutual inductances ($M_1=M_2=M_3=M$); then, based on the T-equivalent circuit of the two coupled coils, the coupled coils of the full-autonomous converter can be modeled as shown in Figure 7-11 (c). It should be noted that the push-push oscillators have been physically coupled either in a ring or star via additional circuitry [161, 162], while in the proposed full-autonomous topology the individual oscillating circuits are magnetically coupled via their resonant inductances without any direct electrical contacts.

(a) The proposed poly-phase converter with balanced mutual magnetic field coupling ($k_{12}$, $k_{13}$ and $k_{23}$)

(b) Equivalent circuit of one phase
In poly-phase IPT systems, the mutual interaction between the primary phases is a problem, and it is normally cancelled by using phase rotation, flux compensation, etc. as presented in [63, 154]. This is because if the mutual interaction between the phases is not cancelled, it will cause power exchange between phases, which may cause disruption or failure of the power supply. The proposed full-autonomous converter utilizes the mutual magnetic coupling between the phases to form a triple-push coupled ring oscillator for full-autonomous operation. However, the phases should be symmetrically coupled as shown in Figure 7-12 (a). If the magnetic structure of the system is balanced, the required symmetrical magnetic coupling would naturally exist for the converter operation; otherwise, the symmetrical coupling can be achieved by adding extra magnetics in the poly phase circuit as illustrated in Figure 7-12 (b). Note that the coupling between the phases should be sufficiently strong (typically $k_{12} = k_{13} = k_{23} = 0.2 \sim 0.3$) to establish and sustain the balanced phase shift between the phases.
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7.9.2 Basic principle of operation

In principle the proposed full-autonomous poly-phase converter operates based on ring-coupled push-push oscillator’s theory [163-167]. If three identical autonomous single-phase converters are magnetically coupled symmetrically via their resonant inductances, they form a ring-coupled push-push oscillator. N-push oscillator topologies comprise identical oscillators which are symmetrically coupled as a ring to generate high frequency components [162, 167, 168]. In order to utilize N-push circuits for power combination, the oscillating circuits have to present a phase shift distribution of $2\pi/N$ among them. The triple-push oscillator design is a specific case of a ring-coupled oscillators for high frequency generation [161, 169]. The triple-push oscillators have been physically coupled either in a ring or star via additional circuitry [161, 162], while in the proposed topology the individual oscillating circuits are magnetically coupled via their resonant inductances without any direct electrical contacts.

A complete dynamic analysis of such a complex nonlinear system (high order autonomous circuit) is very difficult. The solution may be sensitive to initial conditions and parameter variations which may cause uncertainties such as chaos and bifurcation [48, 170-173]. Due to the symmetry properties of the ring topologies and high order switching operation, they present multiple modes of operation, which may create a chaotic unstable situation [173]. Depending on the initial conditions, the system can enter one or more operating modes [161, 162, 173]. The presence of the different modes depends on the number of oscillators that form the ring and the coupling strength between them [173, 174]. The possible modes of operation in a ring of three coupled oscillators have been identified and reported in [161-163, 173].

(b) Asymmetrical coupling structure

Figure 7-12: Fundamental diagram of a poly-phase full-autonomous converter.
They are listed in Table 7-1, where \( \phi_1, \phi_2, \phi_3 \) are the output phases of the three oscillators. It shows that Mode 1 is the in-phase mode, meaning all the oscillators are in phase, Mode 2 is the mode where the oscillators present a balanced phase shift of 120º, and Mode 3 is when two of the oscillators are in phase, and the third one is in the opposite direction.

### Table 7-1: Possible operating modes in a ring oscillator (N=3)

<table>
<thead>
<tr>
<th>Mode of operation</th>
<th>Oscillation phase (relative to ( \omega_0 t ))</th>
<th>Feature</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mode 1</td>
<td>( \Phi ) ( \Phi ) ( \Phi )</td>
<td>All in phase</td>
</tr>
<tr>
<td>Mode 2</td>
<td>( \Phi ) ( \Phi +120 ) ( \Phi +240 )</td>
<td>Balanced phase shift</td>
</tr>
<tr>
<td>Mode 3</td>
<td>( \Phi ) ( \Phi ) ( \Phi +180 )</td>
<td>Two in phase and one opposite</td>
</tr>
</tbody>
</table>

The mathematical formulation for triple-push oscillators was reported in [175] using three port Z-parameter representation, which showed that three identical oscillators coupled with a common load can generate balanced mode of operation with equal phase shift of 120 degrees. From the theoretical analyses it has confirmed that the circuit would enter the balanced mode (Mode 2 in Table 7-1) of operation instead of the other two modes. This is in consistency with what is observed in this research which demonstrates that coupled three push-pull converters generate balanced three phase currents at the output.

### 7.9.3 Simulation and experimental results on the poly-phase full-autonomous converter

A complete dynamic analysis of such a complex nonlinear system (high order autonomous circuit) is very difficult. The solution may be sensitive to initial conditions and parameter variations which may cause uncertainties such as chaos and bifurcation [48, 170-173]. In this section a simulation study as well as experimental evaluation of the proposed full-autonomous converter is undertaken using the specification of Table 7-2.

As stated in the previous section, the possibility of different modes heavily depends on the coupling strength between the autonomously oscillating phases of the converter. Figure 7-13 shows the resonant currents of the converter for various levels of symmetrical mutual coupling between coupling the phases. As it can be seen from Figure 7-13 (a), for weak coupling about 0.0001 between the phases, the converter evolves to have all the phases in-phase (Mode 1) as the stable mode of operation. For a coupling of 0.001, the converter tends
to have two of the phases in-phase and the other phase is about 180° out of phase (Mode 3) as illustrated in Figure 7-13 (b). As the coupling between the phases increases, the phases tend to push each other within the coupling loop and share an equal phase shift. This can be observed from Figure 7-13(c, d, f) the converter reached the desired mode of operation at a coupling of 0.01 (Figure 7-13 (f)). After the converter reached the proper phase delays between the phases, strengthening the coupling does not affect the mode of operation. Figure 7-14 shows the resonant voltages (full-wave) and the gate waveforms over one side of the resonant voltage for the case with k=0.2. As it can be seen they are good sinusoidal with accurate 120° phase shift and ZVS operation.

Table 7-2: Circuit specifications (full-autonomous poly-phase converter)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1 = L_2 = L_3$ (µH)</td>
<td>100</td>
</tr>
<tr>
<td>$C_1 = C_2 = C_3$ (nF)</td>
<td>220</td>
</tr>
<tr>
<td>$L_{dc1} \sim L_{dc6}$ (mH)</td>
<td>1</td>
</tr>
<tr>
<td>$R$ (Ω)</td>
<td>3</td>
</tr>
<tr>
<td>$V_{dc}$ (V)</td>
<td>12</td>
</tr>
<tr>
<td>$R_1 \sim R_6$ (kΩ)</td>
<td>1</td>
</tr>
<tr>
<td>$C_{S1} \sim C_{S6}$ (nF)</td>
<td>1.5</td>
</tr>
<tr>
<td>N-channel MOSFET (IRFP240)</td>
<td>200V/20A/0.18 Ω</td>
</tr>
</tbody>
</table>

(a) $k = 0.00001$  
(b) $k = 0.001$  
(c) $k = 0.003$  
(d) $k = 0.005$
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Figure 7-13: Resonant currents for various mutual couplings.

(a) Resonant voltages across the resonant tanks (full-wave)
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(b) Gate waveforms following the resonant voltages (half-wave)

Figure 7-14: Simulation results for k = 0.2.

A prototype of the poly-phase converter is built and tested in the laboratory at a frequency of about 73.5 kHz to drive the RMF-based systems. As the structure is symmetrical, the magnetic coupling coefficients between the phases are equal \((k_{12}=k_{13}=k_{23})\) and measured about 0.2. This level of mutual coupling is sufficient to sustain the poly-phase autonomous operation with ZVS. From the experiments, the full-autonomous converter can start up automatically by just turning on the main switch of a regulated DC power supply without employing any start-up gate control. At steady-state on the other hand, the converter maintains self-regulating phase delay scheme with ZVS operation for all phases without any controllers used. Figure 7-15 shows the captured waveforms of the resonant current flowing through the resonant inductors (full-wave) and the resonant voltage across the resonant capacitor (half-wave) at about 73.5 kHz operating resonant frequency. As it can be seen, the waveforms are sinusoidal with a very good quality and exact phase delays. Figure 7-16 shows the measured waveforms of the switching voltage of each phase over switch \(S_1\) (a half wave) and the gate drive signal of the other switch \(V_{G2}\). It can be seen from this result that switching occurs precisely at the zero voltage crossing points following the resonant frequency of its own phase. The phases are adjusting proper 120° phase shift and is flawlessly accepted for poly-phase IPT systems [63, 116].

It should be noted that although the results presented in this section are conducted for a symmetrical system. However, the converter is tested for an asymmetrical system and after balancing the mutual coupling between the phases, it gives similar results with accurate phase 120° delays.
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Figure 7-15: Steady-state measured practical waveforms of the resonant tank at about 73.5 kHz.
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Figure 7-16: ZVS operation of the full-autonomous converter
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7.10 Simulation and experimental results of the poly-phase system

The two single-phase power pickups (Bipolar and Unipolar) presented in Section 7.4 are studied with RMF-D1 and the poly-phase power pickup is studied with RMF-D2. This was based on the structure improvements at different stages of the research.

7.11 RMF-D1

A 3D-FEM model is developed as shown in Figure 7-17 to study the magnetic properties of RMF-D1 using the data shown in Table 7-3.

![Figure 7-17: 3D-FEM model of RMF-D1.](image)

Table 7-3: System specifications (RMF-D1)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_p$ (A)</td>
<td>5</td>
</tr>
<tr>
<td>$N_p$ per phase</td>
<td>34</td>
</tr>
<tr>
<td>$N_s$ per coil</td>
<td>40</td>
</tr>
<tr>
<td>$f$ (kHz)</td>
<td>32</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>5</td>
</tr>
<tr>
<td>Shaft diameter (mm)</td>
<td>14</td>
</tr>
<tr>
<td>Ferrite</td>
<td>Mn_Zn2500</td>
</tr>
</tbody>
</table>

7.11.1.1 Power transfer capability analysis

The power transfer capability of the system can be roughly evaluated by its open-circuit voltage and short-circuit current as demonstrated in Table 7-4. As it can be seen from Table...
7-4, both power pickups (bipolar and unipolar) are able to capture almost same amount of power from the primary magnetic field. The uncompensated power reached to about 157.7 VA and 153.6 VA for the bipolar and unipolar pickups respectively without magnetic shielding. Then for the second stage, an aluminium sleeve is added over the shaft for shielding. Normally, aluminium creates losses in IPT systems and is one of the trade-off designs. For the RMF-based systems however, aluminium shield contributes to the flux linkage between the primary and the secondary sides. The power is increased about 20.35% to 198 VA for the bipolar power pickup, and about 23.9 % to 190.4 VA for the unipolar power pickup. In fact, due to the induced eddy currents in the aluminium, a magnetic flux is produced which opposes the primary flux lines and forcing them back into the secondary ferrite. This allows more flux lines to contribute to the flux linkage between the primary and secondary side and increases the power transfer capability of the system (see Figure 7-18) [116].

Table 7-4: Simulation results of RMF_D1: with single-phase bipolar pickup (left side table) and with single-phase unipolar pickup (right side table)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Without aluminium</th>
<th>With aluminium</th>
<th>Parameter</th>
<th>Without aluminium</th>
<th>With aluminium</th>
</tr>
</thead>
<tbody>
<tr>
<td>V_{OC} (V)</td>
<td>68.27</td>
<td>68.27</td>
<td>V_{OC} (V)</td>
<td>66.8</td>
<td>66.8</td>
</tr>
<tr>
<td>I_{SC} (A)</td>
<td>2.31</td>
<td>2.9</td>
<td>I_{SC} (A)</td>
<td>2.3</td>
<td>2.85</td>
</tr>
<tr>
<td>S_u (VA)</td>
<td>157.7</td>
<td>198</td>
<td>S_u (VA)</td>
<td>153.64</td>
<td>190.4</td>
</tr>
<tr>
<td>Power increase</td>
<td>------</td>
<td>20.35 %</td>
<td>Power increase</td>
<td>------</td>
<td>23.9 %</td>
</tr>
</tbody>
</table>

Figure 7-18: Effect of aluminium shield on the primary flux lines.
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7.11.1.2 Load test

A loading test is also performed by connecting a pure resistive load to the secondary coil without any tuning. The graph of Figure 7-19 (a) shows the simulated output power and efficiency versus the load. The maximum power about 75 W at about 83% efficiency transferred to the load with 30Ω. The maximum simulated efficiency is obtained about 84% for 71.82 W output power and 40Ω load. The power loss for this case is studied based on FEM magnetic simulation and added to Figure 7-19 (a). The major portion of power loss is found to be within the primary and secondary windings (Pw1) as demonstrated in green as compared to the power loss within the ferrite material (Pc1). The percentage of power loss for this case (Pw1 and Pc1) can also be observed from Figure 7-20 as fraction of the total power loss (Pin-Pout) of the system. The subscript ‘1’ here denotes the first case of simulation which is without magnetic shielding added to the system.

Figure 7-19 (b) on the other hand shows the results of the simulated case (second case) with the inclusion of a still shaft and an aluminum sleeve added for magnetic shielding. The power fed to the load is not regulated and depends on the load value (Vo and Io are varying). As compared to the previous case, maximum power transferred to the load is increased to 91.5 W with the same 83% efficiency. However maximum efficiency for this case is slightly increased from 84% to 84.6% at 83.15 W output power. Again for this case the power loss is investigated and added to the graph of Figure 7-19 (b). The power loss within the primary and secondary windings together (Pw2), ferrite material (Pc2) and the aluminum shield (AL2) is simulated in terms of fraction from the total power loss and demonstrated in Fig.9. Similarly the subscript ‘2’ here denotes the second stage of the simulation which includes shielding sleeve over the shaft. It can be seen that again the major portion of power loss for this case is within the coils and it can be decreased by using a wire with greater cross section area and smaller ESR. The power loss within the aluminum is the lowest percentage as the load increases as compared to the core and coils losses. Meaning that adding aluminum material for shielding in RMF-based system improving the power transfer capability of the system, rather than create power loss and appears as a design trade-off.
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![Diagram](image)

(a) Without metal shaft and aluminium sleeve

(b) With steel shaft and aluminium sleeve

Figure 7-19: Simulation results of the RMF_D1 system.
In order to practically evaluate the system, a prototype of the RMF_D1 structure is built and examined as demonstrated in Figure 7-21. The practical structure is driven by a three-phase push-pull converter presented in Section 7.9. The practical model is powered with constant primary current ($I_P=5A$) and 32 kHz. The results of open-circuit voltage and short-circuit current achieved 78V and 2.04A respectively. The un-tuned power as a product of $V_{oc}$ and $I_{sc}$ then is 159.12VA. Another experiment is performed by including a steel shaft and an aluminum shield and is carried out to about 195VA (=78V*2.5A) which is about 22% greater. A very good agreement between the practical and the simulation results can be seen which is indicating the accuracy and feasibility of the system.
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7.12 RMF-D2

The 3D-FEM model shown in Figure 7-22 is developed to study the second (RMF-D2) system. The simulation data are shown in Table 7-5. Similar to RMF_D1, the first stage of simulation is carried out without any shielding considered for the system and the results shown in Table 7-6. The maximum total un-tuned power is reached to about 858 VA. Then for the second stage, an aluminum sleeve is added over the shaft for shielding. In the case of RMF-D2, adding aluminum has also increased the system power transfer capability. This feature of RMF-based systems makes these systems compatible with conductive materials. The power transfer capability is increased to about 15.07% from 858 to 991VA as depicted in Table 7-6. A practical set-up of the proposed system with the same dimensions and simulation data given in Table 7-5 is constructed for system evaluation as shown in Figure 7-23. The experimental results of open-circuit voltage and short-circuit current are shown in Table 7-7. It can be seen that the power is increased to about 14% from 887 to 1014VA after adding the aluminum shield to the system.

![Figure 7-22: 3D-FEM model (RMF_D2).](image)

Based on the theoretical analysis presented in Section 7.6, the open circuit and short-circuit currents are calculated based on the mutual inductance per pole from (7.16) as shown in Table 7-8 using the measured inductances and given circuit parameters (primary current and frequency). Comparing the calculated results with the simulation and practical results shows
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that the introduced factor ‘\(M_p\)’ is reliable and can be used to predict the power transfer capability of a poly-phase system with a very good accuracy without performing live testing. Typical waveforms of the output voltages and currents taken from the practical testing for the case without shielding are shown in Figure 7-24. As it can be seen the output waveforms are reasonably sinusoidal with proper phase shift forming a balanced three-phase output at the secondary side suitable for both single-phase and three-phase electrical loads. A very good agreement can be seen between the achieved results based on three approaches (simulation, experimentation and calculation), signifying the feasibility of the proposed system.

Table 7-5: System specifications (RMF_D2)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>(I_p) (A)</td>
<td>6</td>
</tr>
<tr>
<td>(N_p) per phase</td>
<td>26</td>
</tr>
<tr>
<td>(N_s) per phase</td>
<td>24</td>
</tr>
<tr>
<td>(f) (kHz)</td>
<td>34</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>3</td>
</tr>
<tr>
<td>Shaft diameter (mm)</td>
<td>14</td>
</tr>
<tr>
<td>Ferrite (Mn_Zn)</td>
<td>((\mu=2500, B_s=0.5T))</td>
</tr>
</tbody>
</table>

Figure 7-23: Complete practical set-up of the RMF-D2.
Table 7-6: Simulation results (RMF_D2)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Open circuit test</th>
<th>Parameter</th>
<th>Short circuit test</th>
</tr>
</thead>
<tbody>
<tr>
<td>Without shielding</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{OC1}$ (V)</td>
<td>57.2</td>
<td>$I_{SC1}$ (A)</td>
<td>5</td>
</tr>
<tr>
<td>$V_{OC2}$ (V)</td>
<td>57.2</td>
<td>$I_{SC1}$ (A)</td>
<td>5</td>
</tr>
<tr>
<td>$V_{OC3}$ (V)</td>
<td>57.2</td>
<td>$I_{SC1}$ (A)</td>
<td>5</td>
</tr>
<tr>
<td>Total $S_u$ (VA) = 858</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>With shielding</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{OC1}$ (V)</td>
<td>57.5</td>
<td>$I_{SC1}$ (A)</td>
<td>5.75</td>
</tr>
<tr>
<td>$V_{OC2}$ (V)</td>
<td>57.5</td>
<td>$I_{SC1}$ (A)</td>
<td>5.75</td>
</tr>
<tr>
<td>$V_{OC3}$ (V)</td>
<td>57.5</td>
<td>$I_{SC1}$ (A)</td>
<td>5.75</td>
</tr>
<tr>
<td>Total $S_u$ (VA) = 991.88 (15 % power increase)</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 7-7: Practical results (RMF-D2)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Open circuit test</th>
<th>Parameter</th>
<th>Short circuit test</th>
</tr>
</thead>
<tbody>
<tr>
<td>Without shielding</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{OC1}$ (V)</td>
<td>59.6</td>
<td>$I_{SC1}$ (A)</td>
<td>4.93</td>
</tr>
<tr>
<td>$V_{OC2}$ (V)</td>
<td>60.4</td>
<td>$I_{SC2}$ (A)</td>
<td>4.71</td>
</tr>
<tr>
<td>$V_{OC3}$ (V)</td>
<td>61.2</td>
<td>$I_{SC3}$ (A)</td>
<td>5.05</td>
</tr>
<tr>
<td>Total $S_u$ (VA) = 887.37</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>With shielding</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{OC1}$ (V)</td>
<td>60</td>
<td>$I_{SC1}$ (A)</td>
<td>5.62</td>
</tr>
<tr>
<td>$V_{OC2}$ (V)</td>
<td>60.5</td>
<td>$I_{SC2}$ (A)</td>
<td>5.34</td>
</tr>
<tr>
<td>$V_{OC3}$ (V)</td>
<td>61.5</td>
<td>$I_{SC3}$ (A)</td>
<td>5.75</td>
</tr>
<tr>
<td>Total $S_u$ (VA) = 1013.9 (14 % power increase)</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 7-8: Calculated results based on “Mp” (without shielding)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Open circuit test</th>
<th>Parameter</th>
<th>Short circuit test</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{OC1}$ (V)</td>
<td>62.1</td>
<td>$I_{SC1}$ (A)</td>
<td>4.44</td>
</tr>
<tr>
<td>$V_{OC2}$ (V)</td>
<td>62.6</td>
<td>$I_{SC2}$ (A)</td>
<td>4.31</td>
</tr>
<tr>
<td>$V_{OC3}$ (V)</td>
<td>63.5</td>
<td>$I_{SC3}$ (A)</td>
<td>4.56</td>
</tr>
<tr>
<td>Total $S_u$ (VA) = 835</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
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![Figure 7-24: Measured output results at f ≈ 34 kHz.](image)

**7.13 Comparison and discussion**

**7.13.1 Power transfer capability analysis**

A 3D-FEM model of an existing coaxial single-phase system is developed for comparison with the rotating magnetic field-based system (RMF-D2) as shown in Figure 7-25. The existing single-phase system is modeled with the same overall physical dimensions as that of the RMF_D2 system (OD=95mm, height=25mm, air gap=3mm and shaft diameter=14mm) to analyze the power density of both systems. The primary number of turns \(N_1=20\) and the secondary number of turns \(N_2=35\) are used to fill the entire core window area using the same
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Litz wires as in the RMF_D2 system. However, because of their different layouts, the total amount of ferrite and copper materials are calculated to fill in the given volume for both the designs. It turned out that the proposed system used about 105 cm$^3$ of ferrites in total, being 54.4% more than the existing single phase system (68 cm$^3$). However, its copper use is only about 26 cm$^3$, only half of that used in the single phase system (52 cm$^3$). The comparison between the two systems is conducted based on the same flux density (0.14T) in the air gap, constant primary current of 6A and operating frequency of 34 kHz, and the results are shown in Table 7-9. A prototype of the single-phase system is also constructed and then driven under the same operating conditions as in the simulation study, and the results are shown in Table 7-10.

![Image](image_url)

**Figure 7-25:** 3D model of an existing single-phase system.

**Table 7-9: Simulation results (single-phase system)**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Without shielding</th>
<th>With shielding</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_u$ (VA)</td>
<td>665</td>
<td>656</td>
</tr>
<tr>
<td>Power loss</td>
<td></td>
<td>1.3 %</td>
</tr>
</tbody>
</table>

**Table 7-10: Practical results (single-phase system)**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Without shielding</th>
<th>With shielding</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_u$ (VA)</td>
<td>662</td>
<td>652</td>
</tr>
<tr>
<td>Power loss</td>
<td></td>
<td>1.5 %</td>
</tr>
</tbody>
</table>
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7.13.2 Impact of the contactless slipring system on the shaft

In contactless slipring applications, the speed of the shaft (rotor) is not determined by the contactless slipring system itself, which is designed to transfer electrical power to a rotatable electrical load mounted on the shaft rather than mechanically drive the shaft. A contactless slipring system should be designed as such that it does not affect the shaft rotation. On the other hand, the contactless slipring system should not be affected by the shaft rotation, or simply speaking, it should not exert any force on the shaft by its associated magnetic field. In the case of the existing single-phase system, there is no rotational torque associated as its operation is based on pulsating time-variable type of magnetic field. Thus it exerts radial force on the shaft than rotational force. However, the proposed system is based on rotating field in the air gap, so special care needs to be taken to avoid rotational force on the shaft. In the proposed design, the applied force due to the rotating magnetic field is totally eliminated by using a toroidal core and winding at the secondary side, very different from the traditional winding of an induction machine. Use of toroidal shape provides zero angle between the primary and the secondary magnetic fields, consequently the force on the system shaft becomes zero [116].

![Figure 7-26: Exerted force on the shaft due to the primary magnetic field.](image)
In principle, when there is no current flowing through the secondary coil under open-circuit unloaded conditions, there is no magnetic field generated by the secondary current, so there is no magnetic interaction and force. And when a load current is flowing through the secondary coil, a magnetic field will be generated by the secondary current which will interact with the primary magnetic field, thus a force may be generated. The maximum applied force would occur when the secondary coil is shorted-circuited acting like a shorted cage as in induction motors. Thus, the impact of both existing single-phase and the proposed poly-phase systems on the shaft are studied when the secondary coil is shorted. The results are shown in Figure 7-26 demonstrating that there is no force applied on the shaft.

7.13.3 Load condition analysis

The secondary power pickup is parallel tuned, to boost the voltage and output power under loaded conditions and to study the efficiency of both the existing and the RMF-D2 systems. The power efficiency is calculated based on the defined input and output power (P_in and P_out) as shown in Figure 7-27 for each phase of the poly-phase and the existing single-phase systems.

For a parallel-tuned secondary power pickup shown in Figure 7-27, the resonant voltage and current can be expressed by \[ V_{\text{res}} = Q_s V_{sc} \] \[ (7.32) \]

\[ I_{\text{res}} = I_{sc} \sqrt{1 + Q_s^2} \]

The practical Litz wire used (AWG12) for the secondary pickup coil can handle up to about 10.8 A of current @40°C. Thus, for the \( I_{sc} \) of 5.75 A from Table 7-6 (for the case with shielding), \( Q_s \) in (7.32) should be designed so that \( I_{res} < 10.8 \text{A} \). Since the secondary
inductance $L_s$ and the operating frequency are known circuit parameters, $Q_s$ is then defined by the load. Table 7-11 and Table 7-12 show the simulation and the practical results when the circuit is loaded with a $Q_s$ of 1.5. The power losses shown in Table 7-11 are obtained by FEM simulation, while the power losses shown in Table 7-12 for the practical results are estimated based on their measured Equivalent Series Resistance (ESR) using the following relationships:

\[
P_{\text{loss-total}} = P_{\text{in}} - P_{\text{out}} \quad (7.33)
\]

\[
P_{w1} = I_{1}^2 \cdot R_1 \quad (7.34)
\]

\[
P_{w2} = I_{\text{res}}^2 \cdot R_2 = \left[ I_{sc}^2 (1 + Q_S^2) \right] \cdot R_2 \quad (7.35)
\]

\[
P_{Cs} = I_{\text{res}}^2 \cdot R_{Cs} = \left[ I_{sc}^2 (1 + Q_S^2) \right] \cdot R_{Cs} \quad (7.36)
\]

\[
P_c = P_{\text{loss-total}} - (P_{w1} + P_{w2} + P_{Cs}) \quad (7.37)
\]

where $P_{w1}$ and $P_{w2}$ are the power losses of the primary and the secondary windings, $P_{Cs}$ is the power loss of the secondary tuning capacitor, and $P_c$ is the power loss within the ferrite.

### Table 7-11: FEM simulation results at $Q_s = 1.5$

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Proposed system</th>
<th>Existing system</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{out}$ (W)</td>
<td>1487</td>
<td>980</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>98.28</td>
<td>97.49</td>
</tr>
<tr>
<td>$P_{w1} + P_{w2}$ (W)</td>
<td>18.48</td>
<td>14.7</td>
</tr>
<tr>
<td>$P_{Cs}$ (W)</td>
<td>4</td>
<td>3</td>
</tr>
<tr>
<td>$P_c$ (W)</td>
<td>3.52</td>
<td>7.5</td>
</tr>
<tr>
<td>Total loss (W)</td>
<td>26</td>
<td>25.2</td>
</tr>
</tbody>
</table>

### Table 7-12: Experimental results at $Q_s = 1.5$

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Proposed system</th>
<th>Existing system</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{out}$ (W)</td>
<td>1515</td>
<td>973</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>97.88</td>
<td>96.96</td>
</tr>
<tr>
<td>$P_{w1} + P_{w2}$ (W)</td>
<td>6.5 + 14 = 20.5</td>
<td>7.5 + 8.5 = 16</td>
</tr>
<tr>
<td>$P_{Cs}$ (W)</td>
<td>6.78</td>
<td>5</td>
</tr>
<tr>
<td>$P_c$ (W)</td>
<td>5.52</td>
<td>9.42</td>
</tr>
<tr>
<td>Total loss (W)</td>
<td>32.8</td>
<td>30.42</td>
</tr>
</tbody>
</table>
Chapter 7: Poly-phase contactless slipring system based on rotating magnetic field

From the results shown in Table 7-12 it can be seen that under load conditions (at Qs of 1.5); the total power loss in the proposed system is about 32.8 W which is marginally higher than the power loss of the existing system (30.42 W). But, the proposed system delivers 1515 W of power to the load which is 55% more than that of the existing system (973 W), which leads to a higher overall efficiency. The maximum efficiencies are measured to be about 96.96% for the existing system and slightly greater about 97.88% for the proposed system. Although the total power loss is almost the same for both the systems, the copper loss of the proposed system is about 20.5 W, being higher than 16 W of the existing system due its higher resonant current. The core loss, on the other hand, of the proposed system is about 5.52 W, being lower than 9.42 W of the existing system. These results are in a good agreement with the simulation results shown by Table 7-11.

Both the FEM study and practical testing have demonstrated that the proposed poly-phase contactless slipring system has the following superior features compared to its single-phase counterpart.

1. **Integrated design.** The primary and the secondary sides of the proposed system share a common core allowing the system design to be more compact, which greatly improves the power density of the system. A compact system is often required for rotary applications with a limited space around the shaft.

2. **Higher power transfer capability.** In the proposed system, each secondary phase is coupled to a resultant magnetic field which is 1.5 times that of a single-phase pulsating magnetic field. Thus the voltage induced in each secondary phase is 50% higher. Similar conditions are valid for the other two secondary phases. This has been verified by the simulation and the practical results of both the systems. It has found that the power transfer capability of the proposed system is 1.55 times that of the existing coaxial single-phase system with the same size, corresponding to about 55% increase in power density.

3. **Lower core loss.** The resultant magnetic field of the proposed method is of a special type travelling with a constant magnitude; therefore it does not go into the cycling of the hysteresis loop as in the case of pulsating magnetic field. This would result in lower core losses in the system as demonstrated in Table 7-11 and Table 7-12 from simulation and practical results.
Chapter 7: Poly-phase contactless slipring system based on rotating magnetic field

4. **Self-shielded.** The magnetic field of the proposed system is a resultant vector with a constant magnitude radially pointing towards the secondary (see Figure 7-28a). This spatial magnetic field rotating in the air gap does not link the secondary side through two surfaces of cores as in the case of the existing single-phase system (see Figure 7-28b) [69, 97]. This feature in combination with the toroidal shape of primary and secondary cores confines the magnetic field within the toroidal-shape of cores, thereby results in very low fringing magnetic field associated with the system. This would result in low EMI with little fringing field effect on the adjacent circuitry.

![Figure 7-28: Flux pattern for (a) the proposed system and (b) existing system.](image-url)
Chapter 7: Poly-phase contactless slipring system based on rotating magnetic field

7.14 Summary

A new type of contactless slipring system by generating rotating mmf has been proposed in this chapter to transfer electrical power to a rotating load. Two different primary designs (RMF-D1 and RMF-D2) and three different power pickups designs have been proposed and investigated. The magnetic structure of the system has been designed as such that the net mechanical torque generated on the shaft is zero, so the electric power transfer would not affect the dynamic rotation of the rotor. This has been achieved by proposing a unique toroidal design for the secondary side.

A poly-phase current-fed push-pull resonant converter with ZVS operation is proposed to drive the poly-phase system which operates autonomously without any external controller required. It has been shown that full ZVS for all phases can be achieved with accurate dynamic 120° phase displacements verified by experimenting at about 34 kHz.

The simulation and experimental results have demonstrated that the proposed rotating magnetic field-based system has increased the power transfer capability by about 55% compared to the existing single-phase system with the same overall physical size and operating conditions.
Chapter 8: Conclusions and suggestions for future work

8.1 General conclusions

A comprehensive study on IPT-based contactless slipring systems for wireless power transfer in rotary applications has been undertaken in this thesis by theoretical modelling and analysis, simulation, and experiments. The study covers two main areas: one is magnetic coupling structures, and another is power converters for driving single phase and poly-phase magnetically coupled coils. The proposed slipring systems have been integrated and fully evaluated.

A general introduction to contactless/wireless power transfer technology mainly the contactless slipring system has been presented in the first chapter. This has covered the general structure, general features, suitable applications and major advances and challenges of contactless slipring systems. An overview of the technologies involved in power transfer for rotary applications has been presented in the second chapter of this thesis. Their advantages and disadvantages have been discussed in relation to IPT-based contactless slipring systems. The IPT-based contactless slipring, being the subject of this research, has been reviewed in details, which shows that only IPT-based contactless sliprings can transfer power at kW levels to rotating shafts across large air gaps, although conventional mechanical sliprings are widely used now. Capacitive power transfer is another option for rotary applications, but it is limited to very low power levels across very small air gaps.

An extensive study has been undertaken in Chapter 3 and Chapter 5 on the single-phase contactless sliprings based on pulsating magnetic field principle. The basic theoretical proposal and analysis have been presented in Chapter 3 [57], while detailed simulation and experimental study covered in Chapter 5 [95, 96]. Improved pot-core and through-hole magnetic coupling structures with higher magnetic coupling coefficients have been proposed and verified by 3D finite element simulation as well as practical experiments. All these designs are compared against each other in the next section. In the case of the pot-core type contactless slipring systems, for 5 mm air gap, the power almost doubled for the improved system due to the 20% of improvement in the magnetic coupling coefficient with the same
size. The simulation and experimental results have demonstrated that about 14.5 W and 8.5 W of power delivered to the load for the improved and typical pot-core systems. The power efficiency of the improved system has reached about 89% higher than the efficiency of the typical system with 86%.

In the case of the through-hole systems, it has been demonstrated that the magnetic coupling coefficient of the double-stator system has increased to 0.81 (after the secondary coil is sandwiched between two primary coils) compared to 0.65 for a coaxial and face to face contactless sliprings. It has been found that the double stator system can transfer about 4 times more power than a typical contactless slipring system [96]. The single air gap through-hole systems have demonstrated a very good magnetic field coupling due to their unique geometry with one air gap [92, 97, 100]. The magnetic coupling coefficients are 0.93 and 0.87 for the sandwiched and non-sandwiched structures respectively. Due to the constant coupling between the primary and the secondary coils, in the single air gap systems it has been observed that if the secondary circuit is fully series tuned, the output voltage is approximately constant.

A new method for improving the magnetic field coupling by varying the ratio of copper and ferrite usage of the existing UU, EE and pot-core configurations has been proposed for contactless sliprings. The proposed method is based on shaping the magnetic field distribution by reducing the leakage area between the core limbs to improve the overall magnetic coupling between the primary and secondary coils within the same volume. It has been found from the FEM study and practical experiments that the magnetic coupling coefficient increases about 21.3%, 45% and 20% for UU, EE and pot-core arrangements respectively.

In Chapter 4 an autonomous current-fed push-pull resonant converter with improved gate driving has been proposed which can increase the operating frequency to 10MHz level with full resonance and soft switching operation. The proposed autonomous power converter is built to drive single-phase contactless slipring systems presented in this thesis. The key feature of the proposed improvement is that the converter uses the input DC voltage as the driving power in combination with the resonant voltage as the ZVS signals via two cross-connected diodes. Since the turning ‘on’ of the switches is driven by the DC source, the input capacitors of the switches can be charged up quickly; and the turning ‘off’ is achieved by shorting of the main switches via fast diodes, the discharging process is fast. As a result, the quality of the switching signal can be maintained at a higher frequency without high power
losses associated. Basic design concepts and design equations of the proposed autonomous topology have been presented. The validity of the proposed autonomous converter has been investigated using LTSpice simulations. Two practical example circuits have been analysed and implemented to verify the validity of the proposed converter at 92 kHz and 10 MHz. Both simulation and practical results have demonstrated that the proposed inverter is able to generate high quality resonant currents from kHz to MHz for a wide range of applications such as Inductive Power Transfer (IPT), induction heating, and other applications where high frequency currents or voltages are required. To take the reflected load from the secondary into account, the circuit power efficiency has been studied for a full range of load using LTSpice as well as experiments. It has shown that the maximum circuit efficiency at 92 kHz and 10 MHz is about 93% and 80% respectively.

A new poly-phase contactless slipring system has been proposed in Chapter 6 based on axial travelling magnetic field principle[63]. To drive the proposed travelling magnetic field-based system, a semi-autonomous poly-phase current-fed push-pull resonant converter with ZVS operation led by one of the phases has been proposed. It has been shown that designing one autonomous phase with the lowest resonant frequency as the driving phase and allowing other phases to follow, full ZVS for all phases can be achieved although the following phases may not be in full resonance. Accurate dynamic 120° phase displacements with ZVS operation have been verified by experimenting at about 40 kHz. The simulation and experimental results are in a very good agreement, and they have demonstrated that the proposed converter is able to generate poly-phase high quality currents for poly-phase inductive power transfer systems. In comparison, it is shown that the poly-phase travelling magnetic field-based system is able to transfer about 3.7 times more power than a set of 6 single-phase units.

In Chapter 7 another poly-phase contactless slipring system based on rotating magnetic field principle has been proposed [116]. It helps to increase the power transfer capability within a confined physical space. Two different rotating magnetic field-based primary structures (RMF-D1 and RMF-D2) and three different secondary power pickups have been proposed and investigated. Unlike an induction motor, the rotating magnetic field in this research has been designed for wireless electric power transfer purpose rather than generating a mechanical driving torque on the shaft. Thus, the magnetic structure of the system has been designed as such that the net mechanical torque generated on the shaft is zero, so the electric power transfer would not affect the dynamic rotation of the rotor. A new poly-phase resonant
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converter with full-autonomous operation based on mutual magnetic coupling between the phases has also been proposed in this chapter to drive the poly-phase system. Unlike the semi-autonomous poly-phase converter presented in Chapter 6 the control circuitry is totally eliminated in the full-autonomous converter. So a full autonomous operation with automatic start-up, full resonance at ZVS, and accurate dynamic balancing between the phases are achieved by the converter itself. It has been shown by simulation and experiments that full ZVS at all phases can be achieved with accurate dynamic 120° phase displacements. The system is verified by FEM simulation and practical experiments. It is shown that the RMF-based systems can increase the power density by about 55% compared to the equivalent single-phase pulsating magnetic field-based system.

8.2 Comparisons of different contactless slipring systems

Contactless sliprings can be designed to situate at the end of the shaft using ‘Pot-core’ geometry to form a face to face structure, or be designed to form a ring around the shaft to form a ‘Through-hole’ structure. The through-hole designs can be single-phase (Chapter 3), which operates based on time-variable pulsating magnetic field; or poly-phase which operates based on moving (Axial/Rotating) magnetic field (Chapter 6 and Chapter 7). These contactless slipring systems are compared with each other in this section, and their general features are summarised.

8.2.1 Single-phase pot-core face to face type

Rotating transformers with pot-core geometry are the most common type of contactless slipring systems found in the literature and industry, but these are typically associated with high flux leakage because of the middle limb of the pot-core geometry. So, they are suitable for applications where it is possible to design the system with an air gap smaller than the distance between the middle and the outer limbs. Improvements over existing pot-core design were made by shaping the magnetic field distribution by reducing the leakage area between the core limbs to improve the overall magnetic coupling between the primary and secondary coils within the same volume. The improved design of pot-core allows for a larger air gap design as it maintains about 20% higher magnetic field coupling and results in doubled power transfer capability. However, the improved pot-core design creates large loops of magnetic field lines which may increase the EMI and cause problems for some applications. Table 8-1 compares the magnetic coupling, uncompensated power, as well as the air gap flexibility of different pot-core designs.
8.2.2 Single-phase through-hole type

In the case of through-hole contactless sliprings, coaxial design has been developed by PowerbyProxi Ltd and is currently being trialled for a wind turbine pitch control [26]. In this design, the primary side encircles the secondary side fully, which results in a system with a fixed air gap and asymmetrical primary and secondary. In such a design, in addition to individual design procedures for each side, any changes in the air gap requires a complete system redesign, which can be time consuming and costly. Moreover, if there is a barrier wall along the system shaft, the coaxial structure would not work. The face to face design on the other hand, consists of identical primary and secondary magnetic structures and allows for flexible air gaps. In through-hole face to face design, the secondary can be movable (linearly or/and rotating) giving flexibility, mobility and safeness for the supplied loads. This is one of the clear advantages of this system that accommodates the possibility of linear and rotary movement for the secondary as compared to the coaxial contactless slipring. Moreover, its face to face feature provides the possibility of using such a design with a primary located outside and a secondary in the other side of a barrier wall [95]. However, this is possible if the barrier wall is made of a non-magnetic and non-conductive material.

The through-hole face to face design can be further improved to form a double-stator structure by sandwiching a secondary coil between the existing face to face structures[96]. The two existing structure of a face to face type connected in series with aiding polarities to enhance the magnetic field generation and accordingly enhance the magnetic coupling coefficient. The added coil to the system as a secondary power pickup is without any rotating
core which is a great advantage as a ferrite material may break during rotation. Moreover, this helps to eliminate the magnetic force interaction between permeable ferrite surfaces because all of the ferrite material remains fixed [92, 97]. As a unique feature of the double stator system, the magnetic coupling between the primary and secondary is constant for any angle of rotation. The double stator design is able to transfer higher power as compared with other designs in Table 8-2. The single air gap through-hole systems have high magnetic coupling coefficients due to their unique structures with one air gap. However, the cores needed in these structures are not easy to find in the market so customised design and manufacturing are required.

<table>
<thead>
<tr>
<th>Through-hole systems</th>
<th>Magnetic coupling coefficient</th>
<th>Power transfer capability, Su (VA)</th>
<th>Air gap</th>
<th>General layout</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coaxial design</td>
<td>k₁</td>
<td>S₁≥S₄≈1.45*S₄</td>
<td>Fixed</td>
<td></td>
</tr>
<tr>
<td>Face to Face design</td>
<td>k₂ &gt; k₁</td>
<td>S₂&gt; S₁ &gt; S₄</td>
<td>Flexible</td>
<td></td>
</tr>
<tr>
<td></td>
<td>(k₂≈1.07* k₁)</td>
<td>(S₂≈ 1.12* S₁≈ 1.6* S₄)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Double-Stator design</td>
<td>k₃ &gt; k₂</td>
<td>S₃&gt; S₅&gt; S₂&gt; S₁&gt; S₄</td>
<td>Flexible</td>
<td></td>
</tr>
<tr>
<td></td>
<td>(k₃≈ 1.25* k₂)</td>
<td>(S₃≈ 1.4<em>S₅≈ 4</em>S₂≈ 4.1<em>S₁≈ 6</em>S₄)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Single air gap</td>
<td>k₄ &gt; k₃</td>
<td>S₄</td>
<td>Fixed</td>
<td></td>
</tr>
<tr>
<td>(Non-sandwiched)</td>
<td>(k₄≈1.06* k₃)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Single air gap</td>
<td>k₅ &gt; k₄</td>
<td>S₅&gt; S₂&gt; S₁&gt; S₄</td>
<td>Fixed</td>
<td></td>
</tr>
<tr>
<td>(Sandwiched)</td>
<td>(k₅≈1.08* k₄)</td>
<td>(S₅≈ 2.6<em>S₂≈ 2.9</em>S₁≈ 4.2*S₄)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
Chapter 8: Conclusions and suggestions for future work

8.2.3 Poly-phase axial travelling magnetic field type

The resultant magnetic field of the travelling magnetic field-based system is a special type with a constant magnitude (1.5 times the time-variable type of field) which does not go into the cycling of the hysteresis loop as it is the case for a pulsating magnetic field. This results in a very low core loss in the system and accordingly increased power efficiency. Moreover, because of this higher magnetic flux density, the primary current then can be reduced, and consequently the power loss within the coils ($I^2R$) would reduce.

The system validity has been verified by 3D-FEM simulation and practical experiments and compared with a typical single-phase coaxial system. Basically the travelling magnetic field-based system uses 6 single-phase coaxial structures to form a poly-phase system. It has been found that the travelling field-based system can transfer about 3.7 times more power than a set of 6 individual single-phase systems or about 22 times more than one single-phase structure at much higher efficiency. The maximum power transferred to the load is about 68.4W at 90% efficiency for 6 individual single-phase systems and about 255.6 W at 97% efficiency for the proposed system.

8.2.4 Poly-phase rotating magnetic field type

It has been found from the simulation and experimental results that the rotation feature of the rotating magnetic field offers a better wireless power transfer solution in rotating applications as compared to existing single-phase with coaxial structure. The primary and the secondary sides of the RMF-based system share a common core allowing the system design to be more compact, which greatly improves the power density of the system. A compact system is a good candidate for high power applications with a limited space on the shaft for design. In the RMF-based systems, each secondary phase is coupled to a resultant magnetic field which is 1.5 times that of a single-phase pulsating magnetic field. Thus the voltage induced in each secondary phase is 50% higher. Similar conditions are valid for the other two secondary phases. This has been verified by the simulation and the practical results of both the systems.

It has been found that the power transfer capability of the RMF-based systems is 1.55 times that of the existing coaxial single-phase system with the same size, corresponding to about 55% increase in power density.

The magnetic field of the RMF-based systems is a resultant vector with a constant magnitude radially pointing towards the secondary. This spatial magnetic field rotating in the air gap, does not link the secondary side through two surfaces of cores as in the case of the existing
single-phase coaxial system [69, 97]. This feature in combination with the toroidal shape of primary and secondary cores confines the magnetic field within the toroidal-shape of cores, thereby results in very low fringing magnetic field associated with the system. This would result in low EMI with little fringing field effect on the adjacent circuitry.

8.3 Contribution of the thesis

In this thesis, the contactless slipring systems based on different magnetic field coupling theories have been proposed and investigated using theoretical analysis, computer simulation and practical experiments. The basic characteristics and underlying principles of the systems (magnetic structure and power converter) have been studied in order to determine their performance and power transfer capability. Various improved single-phase magnetic coupling structures are proposed with high coupling coefficients. In addition, this research has looked into the capability of moving magnetic fields in contactless slipring applications. The newly developed magnetic coupling structures from this thesis work are listed below:

- Improved single-phase magnetic coupling structures with:
  1. Single air gaps structures,
  2. Improved pot-core,
  3. Through-hole with face to face layout,
  4. Through-hole double-stator.

- New poly-phase magnetic coupling structures:
  1. A poly-phase system based on axial travelling magnetic field principle with two different secondary power pickups: 1) single-phase and 2) poly-phase.
  2. A poly-phase system based on rotating magnetic field principle with two different primary structures and three different secondary power pickups: 1) single-phase with two-pole, 2) single-phase with one shorted pole, and 3) three-phase with two-pole.

To drive the proposed magnetic structures, one single-phase and two poly-phase power converters have been proposed and built:

1. Improved single-phase autonomous current-fed push-pull converter. The proposed converter can increase the operating frequency to MHz level with full resonance and soft switching operation.
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2. A new poly-phase semi-autonomous current-fed push-pull converter. This converter uses one autonomous phase as the driving phase and allows other phases to follow after predetermined phase delays programmed and controlled by an FPGA-mode of a c-RIO controller and graphical coding achieved by Labview package.

3. A new poly-phase full-autonomous current-fed push-pull converter based on mutual magnetic field coupling without any additional controllers.

The contribution from this thesis has led to the following publications with 10 journal papers, 10 conference papers, and 6 patent applications.

Journal papers:


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Conference papers:


3. **A. Abdolkhani**, and A. P. Hu, "A contactless slipring system by means of axially travelling magnetic field" in *Energy Conversion Congress and Exposition (ECCE)*, 2012 IEEE.


6. **A. Abdolkhani** and A. Sarawde, Paper on: “Effects of STATCOM on Distance Relay”, in *the institution of engineers (Maha Engineers-India)*, Feb 2008.


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Patent applications:

1. PCT Patent Publication No. WO2013/187777; Title: “A MAGNETICALLY PERMEABLE CORE FOR USE IN WIRELESS POWER TRANSFER SYSTEMS”; Applicants: PowerbyProxi Limited and Tyco Electronics Nederland B.V.; Inventors: Saining REN; Lawrence Bernardo Dela CRUZ; Rex Pius HUANG; ALI ABDOLKHANI; Gied HABRAKEN; Wijand Van GILS; Peter Dirk JAEGGER; Guus MERTENS.

2. Five provisional patent applications filed in 2013 and assigned to PowerbyProxi Limited; Inventors: A. Abdolkhani and A. P. Hu

8.4 Suggestions for future work

The work conducted in this thesis has focused on increasing the power density of contactless slipring systems for wireless power transfer in rotary applications, including improving the power delivery of the existing single-phase systems and developing new poly-phase systems without significant increase in size. Various theoretical analyses, computer simulations, and experimental tests have been undertaken to achieve this goal. While some achievements have been successfully accomplished in this thesis, there is still a considerable amount of work for further investigation as described below:

8.4.1 Alternative pot-core coupling configuration

An improved magnetic coupling design for the existing pot-core geometry has been proposed in Chapter 3 and practically verified in Chapter 5 (Section 5.3) of this thesis. The method has improved the magnetic field coupling of the system with the same size and resulted in a system with higher power density. However, as the proposed method modifies the magnetic structure by reducing the outer core limbs (see Figure 8-1a), it increases the air gap and
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thereby the EMI. Alternatively, an investigation can be conducted to modify the middle limb (see Figure 8-1b) of the core and keep the outer limbs the same. This may help to reduce the EMI due to more confined magnetic path.

8.4.2 Single air gap through-hole structure

Two single air gap through-hole contactless sliprings have been proposed in this thesis (Chapter 3-Section 3.4.3). As a result of their unique geometry with single air gap, higher magnetic coupling coefficient (above 0.9) is achieved. In these designs only stationary ferrite core is needed which lead to a fixed mutual inductance between the primary and the secondary sides. A system with fixed mutual inductance results in constant electrical properties which makes it easy to predict the system behaviour with simple control strategy. This topic in relation to single air gap can be further researched on to determine its power transfer performance.

![Pot-core designs](image)

(a)
(b)

Figure 8-1: Pot-core designs (a) the system of this thesis and (b) suggestion for future work.

8.4.3 Frequency stabilisation method for autonomous operation

An improved current-fed autonomous topology power converter has been proposed in Chapter 4 of this thesis which is able to generate high quality sine-wave current at MHz. The proposed converter is a push-pull topology with simple circuitry. It can start-up automatically and sustains full resonance and accurate ZVS operation without using any additional controller. However, it has been observed that the circuit is very sensitive as the operating
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frequency increases (e.g. above 3 MHz). This requires an appropriate method for stabilizing the operating frequency to ensure stable wireless power transfer to secondary pickups. Further study could cover topics such as appropriate resonant tank design, dynamic tuning, etc. to achieve a stable frequency with ZVS operation.

8.4.4 Alternative method for mutual magnetic field coupling cancellation

A poly-phase contactless slipring system is proposed in Chapter 6 of this thesis. The system operates based on axial travelling magnetic field along the shaft (the axis of rotation) and it is able to transfer about 3.7 times more power than the single-phase counterpart. However, because of its geometry, there is an impedance imbalance between the primary phases. If this imbalance is too large, it would affect the power converter performance and may even cause system failure. To minimize this imbalance, a cancellation transformer is used in this research. The extra transformer has reduced the system imbalance to almost zero. However, the cancellation transformer is bulky so it is inconvenient to accommodate in most practical rotary applications. Further research can be conducted to overcome this issue by actively controlling the power converter to handle the system imbalance.

8.4.5 Rotating magnetic field-based system with high number of poles

The contactless slipring system presented in Chapter 7 is based on rotating magnetic field principle. The resultant primary rotating magnetic field of the system has been investigated for 2-pole system in this thesis. In a 2-pole system the magnetic flux lines travel a long path in order to link with the power pickups as shown in Figure 8-2. In principle, the mutual inductance between the two sides is directly proportional to the length of flux path. One way of reducing the magnetic flux path and increase the mutual inductance of the system is to increase the number of poles of the system. A system with higher number of poles has a shorter flux path to travel which increases the mutual coupling as shown in Figure 8-3 for 4 poles. Thus, the system can be further improved with higher number of poles in practical engineering design.
Chapter 8: Conclusions and suggestions for future work

8.4.6 Self-balancing of poly-phase full-autonomous operation

The full-autonomous poly-phase converter proposed in Chapter 7 operates based on triple-push coupled oscillators [172-174]. This converter can eliminate all the control circuitry required by conventional converters and generate accurate 120° phase delays to form a balanced three phase system with ZVS operation autonomously. However, due to the
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Symmetrical properties of the ring topology and associated high order nonlinear switching, it is found that the system may present multiple modes of operation and present a chaotic situation. Depending on the initial conditions and the coupling strength between the oscillators, the system may operate with all three phases in-phase, or two phases in phase and another phase in an opposite direction. Further research is needed to investigate this phenomenon so as to develop new methods for stabilizing the converter at a desirable operation mode for poly-phase wireless power transfer system applications.
References


