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Design of IPT EV Battery Charging Systems for Variable Coupling Applications

Chang-Yu (David) Huang

To my loving family
Abstract

Inductive Power Transfer (IPT) is a method and technology through which power may be transferred between two mutually coupled inductors across a relatively large air gap (from 10mm to 100mm). IPT systems have found applications in industry where galvanic isolation is seen as advantageous. This thesis focuses on developing loosely coupled IPT systems for use in hands-free EV battery charging applications.

An IPT system can be divided into three main independent blocks: the primary power supply, the secondary pick-up controller, and the coupled magnetic structure of both the primary and the secondary. This thesis presents a new lumped coil magnetic structure and two new pick-up controllers. The new lumped coil system, called a Double D charging pad, provides superior magnetic coupling performance over large air gaps when compared to conventional circular charging pads which have been the preferred design for loosely coupled lumped coil systems. A number of practical considerations are addressed and a strategy presented for designing tuning networks within lumped coil systems operating in variable coupling applications. The thesis also develops two new pick-up control topologies, a multi-path pick-up controller, and a circulating current controller, both of which are designed for efficient power regulation.

The Double D charging pad consists of a flux pipe with two flat Archimedean spiral-wound coils which sit in a co-planar relationship on top of the flux pipe. This structural design guides the generated magnetic flux to exit and to enter at the extremities of the pad in order to generate a high flux profile between the primary and the secondary. As a result, the developed Double D pad prototype has an improved coupling factor, 50% better than a circular pad with similar physical size, when operating with large vertical and lateral displacement. The Double D pad structure presented generates magnetic flux only in one side of the spiral winding and is naturally insensitive to nearby metallic objects. This is important for EV battery charging applications where metal on the car is largely ignored.

A number of practical design considerations for variable coupled IPT systems using primary side current control are addressed. The self-inductance of the charging pad can vary with relative movement and misalignment of the ferrimagnetic material in the primary with respect to the secondary coil. As a result, tuning networks become mistuned, introducing additional
reactive load into the system. This impact on the power supply has been investigated and a
design approach has been proposed to mitigate the burden on the power supply.

Two new pick-up controllers, one switched asynchronously and the second one switched
synchronously, have been developed to operate efficiently in variable coupling applications.
They are called the multi-path pick-up controller and the circulating current controller
respectively. The multi-path pick-up controller consists of a series-tuned LC pick-up with
multiple series-parallel LCL networks. Power regulation is performed by slowly switching in
or out the LCL networks so as to adjust the equivalent number of active LCL networks
required to match the output power. The circulating current controller is designed for a
series-parallel LCL tuned pick-up. The controller uses two diodes and two switches,
synchronously switched with the track frequency, to perform both power regulation and
rectification. Both controllers were proven to be efficient when operating at power levels of
1-2kW with variable coupling. The measured efficiency of both controller prototypes is near
95% at rated load and above 85% for part load efficiencies greater than 0.25 per unit.

The thesis demonstrates the feasibility of using fixed frequency IPT systems in a single pick-
up variable coupling application, where in the past this application has been dominated by
variable frequency designs. The techniques developed here can be directly applied to systems
where single power supplies are used to transfer power to multiple vehicles. Using the
developed tuning design methodology, the reactive load in the power supply is minimised to
achieve a nearly identical performance to a variable frequency system. Two new pick-up
control topologies have been introduced that tolerate the variations in coupling. The
development presented in this thesis demonstrates the possibility of using IPT systems for
both stationary and Roadway Powering EVs applications, considered not possible in the past.
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### Nomenclature

#### Acronyms

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<th>Description</th>
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<tbody>
<tr>
<td>AGV</td>
<td>Automated Guided Vehicle</td>
</tr>
<tr>
<td>BEV</td>
<td>Battery Electric Vehicle</td>
</tr>
<tr>
<td>CARB</td>
<td>California Air Resource Board</td>
</tr>
<tr>
<td>CT</td>
<td>Current transformer</td>
</tr>
<tr>
<td>DPF</td>
<td>Displacement Power Factor</td>
</tr>
<tr>
<td>EV</td>
<td>Electric Vehicle</td>
</tr>
<tr>
<td>FCEV</td>
<td>Fuel Cell Electric Vehicle</td>
</tr>
<tr>
<td>HEV</td>
<td>Hybrid-Electric Vehicle</td>
</tr>
<tr>
<td>ICE</td>
<td>Internal Combustion Engine</td>
</tr>
<tr>
<td>ICNIRP</td>
<td>International Commission on Non-Ionizing Radiation Protection</td>
</tr>
<tr>
<td>IPT</td>
<td>Inductive Power Transfer</td>
</tr>
<tr>
<td>LF</td>
<td>Low Frequency</td>
</tr>
<tr>
<td>PF</td>
<td>Power Factor</td>
</tr>
<tr>
<td>RPEV</td>
<td>Roadway Powering Electric Vehicle</td>
</tr>
<tr>
<td>SAE</td>
<td>Society of Automotive Engineers</td>
</tr>
<tr>
<td>UPF</td>
<td>Unity Power Factor</td>
</tr>
<tr>
<td>VLF</td>
<td>Very Low Frequency</td>
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#### Symbol

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>AM</td>
<td>Amplitude modulated</td>
</tr>
<tr>
<td>$C_2$</td>
<td>Secondary tuning capacitor</td>
</tr>
<tr>
<td>$C_{20}$</td>
<td>Nominal pick-up tuning capacitance</td>
</tr>
<tr>
<td>$\Delta C_2$</td>
<td>Variation of pick-up tuning capacitance</td>
</tr>
<tr>
<td>$C_{L1}$</td>
<td>Partial series compensation capacitor of primary track</td>
</tr>
<tr>
<td>Symbol</td>
<td>Definition</td>
</tr>
<tr>
<td>--------</td>
<td>------------</td>
</tr>
<tr>
<td>$C_{L2}$</td>
<td>Partial series compensation capacitor of pick-up coil</td>
</tr>
<tr>
<td>$C_{DC}$</td>
<td>DC capacitor in pick-up filter network</td>
</tr>
<tr>
<td>$D$</td>
<td>Switch PWM duty cycle</td>
</tr>
<tr>
<td>$I_1$</td>
<td>Primary track current</td>
</tr>
<tr>
<td>$I_2$</td>
<td>Secondary pick-up coil current</td>
</tr>
<tr>
<td>$I_{DC}$</td>
<td>Pick-up output DC current</td>
</tr>
<tr>
<td>$I_{sc}$</td>
<td>Pick-up short circuit current</td>
</tr>
<tr>
<td>$k$</td>
<td>Magnetic coupling coefficient</td>
</tr>
<tr>
<td>$\kappa$</td>
<td>Pick-up coupling efficiency metric</td>
</tr>
<tr>
<td>$L_1$</td>
<td>Primary track inductance</td>
</tr>
<tr>
<td>$L_2$</td>
<td>Pick-up coil inductance</td>
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<tr>
<td>$\Delta L_2$</td>
<td>Variation of pick-up inductance</td>
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<tr>
<td>$L_{20}$</td>
<td>Designed nominal pick-up tuning inductance</td>
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<td>Bridge inductance of the power supply LCL resonant network</td>
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<td>$L_{DC}$</td>
<td>Pick-up output DC inductor</td>
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<td>$L_{oc}$</td>
<td>Primary track inductance with pick-up coil open circuited</td>
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<td>Pick-up output AC current go into AC resistive load</td>
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<td>Primary track inductance with pick-up coil short circuited</td>
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<td>$L_t$</td>
<td>Designed track inductance of the primary power supply</td>
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<td>$M$</td>
<td>Mutual inductance</td>
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<tr>
<td>$N_1$</td>
<td>Number of turns in the track inductor</td>
</tr>
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<td>$N_2$</td>
<td>Number of turns in the pick-up coil</td>
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<td>$P_{out}$</td>
<td>Pick-up output power</td>
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<tr>
<td>$pu$</td>
<td>per unit</td>
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<td>$Q_2$</td>
<td>Loaded quality factor of pick-up resonant network</td>
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<td>$Q_{20}$</td>
<td>Nominal pick-up loaded quality factor</td>
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<td>Definition</td>
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<tr>
<td>$Q_I$</td>
<td>Pick-up current quality factor</td>
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<td>$Q_L$</td>
<td>Native quality factor of an inductor</td>
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<td>$Q_V$</td>
<td>Pick-up voltage quality factor</td>
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<td>$R_{AC}$</td>
<td>Pick-up output AC resistance</td>
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<td>$R_{load}$</td>
<td>Equivalent pick-up output resistance</td>
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<td>$RX$</td>
<td>Receiver</td>
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<td>$S_U$</td>
<td>Uncompensated power output of the pick-up coil</td>
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<td>$TX$</td>
<td>Transmitter</td>
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<td>$VA_1$</td>
<td>Primary track volt-amp</td>
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<tr>
<td>$VA_2$</td>
<td>Pick-up coil volt-amp</td>
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<tr>
<td>$V_B$</td>
<td>Inverter bridge output voltage</td>
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<td>$V_d$</td>
<td>Primary inverter bridge input DC voltage</td>
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<td>$V_{DC}$</td>
<td>Pick-up output DC voltage</td>
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<td>$V_{in}$</td>
<td>Input voltage of an LCL network</td>
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<tr>
<td>$V_{in,Avg}$</td>
<td>Average RMS voltage of an AM voltage waveform</td>
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<td>$V_{in,RMS}$</td>
<td>RMS value of an AM $V_{in}$ of it’s RMS envelope</td>
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<tr>
<td>$\hat{V}_{in,RMS}$</td>
<td>Peak RMS voltage of an AM voltage waveform</td>
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<td>$\hat{V}_{in}$</td>
<td>Peak value of the peak RMS voltage of an AM voltage waveform</td>
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<td>$V_{LCL}$</td>
<td>LCL network output AC voltage</td>
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<td>$V_{oc}$</td>
<td>Pick-up open circuit voltage</td>
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<td>$V_R$</td>
<td>Pick-up output AC voltage across AC resistive load</td>
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<td>$V_r$</td>
<td>Pick-up reflected voltage back onto the primary track</td>
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<tr>
<td>$\omega$</td>
<td>Operating frequency of the primary power supply</td>
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<tr>
<td>$\omega_0$</td>
<td>Resonant frequency of the pick-up or track network</td>
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<tr>
<td>$\omega_m$</td>
<td>Mains frequency</td>
</tr>
<tr>
<td>$X$</td>
<td>Characteristic impedance of an LCL network</td>
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<td>Symbol</td>
<td>Description</td>
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<td>$X_{load}$</td>
<td>Equivalent pick-up output reactance</td>
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<td>Mismatch reactance in an LCL network</td>
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<td>Equivalent reactance of pick-up bridge rectifier</td>
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<td>$Z_{in}$</td>
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<td>$Z_{load}$</td>
<td>Equivalent pick-up output impedance</td>
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<td>$Z_{out}$</td>
<td>LCL network impedance seen by the output voltage source</td>
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<td>$Z_r$</td>
<td>Reflected impedance from the pick-up to the track</td>
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Chapter 1.

Introduction

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1.2. Background of Electric Vehicles
1.3. Conductive (SAE J1772) and Inductive (SAE J1773) Charging Standards
1.4. Background to IPT
1.5. Industrial Applications
1.6. Motivations for adopting IPT systems for EV battery charging
1.7. Objectives of the Thesis
1.8. Contributions of the Thesis
1.9. Outline of the Thesis

1.1. Introduction

This thesis is about developing new techniques for a stationary Electric Vehicle (EV) hands-free charging system using loosely coupled coils. Initially, a brief background of EVs is presented in this chapter, followed by a brief review of the current standards for the conventional direct plug-in conductive coupler and a plug-in inductive coupler. The concept of an Inductive Power Transfer (IPT) system is explained and a number of commercially implemented IPT systems are introduced with the focus on EV battery charging applications. This is then followed by a discussion on what is required to further advance the development of IPT systems for use in such applications.

1.2. Background of Electric Vehicles

EVs were first invented in 1834 [1], and were the mainstream of the vehicular market in the late 1800s and early 1900s. Their success was short lived as the Internal Combustion Engine (ICE) became more advanced and popular due to several factors which include cost, high power to weight ratio, and excellent fuel energy density. EVs were soon replaced by ICE vehicles and had almost vanished from the market since the 1930s. However, with the continual decline in petroleum reserves and the increase of public awareness in the environmental impact of fossil fuels, the interest in EVs has rapidly increased in the late 20th century. Since then, the automotive industry has developed new generations of Hybrid and Battery EVs which are increasingly competitive with conventional vehicles and not as dependent on petroleum.
The modern definition of an EV is broad and includes Battery EV (BEV), Hybrid EV (HEV) and Fuel Cell EV (FCEV). In this thesis, the term EV refers to a vehicle powered by an on-board battery that requires external electricity generation for charging the battery, e.g. BEV and plug-in HEV (PHEV). A typical EV built in the early 1900s and a modern EV, Model S, built by Tesla Motors, Inc. are shown in Figure 1-1 (a) and (b) respectively. The modern EVs are more advanced in all aspects, however for modern EVs to meet the demands of current society, they still have challenges to overcome before the uptake of EVs is dominant among the general public. These hurdles include the battery cost and range anxiety of pure battery EVs. Even with state of the art EV battery technology, the energy density of a Lithium ion battery is still 50 times less than that of conventional gasoline (233Wh/kg versus 12.2kWh/kg) [2, 3]. These concerns have led to the idea of increasing the availability of publicly accessible charging points so that opportunistic charging can be used after each short trip rather than a long charge at the end of the day. As a result the driving range can be extended by minimizing the depth of discharge throughout the day and the EV has a lower cost since a smaller battery is required [4].

![Figure 1-1: (a) A typical EV in the early 20th century. (b) A modern EV (Model S) made by Tesla Motor, Inc.]

1 Image from http://edison.rutgers.edu/elecar.htm.
2 Image from http://www.teslamotors.com/models/
The most conventional way of transferring power is by means of a conductive coupler which involves direct metal contact. This requires a plug with a trailing cable to be connected between an EV and its charging point, and thus a comprehensive design of the plug to meet public safety regulations is necessary. The need for a trailing cable to the vehicle not only inherently presents safety and security concerns, but also limits the realisation of opportunistic charging at places like bus stops, taxi stands and traffic lights.

1.3. Conductive (SAE J1772) and Inductive (SAE J1773) Charging Standards

Performing EV battery charging by means of conductive couplers inherently presents a significant safety risk as it involves voltage and power levels that are substantially more dangerous than normal household appliances. To address this concern and also to standardise many different charging mechanism designs, the Society of Automotive Engineers (SAE) has introduced guidelines for both conductive and inductive couplers: these are SAE J1772 and SAE J1773 respectively [5, 6].

1.3.1. SAE J1772: Electric Vehicle and Plug in Hybrid Electric Vehicle Conductive Charge Coupler

SAE J1772 was first issued in 1996 and SAE J1772-2010 was approved on 14th January 2010 by the SAE Motor Vehicle council [6, 7]. This standard covers the general electrical, physical, communication protocol and performance requirements for an EV conductive coupler. Its automotive industry supporters include Chrysler, Ford, General Motors (GM), Honda, Nissan, Tesla Motors and Toyota [7].

In the SAE J1772-2010 standard, the charger plug design, shown in Figure 1-2, is a round connector with 5 pins including two AC lines, one ground pin, one proximity detection and one communication pin [6]. The power flow control is determined by the onboard charger which communicates with the external power supply via the communication pin. The onboard charger and a proximity circuit are used to ensure no voltage is present on exposed terminals while the plug is not in situ. The transfer power level decided in this revision is listed in Table 1-1. The utility infrastructure specified in the standard has the chargers connected to a single phase AC mains supply. Table 1-1 also shows the provision of 3 fast rate DC charging levels specifications using an external battery charger for voltage levels between 200 – 600V DC and power ratings between 19.2kW and 240kW. The finalised DC
levels 1 and 2 for fast rate DC charging specifications are scheduled to be published in December 2011 [8].

<table>
<thead>
<tr>
<th>Charging Level</th>
<th>Voltage</th>
<th>Power Source</th>
<th>Rated Current</th>
<th>Rated Power</th>
<th>Finalised</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC Level 1</td>
<td>120V AC</td>
<td>Single phase</td>
<td>$\leq 16A$</td>
<td>$\leq 1.92kW$</td>
<td>Yes</td>
</tr>
<tr>
<td>AC Level 2</td>
<td>240V AC</td>
<td>Single phase</td>
<td>$\leq 80A$</td>
<td>$\leq 19.2kw$</td>
<td>Yes</td>
</tr>
<tr>
<td>DC Level 1</td>
<td>200 – 450V DC</td>
<td>DC</td>
<td>$\leq 80A$</td>
<td>$\leq 19.2kw$</td>
<td>No</td>
</tr>
<tr>
<td>DC Level 2</td>
<td>200 – 450V DC</td>
<td>DC</td>
<td>$\leq 200A$</td>
<td>$\leq 90kw$</td>
<td>No</td>
</tr>
<tr>
<td>DC Level 3</td>
<td>200 – 600V DC</td>
<td>DC</td>
<td>$\leq 400A$</td>
<td>$\leq 240kw$</td>
<td>No</td>
</tr>
</tbody>
</table>

Table 1-1: Charging power level specified in SAE J1772-2010 [6-8]

With the conductive charging standards becoming affirmed, the deployment of public charging stations continues to grow. As of May 2011, there are more than 1,800 charging stations been installed in the major cities in the U.S. [9]. A plan called “Electric Highways Project” is currently being undertaken in the west coast of the U.S. by the Washington State Department of Transportation to install fast charging stations (AC Level 2) along interstate highways to assist the EVs being able to travel between states [10].

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1.3.2. SAE J1773: Electric Vehicle Inductively Charging

In 1990s General Motors developed an inductively coupled paddle system, shown in Figure 1-3, called MagneCharge for charging their newly developed EV1. Subsequently, it was adopted in the Toyota RAV4 and Nissan Altra EV [11]. The design of the MagneCharge system forms the basis for the SAE J1773 standard [5]. This system consists of a primary converter, a high frequency transformer split between the charging paddle and the charger port, and a diode bridge rectifier in the vehicle [5, 12]. The prime motivations for developing an inductive charging system were safety concerns largely due to the battery pack floating with respect to ground and the risk of damaging litigation should a user be injured. The charging system was designed to provide 3 different levels of charging and these are given in Table 1-2:

<table>
<thead>
<tr>
<th>Charging Level</th>
<th>Voltage</th>
<th>Power Source</th>
<th>Rated Current</th>
<th>Rated Power</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC Level 1</td>
<td>120V AC</td>
<td>Single phase</td>
<td>≤ 12A</td>
<td>≤ 1.44kW</td>
</tr>
<tr>
<td>AC Level 2</td>
<td>208 -240V AC</td>
<td>Single phase</td>
<td>≤ 32A</td>
<td>6.66 to 7.68kw</td>
</tr>
<tr>
<td>AC Level 1</td>
<td>200 – 600V AC</td>
<td>Three phase</td>
<td>≤ 400A</td>
<td>25 to 160kw</td>
</tr>
</tbody>
</table>

Table 1-2: Charging power level specified in SAE J1773-2009 [5, 13]

Although the development of this inductive charging paddle system was proven to be a success for providing safe domestic charging for EVs, it was short lived due to both its cost and that the safety of conductive chargers was simultaneously addressed in the standards. In

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5 Image from: [http://www.promotingevs.com/MagneCharger.html](http://www.promotingevs.com/MagneCharger.html)
June 2001, the California Air Resource Board (CARB) decided to support a single, conductive charging standard SAE J1772-2001 (which at the time was more cost effective) as the charging interface for EVs in California. [14]. This was then followed by General Motors’ decision to suspend further development of inductive charging systems in March 2002 [15].

1.4. Background to IPT

In the past few decades, transferring useful amount of electrical energy over large air gaps through inductively coupled coils has become practically feasible in an efficient and cost effective manner due to advances in materials, power electronics devices, and magnetics. Inductive Power Transfer (IPT) is now widely accepted in industry as a clean, safe and robust method of transferring power across air gaps without electrical contact. IPT has found many applications in various fields including transferring power to an EV battery charger.

1.4.1. Fundamental theory of IPT

IPT is based on Ampère’s and Faraday’s Laws, both proposed in the 19th century. Using the concept diagram shown in Figure 1-4, a current $I$ flows through a wire and a pick-up coil is mounted in close proximity to the current.

According to Ampère’s Law, the current $I$ must produce a magnetic field $H$, some of which will intercept the pick-up coil. Then by Faraday’s Law, shown in equation (1.1), the flux intercepting the pick-up coil will cause a voltage to be induced in the pick-up coil. Conceptually this voltage may then be used to drive loads as required.

$$V = -N \frac{d\Phi}{dt}$$ (1.1)
Thus, using an AC current and a coil, power can be transferred across an air gap by converting power back and forth between the electrical and magnetic forms as described by Ampère’s and Faraday’s Laws. Although these theories have been understood for more than a century, previous applications have been restricted to transformers and induction motors. Since the AC current source and the coil of an IPT system are widely separated by design compared with a traditional transformer, the resultant flux linkage is very poor. As a result, if an IPT system is treated as a standard mains transformer, the power transfer is poor.

To deliver useful power for industrial applications the magnetic field generated by the AC current source needs to be at a much higher frequency than the utility supply of 50-60 Hz [16-18]. As illustrated in (1.1), the induced voltage is directly proportional to $\frac{d\Phi}{dt}$, so if the magnetic coupling between the wire and the coil, shown in Figure 1-4, was lower than a standard transformer by 1000 times, then the IPT operating frequency needs to be 1000 times higher than the utility frequency to get comparable voltage induced in the secondary coil. Usable amounts of power could not therefore be effectively transferred across a large air gap until the availability of modern power electronics technology that could generate power at VLF-LF frequencies. IPT technology requires the use of high power electronic devices operating at these VLF-LF frequencies, Litz cables that minimise eddy current loss for carrying VLF-LF AC currents, magnetic materials such as ferrites, and sophisticated embedded controllers. It is only recently that all of these technologies have been available.

### 1.4.2. Basic structure and operating principle of IPT system

The basic structure of a typical IPT system is shown in Figure 1-5. It comprises two galvanically isolated but magnetically coupled subsystems, which are labelled as the primary VLF-LF switching power supply and the secondary power controller.
The objective of the primary power supply is to convert a rectified mains or DC input into an AC current in the VLF-LF frequency range along an elongated loop or a lumped coil, which is often referenced as the “track” in an IPT system. The VLF-LF AC track current is used to generate and maintain a magnetic field around the track at a desirable level. Due to the mutual magnetic coupling between the primary track and the secondary pick-up coil, an emf is induced across the pick-up coil terminals. This induced emf is then used to deliver power to the load. The power flow required by the load can be controlled from either or both the primary or the secondary side.

### 1.4.3. Features of IPT systems

The structure of the IPT system outlined earlier in Section 1.4.2 offers a number of advantages over traditional conductive coupler systems due to its contactless nature. These advantages include:

- **Electrically isolated and intrinsically safe:** Similar to the traditional transformer, no direct electrical contact is required from the primary side to the load on the secondary side.
- **Physically contactless:** As no electrical contact between the primary and the secondary side is required, IPT becomes attractive for applications where direct electrical contact is either not feasible or not preferred, such as in a biomedical implant devices \[19-21\] or battery charging applications \[22-27\].
- **Robust and durable:** As no direct physical contact between the primary track and the secondary pick-up is needed, the IPT system can be completely sealed and is extremely reliable even in wet or hazardous environments and with little maintenance required.
- **Mobility and rapid movement:** With loose coupling between the primary and the secondary, power transfer can be realised with much greater tolerances to pick-up movement relative to the track compared to a standard transformer. This allows intermittent EV charging to be possible in places where rapid movement is required e.g.: traffic lights, taxi stands and bus stops.
1.5. Industrial Applications

1.5.1. Monorail systems

Daifuku Co., Ltd, based in Japan, is one of the largest manufacturers of material handling systems and conveyors in the world. They have successfully implemented IPT into their material handling systems (Figure 1-6) and their automated clean-room line of products (Figure 1-7). In these systems, the pick-up is designed to move along the track loop while the relative air gap between the pick-up and the track is ideally constant. The contactless nature of IPT means that no spark or debris is produced compared with conventional conductive brushes. This makes the IPT system the preferred option for clean-room factory automation systems. In fact, IPT systems have been widely employed in more than 80% of all clean-room facilities built in the last few decades.

Figure 1-6: Daifuku material handling systems powered by IPT for automotive assembly plant

Figure 1-7: Daifuku clean-room factory automation systems powered by IPT
1.5.2. EV Battery charging applications

1.5.2.1. Santa Barbara electric bus project

In the 1980’s, a study was commissioned to establish the feasibility of a Roadway Powered Electric Vehicle (RPEV) system for Santa Barbara. A battery charging system using IPT techniques for a 35-passenger electric bus was implemented at the Richmond Field Station by the PATH program of the University of California, Berkeley. In this system, the primary track is 1m wide and was constructed using a number of 15mm diameter aluminium cables buried in the roadway surface with steel C-cores to improve magnetic coupling with the pick-up coil. The pick-up inductor was (measured) 4.3m long, 1m wide and installed underneath the vehicle. During operation, the pick-up was mechanically lowered to within 50-80mm of the road surface [28]. A cross-section diagram of the primary track and secondary inductor is demonstrated in Figure 1-8.

![Figure 1-8: Pick-up and track configuration used in the Santa Barbara Electric Bus Project (reproduced from [28])](image)

The primary track current was 2,000A at an adjustable frequency between 180 to 400Hz. In this system, an overall system efficiency of 65% was achieved at an output power of 64kW [28].

1.5.2.2. Whakarewarewa IPT Bus charging project

As part of a redevelopment of the Whakarewarewa geothermal park in New Zealand in 1998, an IPT battery charging system for a 14-passenger electric bus was implemented (shown in Figure 1-9). This system was developed by The University of Auckland in association with Wampfler AG from Germany. The charger was designed to charge the on-board battery of the electric bus, at 20kW with a nominal vertical displacement of 50mm and a horizontal tolerance of ±50mm [22, 29]. The primary track current was designed to be
160A per turn with 5 turns in total at an operating frequency of 12.9 kHz. The charging process was automatically initiated if needed when the vehicles stopped at the loading platform, which was designed to be the charging bay.

![Figure 1-9: Whakarewarewa IPT bus charging project. (a) Electric bus and charging stand, (b) Pick-up produced by Wampfler AG.](image)

1.5.2.3. Commercial systems of IPT battery charging

After the success of implementing the IPT battery charging system at the Whakarewarewa geothermal park, Wampfler AG started commercialising IPT technology into various applications and one of them is the IPT® Charge system (conceptually shown in Figure 1-10 (a)) for charging electric buses. This IPT® Charge system was first installed in Genoa, Italy in 2002 and later in Turin, Italy in 2003.
Figure 1-10: Wampfler IPT® Charge system on electric buses in Genoa, Italy.  (a) Conceptual IPT® system, (b) buried IPT track, (c) IPT pick-up and (d) pick-up lowered into position for charging.

In this design, the primary track coil, shown in Figure 1-10 (b), is buried underground and a circular pad, shown in Figure 1-10 (c), is mounted underneath the bus. This system is designed to use two 30kW circular pad pick-ups to deliver a total power of 60kW. The nominal air gap between the pick-up coil to ground is 30mm with a vertical tolerance of ±10mm and a horizontal tolerance of ±50mm. Similar to the Santa Barbara project mentioned in Section 1.5.2.1, the pick-up coil is lowered by mechanical means to be within the designed vertical displacement during operation.

1.5.3. Automatic Guided Vehicles (AGV)

Wampfler and VAHLE, Inc, from Germany, have both been developing IPT technologies for Automatic Guided Vehicle (AGV) systems, shown in Figure 1-11, used for car assembly plants. The characteristic of these systems is that the pick-up moves along the primary track loop (similar to the monorail systems discussed earlier), but the pick-up of the AGV system may be designed with greater lateral tolerance of movement [30].
Figure 1-11: Wampfler AGV systems at (a) BMW and (b) Audi assembly plant in Germany.

1.6. Motivations for adopting IPT systems for EV battery charging

Although SAE J1772 specifies a comprehensive design with which the conductive coupler is nearly as safe as an inductive coupler during the plug in process, the presence of a physical connection joint between the vehicle and the charging cord still inherently presents security and safety issues for vehicles in public areas. These include:

- Plugs being removed by people with malice [31]
- Plugs and cables being damaged by vandalism.
- Icing at the plugging connection joint in regions where snow is common, such as Canada or Northern Europe.
- Plug requires maintenance due to wear and tear.
- Trailing cables present a tripping hazard in public parking lot.

The physical connection between the vehicle and the charger also limits opportunistic charging at places like taxi stands and traffic lights. Opportunistic charging can help ease driver range anxiety, because the battery energy density is still far less than the energy density of conventional gasoline [2, 3]. By introducing opportunistic charging while the vehicle is in operation, this critical issue of range anxiety can be addressed as required before EVs are widely deployed and accepted.

1.7. Objectives of the Thesis

In the existing IPT battery charging systems outlined previously in Section 1.5.2, both the vertical and horizontal tolerances are very limited, requiring mechanical assistance to
lower the pick-up coil in order to perform charging. This is impractical for opportunistic charging where a good vehicular ground clearance is always required and a decent lateral tolerance is preferable. Keeping this need for tolerance to movement in mind, the goal of this thesis is to develop a stationary IPT EV battery charging system for domestic household and opportunistic charging applications, with a view that the developed system does not compromise the possibility of RPEVs application. This goal is then subdivided into three objectives:

1.) To develop a new magnetic structure that maintains good coupling performance within a given vehicular ground clearance and lateral tolerance without mechanical assistance.

2.) To develop insight into the impact of coupling variation, due to the pick-up misalignment, on the overall system behaviour.

3.) To develop an efficient pick-up controller capable of handling variations in coupling and the resulting fluctuations in power from the primary.

1.8. Contributions of the Thesis

As a consequence of the above considerations, the thesis proposes a new magnetic structure design and two new pick-up controllers for use in EV battery charging applications. This new magnetic structure, called the Double D charging pad, offers the advantages of superior magnetic coupling and efficient power transfer within a given tolerance of movement over existing lumped system designs of equivalent physical size.

The basic pick-up controller categories that are known and in use today are the slow switching and fast switching groups as explained in detail in Chapter 2. Each has their own strengths and weakness dependent on the system design aspects. In this thesis, two new pick-up power flow controllers (one from each category) are proposed and investigated for applications with variable magnetic coupling. They are called the multi-path pick-up (slow switching) and the circulating current controller (fast switching). Both controllers allow efficient power flow control to be implemented.
1.9. Outline of the Thesis

The following list details the focus of each chapter. Chapter 3 presents the magnetic design aspects of an EV battery charging system while Chapter 4 discusses the resonant network design considerations of such a system. Chapter 5 to Chapter 7 present the proposed pick-up controllers with design methodologies and practical performances.

- Chapter 2 is a literature review on current existing IPT system technologies. It covers magnetic design, pick-up controllers, and basic power supply operating principles and it focuses on techniques which contribute to the development of the battery charging system.
- Chapter 3 presents the concept, operating principle and practical measurements with the proposed single sided flux Double D charging pad. Its coupling and efficiency performances are compared against an existing circular charging pad of equivalent physical size at vertical displacements between 110 to 250mm and horizontal tolerance between ±120mm at a height of 200mm.
- Chapter 4 investigates the impact of coupling variations in a lumped coil system on the overall system behaviour. An analysis of the reactive power flow between the pick-up and the primary power supply in a variable coupling system is presented and compared with SPICE simulation results. Recommendations on designing the resonant network to minimise the impact of these variations are presented.
- Chapter 5 presents the concept and design methodology of a slow switching multi-path pick-up controller which uses a series-tuned pick-up and a number of series-parallel tuned LCL networks for power regulation. The design which uses an LCL network with rectifier DC output and variable input voltages is presented and verified with practical measurements. Lastly, the design and measured performance of a 2kW multi-path pick-up prototype is presented.
- Chapter 6 presents the concept and operation of the proposed fast switching circulating current controller. A steady state analysis using fundamental components is presented. This analysis is verified against SPICE simulation results, and practical measurements on a 1.5kW prototype.
- Chapter 7 investigates the performance of the circulating current controller operating in a system with variable coupling between the charging pads. The analysis presented in Chapter 6 is modified to describe the circuit behaviour with variable input voltages. The same 1.5kW prototype presented in Chapter 6 is adopted to be used with variable
input voltages, and practical measurements are presented and compared using SPICE simulations.

- Chapter 8 summarises the main achievements presented in this thesis. Possible directions that may be taken in order to further develop the results are presented and discussed.
Chapter 2.

**Fundamental of Inductive Power Transfer Systems**

2.1. Introduction
2.2. System Overview
2.3. Magnetic Structures
2.4. The IPT Pick-Up
2.5. IPT Power Supplies
2.6. Power flow control topologies
2.7. Conclusion

2.1. Introduction

IPT systems can be broadly grouped into two categories: distributed systems and lumped systems. These two classifications share the same basic structure but the design focuses are different. This chapter contains reviews of the basic IPT systems and their components for both categories. As the purpose of this thesis is to develop new techniques in lumped systems for stationary IPT Electric Vehicle (EV) battery chargers for domestic use, the reviews will emphasize related components for designing such systems at the required power level (2-7kW).

An overview of a typical IPT system is presented first with discussions on the distinctive difference between a conventional lumped system and a distributed system. In the presented overview, the IPT system is broken into three main independent blocks, namely the primary power supply, the secondary pick-up power flow controller and the coupled magnetic structure of both the primary and the secondary. The magnetic designs of the primary and the pick-up are examined followed by the common types of tuning topologies for each of them. Conventionally, in a lumped system for EV charging the power flow is normally controlled on the primary side [12, 32]. However, as one objective of this thesis is to develop new pick-up controllers that are suitable for this application, practical techniques for controlling power flow on both the primary and the secondary sides are examined but with a focus on secondary side control.
2.2. System Overview

The three main components of an IPT system, as shown in Figure 2-1, are the primary power supply, the secondary pick-up power flow controller, and the mutually coupled primary and secondary magnetic structure. Each component can be further broken down into several functional blocks. Although each functional block is designed and operates in conjunction with others, they can be all examined independently.

![Figure 2-1: Fundamental elements of an IPT system](image_url)

The input power source and DC output load of an IPT system will vary based on the available energy sources and the requirements of the application where the system is to be used. Automotive or low voltage electronic systems may run from 12-48V DC inputs, while higher power systems, particularly for EV charging, will run from a single or three-phase AC mains supply. Similar to the primary power supply, the DC output of the pick-up could vary from a low voltage output of 3.6V for charging mobile electronic devices, to the common medium voltage level of 330-560V for industrial use.

The primary power supply can be further divided into two main blocks. Firstly, the inverter, which converts the DC or rectified mains input into an AC source at VLF-LF frequencies. The frequency of an IPT system is typically in the range of 10-150kHz depending on the application [23, 33-37]. Secondly, the resonant/compensation network, which converts the generated AC source into a sinusoidal AC current $I_1$ in the primary inductor $L_1$.

The primary inductor $L_1$, which is often referred to as the track in an IPT application, can be either an elongated loop for a multiple pick-up distributed track system [22, 35, 38-44] or a lumped planar coil for point to point systems [45-50], as shown in Figure 2-2 (a) and (b) respectively. There are a variety of magnetic configurations for the primary inductor, which
are application dependent and are often based upon commercially available ferrimagnetic geometries used for high frequency transformers such as pot cores, U-cores, E-cores and I-cores. These are discussed further in the next section.

![Conceptual diagram of a (a) distributed track system and a (b) lumped coil system](image)

**Figure 2-2: Conceptual diagram of a (a) distributed track system and a (b) lumped coil system**

The pick-up power flow controller consists of a resonant/compensation network and a power conditioner. The resonant network is designed to compensate the inductance of the pick-up coil to boost its output power capacity for practical use. The power conditioner regulates and converts the power from the resonant tank to the DC voltage level required by the load. The pick-up compensation network and controller will be discussed in depth in Section 2.4.

In a distributed system that is normally designed for multiple pick-up applications, power delivered to the load is normally controlled by each individual pick-up’s controller [35]. In this application controlling power from the primary side by means of variable track current or variable operating frequency to meet the load requirement at the output of each pick-up is not feasible as it would affect all of the pick-ups. On the other hand the conventional lumped system is usually designed for a single pick-up point to point application, so the system design focus is different than that of distributed track systems and
power regulation can be done completely on the primary side by modifying the primary current or frequency while keeping the secondary side pick-up as simple as possible.

### 2.3. Magnetic Structures

The magnetic structure of an IPT system has a significant impact on the effectiveness of the power transfer, which is dependent on the mutual coupling between the primary and the secondary. Similar to a conventional transformer, ferrimagnetic material is often used in the design, where appropriate, to provide a low reluctance path for the magnetic flux and hence an improved mutual coupling.

The geometry of the primary magnetic material and the coil layout are designed to generate a desirable pattern of flux distribution to improve the mutual coupling for given restrictions on movement and space. Likewise, the pick-up geometry is designed to intercept the most magnetic flux it can, within a given space constraint, to improve the power transfer. The primary and the secondary magnetic designs of both distributed and lumped systems have been developed into many different structures to suit the needs of particular applications.

The degree of coupling between a primary and a secondary coil may be expressed by the conventional coupling coefficient $k$:

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (2.1)$$

where $L_1$ is the primary self inductance, $L_2$ is the secondary self inductance and $M$ is the mutual inductance between these two coils.

This coupling coefficient is widely used in transformers and induction motor designs. However, in a distributed system a single pick-up only couples to a portion of the primary track length at any given time, so $k$ provides insight only in terms of the system coupling seen by the primary power supply, but gives no information about individual pick-ups [51]. Recently, a new term called $\kappa$ was introduced in [51, 52] and is expressed in (2.2). It is a direct measure of the magnetic performance of a single pick-up coil and is independent of the number of turns on both the primary and secondary. This allows sensible comparisons between various pick-up magnetic geometry designs. In a single pick-up lumped system the secondary pick-up coil normally couples with the full length of the primary coil and under
this circumstance, \( k \approx \kappa \). Therefore, the coupling coefficient \( k \) may still be used to benchmark the performance of a single pick-up magnetic design in a lumped system.

\[
\kappa = \frac{N_2 M}{N_1 L_2}
\]  
(2.2)

For a lumped system, the magnetic coefficient \( k \) is normally greater than 0.2. In contrast, for a distributed system, the \( k \) value for a single pick-up on a primary track inductor is normally between 0.01 and 0.1, but \( \kappa \) is normally in the range of 0.35 to 0.95 depending on the geometry [51].

### 2.3.1. Distributed System

In a distributed system, there are two main track configurations. One is called a uni-polar configuration, where the return wire of a single conductor track is sufficiently distant that no single pick-up couples to both track conductors at the same time [41, 42]. The other one is called bi-polar where the single conductor track and its return wire form an elongated loop and both are coupled to the same pick-up [43, 44, 53].

There are a large number of pick-up magnetic designs as shown in Figure 2-3 which depend on the application requirements, but they can be broadly separated into two groups which are monorail applications (material handling and clean room factory automation) and Automatic Guided Vehicles (AGV). In both applications the pick-ups are designed to move along the length of the primary track. The major difference between them is that the pick-ups used in monorail systems protrude into the track loop restricting sideways motion to essentially zero whereas the pick-ups of AGV systems have a clearance to the track loop allowing a certain degree of tolerance to lateral movement at the cost of lower coupling.

Some common ferrimagnetic geometries are shown in Figure 2-3. The U-core [54-56], E-core [35, 43, 57-62] and H-core, shown in Figure 2-3 (a) to (c) respectively, are the common geometries for monorail applications. Recently, an improved non-standard S-core geometry [39, 63, 64] was proposed and has been proven to provide a much higher coupling performance than U, E and H-core geometries. The flat bar and the flat-E core, shown in Figure 2-3 (e) and (f) are commonly used in AGV applications [30, 65, 66].
2.3.2. Lumped System

In a lumped system, the primary and the secondary windings are two mutually coupled coils which are normally of a similar size and consist of multiple turns. An example of a typical coil layout is planar coils as shown in Figure 2-4, where the coils are axially aligned. Lumped systems are normally designed for fixed point applications where the pick-up can be positioned in close proximity to the primary before power is required. As the lumped system has a higher magnetic coupling coefficient $k$, the required magnetic flux generation in the primary is less for the same power transfer, but the trade-off is that the pick-up movement relative to the primary track is restricted in all directions.

Applications of lumped IPT systems for low power levels are transcutaneous implants [19-21] and low power consumer electronics [34, 67-69]. For medium to high power levels, they are battery charging for electric vehicles [23-27, 36, 70-72], connector replacement [32, 73-79] and rotational joints for industrial and robotic use [80-86]. Typically the primary track is designed to power one single pick-up at a time, but recently for charging very low power
consumer devices the primary track is designed to power multiple pick-up coils simultaneously [87].

An air core lumped system (shown in Figure 2-4) does not have any ferrimagnetic material in either the primary or the secondary coil, so it is difficult to “guide” the generated flux towards the direction of interest (the pick-up coil). Thus, for higher power applications ferrimagnetic materials are often used on either or both the primary and the secondary side to enhance the magnetic coupling and to reduce leakage flux outside the space between them.

The geometry of the ferrite core is application dependant. Several examples of these ferrite cores are shown in Figure 2-5. For transferring power across rotational joints or rotating machines, pot cores, shown in Figure 2-5 (a), are ideal due to their symmetry [80-85, 88, 89]. U-cores in Figure 2-5 (b) [90], pot cores [77, 91-95] and toroidal co-axial transformers (Figure 2-5 (c)) [73, 74, 96, 97], are often used for connector replacements and plug-in battery charging system to provide safety and reliability in wet or hazardous environment.

![Figure 2-5: Ferrite structures used in lumped IPT systems: (a) pot core, (b) U-core, (c) co-axial and (d) inductive charging paddle, joined E-cores with a circular centre post (used in GM EV1)](image)

As mentioned in Chapter 1, an inductive charging paddle system, shown in Figure 2-5 (d), was developed for the GM EV1 electric vehicle [76, 78]. The paddle system consists of a modified E-core with a circular centre post. These closely coupled lumped systems enable small magnetic designs which are suitable for installing inside a vehicle. Such systems must
adhere to the standard published by the Society of Automotive Engineers that provides guidelines when implementing an inductive charging interface for electric vehicles; SAE J-1773 [5].

2.3.3. Hands-free Battery Charging System

Although a lumped system utilising ferrite cores shown in Figure 2-5 provides safety and reliability over conventional metal contact plugs, it requires user intervention to physically plug it into the vehicle to commence charging. This limits the charging to take place only at dedicated charging stations like a garage or a parking lot and hinders possible opportunistic charging at bus stops or traffic lights.

To utilise the advantages of an IPT system to the full extent, the primary coil should be placed on or slightly under the road surface while the pick-up is mounted underneath the vehicle so that no user intervention is required. Both the circular shaped pad [45, 49, 50] and the flat E-core with a planar coil [98, 99], shown conceptually in Figure 2-6, are capable of being used in this application due to their low physical profile. Early developments using circular pads and flat E-cores achieved an air gap separation of 30-60mm for a power transfer of 30kW [22, 99]. As mentioned in Chapter 1, a practical system using circular pads has been implemented by Wampfler IPT® charging electric buses in Genoa and Turin over a nominal air gap of 30mm with ±50mm lateral tolerance [100, 101]. In this system, the pick-up pad in the vehicle needs to be lowered to be within the designed operational air gap before charging can commence.

With renewed interest in EVs, developments using low profile circular lumped systems for power levels in the range of 1-7kW for EVs and 30kW for electric buses are in progress [27, 70, 71, 102]. These developments focus on transferring power across an air gap that meets the vehicular ground clearance requirements, so that the vehicular pad does not need to be lowered in order to commence charging, and can operate over a wide tolerance so that parking does not need to be overly accurate. As this thesis is focusing on developing IPT systems for EV battery charging, the concept of the circular pad will be discussed in depth in Chapter 3 and its performance will be compared with the newly developed Double D lumped system that operates in a similar physical space but with an improved performance.
2.4. The IPT Pick-Up

2.4.1. Coupling model

Although the power transfer in an IPT system occurs in a similar fashion to that of a conventional transformer, the coupling between the primary and the secondary coil is significantly lower, even for a lumped system. Furthermore, the primary power supply is normally designed to be a controllable current source. Therefore, it is no longer convenient to model the system using the conventional equivalent T circuit, shown in Figure 2-7 (a) [56, 91, 103, 104]. Instead, the interactions between the primary and the secondary coil are described using a mutual inductance model, shown in Figure 2-7 (b) [48, 105-107]. The advantage of this model is that the primary and the secondary can be designed largely independently. Note that winding and core losses are not considered in this simplified model.
In the mutual inductance model the primary current $I_1$, the mutual inductance $M$, and the primary operating frequency $\omega$, cause an induced open circuit voltage $V_{oc}$ in the secondary coil given by \[16, 35, 108\]:

$$V_{oc} = j\omega MI_1$$  \hspace{1cm} (2.3)$$

This induced voltage causes a current to flow in the pick-up coil while a load $R_{EQ}$ is connected to the pick-up terminal. When the load is a short circuit ($R_{EQ}=0$), the output current, called the short circuited current $I_{sc}$, is then given by \[16, 35, 108\]:

$$I_{sc} = \frac{j\omega M I_1}{j\omega L_2} = \frac{M}{L_2} I_1$$  \hspace{1cm} (2.4)$$

The apparent volt-ampere output power capacity $S_U$, which is also called the uncompensated power, of the pick-up is given by the product of the open circuit voltage and the short circuit current as given in equation (2-5). This is a useful measure of the effectiveness of the pick-up and may be used to compare one pick-up with another.

$$S_U = |V_{oc} I_{sc}|$$

$$= \omega \frac{M^2}{L_2} I_1^2$$  \hspace{1cm} (2.5)$$
In a similar way to $V_{\text{oc}}$, the reflected voltage $V_r$ induced in the primary coil due to the influence of the secondary current $I_2$ is represented by [16, 29]:

$$V_r = -j\omega M I_2$$

(2.6)

Defining the pick-up input impedance seen by the open circuit voltage $V_{\text{oc}}$ to be $Z_2$, the pick-up coil current $I_2$ is then given by:

$$I_2 = \frac{V_{\text{oc}}}{Z_2} = \frac{j\omega M I_1}{Z_2}$$

(2.7)

Combining (2.6) and (2.7), the equivalent reflected impedance of the pick-up back on to the track is expressed as [45, 107, 109, 110]:

$$Z_r = \frac{\omega^2 M^2}{Z_2}$$

(2.8)

### 2.4.2. Pick-up Compensation Topologies and Power Transfer

#### 2.4.2.1. Series and parallel tuning topologies

By examining the conceptual circuit diagram of the pick-up shown in Figure 2-7 (b), it can be noted that maximum power transfer to the resistive load ($R_{\text{EQ}}$) is realised when $R_{\text{EQ}}$ matches the impedance of the pick-up coil $\omega L_2$. The power transfer to the load under this condition is given by:

$$P_{\text{Max}} = \frac{S_{\text{in}}}{2} = \frac{1}{2} \frac{\omega M^2}{L_2} I_1^2$$

(2.9)

Due to the low coupling nature of the IPT system, this output power is normally insufficient for practical use. To improve the power transfer, the pick-up coil is almost invariably tuned with a capacitor to compensate its inductive impedance. The value of the tuning capacitor is designed to ensure that it is tuned with the pick-up inductance $L_1$ at the primary operating frequency $\omega$, so $\omega_0 = \omega$ where $\omega_0$ is the resonant frequency of the pick-up resonant network. Thus, the value of the secondary tuning capacitor required is:

$$C_2 = \frac{1}{\omega^2 L_2}$$

(2.10)
The two most common tuning topologies are a series tuned network [86, 91, 111] and a parallel tuned network [16, 112, 113], shown in Figure 2-8 and Figure 2-9 respectively.

![Figure 2-8: An ideal series tuned pick-up and its equivalent circuit with a load](image)

For an ideal series tuned pick-up operating at resonance, the combined reactance of $L_2$ and $C_2$ simplify to a short circuit. Thus its equivalent circuit appears to the load $R_{AC}$ as a voltage source equivalent to the open circuit voltage $V_{oc}$. Therefore, the output voltage $V_R$ is fixed by $V_{oc}$ while its output current $I_R$ is determined by $V_{oc}/R_{AC}$.

![Figure 2-9: An ideal parallel tuned pick-up and its equivalent circuit](image)

For an ideal parallel tuned pick-up operating at resonance, a Norton transformation of $V_{oc}$, and the combined reactance of $L_2$ and $C_2$ simplify to an open circuit. Thus, the equivalent circuit appears to the load $R_{AC}$ as a current source equal to the short circuit current $I_{sc}$. Therefore, the output current $I_R$ is fixed by $I_{sc}$ while its output voltage $V_R$ is determined by $I_{sc}R_{AC}$.

A standard measure of a resonant network is the quality factor $Q$, which describes the ratio between the reactive and real power of the resonant circuit [114, 115]. For ideally tuned resonant networks, the pick-up loaded $Q_2$ for series and parallel resonant networks with ideal lossless inductors are defined by:
Adding the compensation capacitor has the effect of boosting the voltage and current in the pick-up coil to increase the power transfer. However, in a standard LC circuit, the load can only take advantage of the increase in either the voltage or the current [116]. This amount of increase is called the total apparent $Q^2$ of the pick-up, given in (2.12). It is defined by the product of the increase in voltage and current seen by the load.

$$Q^2 = Q_v Q_i$$  

(2.12)

Where $Q_v$ is the voltage quality factor and $Q_i$ is the current quality factor.

For a series tuned pick-up, the load sees an increase in current but a constant voltage source equal to $V_{oc}$, so $Q_v = 1$. For a parallel tuned pick-up, the load sees an increased voltage but a constant current source equal to $I_{sc}$, so $Q_i = 1$. This amount of increase is expressed in terms of measurable circuit parameters. When the resonant circuit is ideally tuned, the apparent $Q^2$ is equal to the loaded $Q^2$. Using (2.3) and (2.4), the apparent $Q^2$ of both the series and parallel tuned pick-ups are expressed here:

$$Q^2 = \begin{cases} \frac{I_R}{I_{sc}} = \frac{V_{oc} L_2}{R_{ac} M I_1} = \frac{\omega L_2}{R_{ac}} & \text{series tuned} \\ \frac{V_R}{V_{oc}} = \frac{I_R}{\omega M I_1} = \frac{R_{ac}}{\omega L_2} & \text{parallel tuned} \end{cases}$$  

(2.13)

The drawback of using a tuned pick-up coil is that the tuning must be done carefully to have the desired boosting effect. At the resonant frequency, as the load increases the pick-up $Q^2$ increases and the bandwidth (BW) of the circuit decreases. The output voltage of a parallel-tuned circuit with fixed loading resistance as a function of operating frequency $\omega$ is illustrated in Figure 2-10. The bandwidth of a resonant circuit is defined as the range of frequencies within which the output power capability is greater than or equal to the half of the maximum output power of the circuit and it can be shown that:

$$BW = \frac{\omega_0}{Q}$$  

(2.14)
Operating with an output power that is half of the maximum is impractical for power levels of 2-7kW for charging electric vehicles. Indeed, given the power capacitors will naturally have a tolerance when manufactured, it is practically impossible to achieve perfect tuning for the pick-up resonant network. Thus, this places an effective limit on the value of $Q_2$ to around 10 or below for medium to high power IPT systems [35].

Regardless of the chosen compensation topology, the resulting output power which includes the additional boost in operating current or voltage seen by the load is given by:

$$P_{out} = Q_2 S_U$$

$$= Q_2 \omega \frac{M^2}{L_2} I_1^2$$

This expression indicates that the power transfer to the pick-up can be improved by increasing the loaded $Q_2$, the operating frequency $\omega$, the primary track current, or by improving the magnetic design ($M^2/L_2$ ratio). However, it is difficult in practice for currently available semiconductor switches to operate with the required power level (2-7kW) in the high LF band. Thus, the operating frequency is typically limited to between 10-150kHz [33, 34]. Increasing the track current is possible but this also increases power dissipation in the track, and increases the complexity and cost of the power supply. Increasing $Q_2$ makes the pick-up tuning more sensitive to any variation in component values and operating frequency as outlined earlier. Thus, improving the magnetic design of the IPT system is the most elegant approach to improving the power transfer.

Recently, for parallel tuned networks a partial series compensation capacitor in series with $L_2$ has been suggested to increase the output current capability [116]. This results in an increase in the current quality factor $Q_I$ on top of the conventional voltage boosting factor $Q_V$. 

![Conceptual graph of a parallel tuned circuit resonant voltage over a fixed output resistor as a function of frequency](image)
of the parallel tuned circuit. This partial series-compensated parallel-tuned pick-up is not explicitly discussed further or within the scope of this thesis, although, this technique has been proposed and used in the series-parallel LCL tuned pick-up which is discussed in the next section.

2.4.2.2. Series-parallel LCL network tuning topology

A series-parallel LCL resonant network, originally used as part of the network configurations for the primary power supply [117-121], has also been proposed for the pick-up tuning in monorail applications [116, 122, 123]. This structure, shown in Figure 2-11 (a), is similar to a parallel tuned pick-up but also includes some of the characteristics of a series-tuned network that will be discussed in Section 2.4.3.

The reactance of each branch of the LCL network is designed to be the characteristic impedance X at the designed operating frequency. Under this circumstance X is given by [124]:

$$X = \omega L_2 = \omega L_3 = \frac{1}{\omega C_2}$$

(2.16)

For an ideal LCL pick-up tuned at the operating frequency, using a Norton transformation of V oc as shown in Figure 2-11 (b), the combined reactance of L2 and C2 simplifies to an open circuit similar for the parallel tuned pick-up. Thus, its equivalent circuit appears to the load RAC as a current source equal to the short circuit current Isc in series with the output inductor L3. Therefore, the output current I R is fixed by Isc (V oc/j\omega L2) while the output voltage V R is determined by IscRAC. This demonstrates that the output current is determined by the input voltage and is independent of the load [119]. Since an ideal LCL circuit is a symmetrical network, the pick-up input current I2 is controlled by the output voltage V R.

The circuit loaded Q2 holds the same definition as a parallel tuned pick-up:

$$Q_2 = \frac{R_{AC}}{X} = \frac{R_{AC}}{\omega L_2} = \omega C_2 R_{AC}$$

(2.17)

The derivation of the apparent Q2 of an LCL pick-up is the same as the Q2 of the parallel tuned pick-up. With the output current being fixed by Isc (Q=1), the total apparent Q2
is dependant only on the voltage boosting factor $Q_V$. Using (2.3) and (2.4), total $Q_2$ is given by:

$$Q_2 = \frac{V_R}{V_{\infty}} = \frac{|I_{ac}|}{|V_{\infty}|} = \frac{MI_{ac}R_{AC}}{L_2\omega M I_1} = \frac{R_{AC}}{\omega L_2} X$$

(2.18)

Figure 2-11: (a) A series-parallel LCL tuning topology (b) Norton equivalent circuit of an LCL tuned pick-up

For designs where higher output current is required, a partial series compensation capacitor $C_{L2}$, shown in Figure 2-12, is used to increase the current quality factor $Q_I$ to boost the current [116, 123]. With the introduction of $C_{L2}$, the LCL network characteristic impedance $X$ is designed to be the combined reactance of $L_2$ and $C_{L2}$. The current quality factor $Q_I$ is given by:

$$Q_I = \frac{I_R}{I_{ac}} = \frac{\omega L_2}{X} = \frac{L_2}{L_2 - \frac{1}{\omega^2 C_{L2}}} = \frac{C_{L2}}{C_{L2}} + 1$$

(2.19)

The voltage quality factor $Q_V$ of the partial series compensated circuit remains the same as the original $Q_V$ of the conventional LCL pick-up. As indicated by (2.19), $Q_I$ is determined by the circuit component values and is independent of the load. Therefore, using (2.18) and (2.19), the total loaded $Q_2$ of the LCL pick-up is:

$$Q_2 = Q_I Q_V = \frac{\omega L_2 R_{AC}}{X^2} = \frac{L_2 R_{AC}}{\omega L_2^2}$$

(2.20)
2.4.3. Pick-up equivalent reflected impedance

The equivalent reflected impedances \( (Z_r) \) of all three tuning topologies are given below in terms of their circuit parameters and quality factor \( Q_2 \) \[106, 109\].

\[
Z_r = \begin{cases} 
\frac{\omega^2 M^2}{R_{AC}} = \frac{\omega M^2}{L_2} Q_2 & \text{series tuned} \\
\frac{M^2}{L_2} (R_{AC} - j\omega L_2) = \frac{\omega M^2}{L_2} (Q_2 - j) & \text{parallel tuned} \\
\frac{\omega^2 M^2 R_{AC}}{X^2} = \frac{\omega M^2}{L_2} Q_i Q_j & \text{LCL tuned}
\end{cases}
\] (2.21)

When the primary tuning network is designed with the pick-up inductor in situ but open circuited, the reflected impedance of both the series compensated and the LCL tuned pick-up are similar and only resistive when operating at the designed frequency. On the other hand, the reflected impedance of a parallel tuned pick-up is not purely resistive, but has a known reactive component depending on the coupling. This reflected reactive component is often taken into account during the design and installation of the primary power supply by tuning the primary network with the pick-up inductor in situ but short circuited, to minimise any undesirable reactive load in the power supply inverter bridge. Further discussion on this topic is presented in Section 2.5.2.

2.4.4. Pick-up Rectifier Equivalent Resistance Model

The most common type of load in industrial applications requires a controlled DC output. To deliver a DC load, both a rectifier and filtering network are required to transform the AC power from the resonant network into DC. For a parallel-tuned resonant circuit, a sufficiently large DC inductor is used at the rectifier output to ensure the rectifier operates in continuous conduction. Thus, the rectifier input voltage and current can be assumed to be an
ideal sinusoidal waveform and an ideal square waveform respectively. If only the fundamental component is considered, a “transformer” turns ratio can be used to describe the relationship between the fundamental AC RMS components and the resulting DC output [112]. This allows a simple transformation of the voltages and currents between the DC and the AC sides of the rectifier for simplified analysis. For a parallel-tuned resonant circuit with a rectifier and DC inductor shown in Figure 2-13 (a), this voltage transformer turns ratio \( n_p \) is given by [112]:

\[
 n_p = 1: \frac{2\sqrt{2}}{\pi} \quad (2.22)
\]

A similar approach can be used for a series-tuned and LCL tuned resonant circuit with rectifier and output DC capacitor as shown in Figure 2-13 (b), and its voltage transformer turns ratio \( n_s \) is given by [117]:

\[
 n_s = 1: \frac{\pi}{2\sqrt{2}} \quad (2.23)
\]

![Figure 2-13: The voltage "transformer" turns ratio of (a) Rectifier with output DC inductor and (b) rectifier with output DC capacitor]

2.5. IPT Power Supplies

As outlined previously, an IPT power supply consists of two main blocks; the VLF-LF inverter and the track compensation / resonant network.

2.5.1. Inverter Configurations

The inverter bridge, shown in Figure 2-14, converts a DC or rectified mains input source into an AC output in the VLF-LF range, typically between 10-50kHz for medium to high power levels [23, 35, 36, 120] and up to 150kHz for low power levels [37, 125, 126]. The input to the power converter can be either a voltage-source or a current-source and the converters are fundamentally categorised into two groups based on these two input types [76,
Since only the voltage-sourced inverter topology is used in this thesis, the current-sourced inverter will not be discussed here.

Figure 2-14: Power supply single phase voltage-sourced inverter configuration. (a) full-bridge and (b) half bridge configurations.

A voltage-sourced full-bridge inverter configuration shown in Figure 2-14 (a) uses four reverse conducting switches. For the half-bridge inverter shown in Figure 2-14 (b), one leg of the inverter bridge is replaced with a centre tapped capacitor. This centre tapped capacitor must be sufficiently large that the voltage across it is almost constant under steady state conditions and it thus acts as a constant DC voltage source, with a magnitude of half the input voltage source. Both the full-bridge and half-bridge configurations are controlled by having the switch pairs S1+/− or S2+/− switching in alternating conduction. As a result, the inverter output voltage is approximately rectangular. For a voltage-sourced inverter, care must be taken to make sure one switch is turned off before the other one in the same leg is turned on to prevent a shoot-through condition. As the voltage across the centre tapped capacitor is half of the input voltage, the maximum output voltage of a half-bridge configuration is only +/- Vd/2 instead of +/- Vd for the full-bridge configuration. Thus, the full-bridge configuration is the common topology choice in industry. Therefore, only the operational waveforms and the fundamental magnitude of the full-bridge output voltage are explained further here. The gate drive signals to each leg of the inverter bridge will be at the desired operating frequency with 50% duty cycle as illustrated in Figure 2-15 [128]. The phase shift θ between the transition of the gate drive signals of both pairs of switches determines the fundamental component magnitude of the output voltage Vb. The RMS of this fundamental component V_{B,1} is expressed as [115]:

\[
V_{B,1} = \frac{2\sqrt{2}}{\pi} V_d \sin\left(\frac{\theta}{2}\right)
\]  

(2.24)
2.5.2. Track Compensation Network

Although it is possible to use the inverter bridge to drive a track directly, the track inductor is invariably compensated with capacitors either in series or parallel. This is done not only to minimise the VA rating of the power supply, but also for harmonic filtering purposes. Conventionally, track compensation networks are grouped into two categories, parallel tuned tracks for current-sourced inverters [16, 35, 113, 125], and series tuned tracks for voltage-sourced inverters [22, 67].

In industry, the most commonly used track compensation for voltage-sourced inverters is the series-parallel LCL composite tuned track [117-121, 127, 129]. It has an inductor $L_B$ in series with the inverter bridge output voltage and a parallel tuned track as shown in Figure 2-16. This resonant network configuration for the primary track is also called an impedance converting network.

![Figure 2-15: Phase-shift control of a single-phase voltage-sourced inverter and the resultant square wave output voltage](image)

![Figure 2-16: LCL track compensation network for voltage-sourced inverter](image)
By having the reactance of $L_B$, $L_1$ and $C_1$ the same ($L_1=L_B=L$) and equal to the desired characteristic impedance $X$ at the designed operating frequency $\omega$, the output track current $I_1$ is given by [117]:

$$I_1 = \frac{V_{B,1}}{X}$$ (2.25)

This has the advantage that the track current can be easily controlled by varying the phase shift $\theta$ to control $V_{B,1}$.

The input impedance $Z_{in}$ seen by the inverter bridge voltage $V_{B,1}$ is given by:

$$Z_{in} = \frac{X^2}{R_i + jX_i} = \frac{X^2 R_i}{R_i^2 + X_i^2} - j \frac{X^2 X_i}{R_i^2 + X_i^2}$$ (2.26)

Thus, the LCL network has an impedance converting characteristic, an inductive load $X_1$ in series with the track results in an equivalent capacitive load at the input and vice versa. So the reflected capacitive load from a parallel tuned pick-up will result in an inductive load to the inverter. In practice, the LCL compensation network is normally designed to include any fixed reactive loading from the pick-up to minimise the required VA rating of the power supply and to ensure the switch conduction losses remain low. However, in situations where the pick-up reflects a variable reactive load back on to the track, the primary tuning must be designed to ensure the resultant $Z_{in}$ is only inductive to prevent any undesirable switching losses due to the reverse recovery currents of the switch body diodes [130]. This is examined later in this thesis when a variable reactive load is presented to the track by a pick-up.

### 2.6. Power flow control topologies

As the load on the pick-up varies over time, some form of power regulation is required to maintain a desirable level of pick-up output power and voltage. The power flow in an IPT system can be controlled from either the primary or the secondary side or both. As such, the power control topologies are broadly categorised into these groups.

#### 2.6.1. Primary Side Control Topologies

Generally, primary side control is implemented using two possible methods, operating frequency control or track current control. Operating frequency control varies the frequency of the AC output from the primary inverter bridge to change the overall operating frequency.
of the whole IPT system [125, 126]. Since the pick-up is only tuned at one frequency, the pick-up resonant circuit will vary in tuning, and thus the power transfer is controlled.

As illustrated in (2.15), the pick-up output power is proportional to the square of the primary track current, so the output power can be easily controlled by varying the track current magnitude. One way to control the magnitude of the track current is to vary the input DC voltage to the inverter bridge, but this requires an additional switch mode power supply [131]. For a voltage-sourced inverter, such track current control can be easily realised by varying the phase shift $\theta$ to control the fundamental AC output voltage of the inverter as discussed in Section 2.5 [117, 126, 132]. As the track current is always controlled to be as low as possible while maintaining the required power transfer, the track conduction loss is also minimised.

While these methods ensure that the secondary circuit is as simple as possible, it also means that some form of wireless communication between the primary and the secondary is necessary in order for the primary to perform the control action. Also, the system cannot accommodate more than one pickup at a time since it cannot control the track current or frequency to be correct for two or more different secondary loads at once. As such, primary side control has mainly been implemented in lumped-system battery-charging applications where only one pick-up is active.

Recently, an approach was proposed in [133] using an observer algorithm to monitor the pick-up coupling and output load on the primary side by measuring the reflected voltage. This allows primary side current control to be realised without any feedback from the pick-up. However, this technique relies on the secondary pick-up being tuned such that it reflects only pure resistive loads and fixed reactive loads (for a parallel tuned pick-up) back onto the primary. A hands-free battery charging system, where ferrimagnetic material is often used on both the primary and the secondary magnetic structures, must be designed to accommodate some misalignment between the primary and the secondary and thus this primary side observer technique may be difficult to realise for such applications where both the pick-up and the track cannot always be on tune within the desirable tolerances.

**2.6.2. Secondary Decoupling Control Topologies**

Secondary side decoupling control is the most common control technique used in industrial IPT systems. This technique effectively controls the pick-up quality factor $Q_2$ [16].
The two common switch-mode IPT decoupling controllers are the parallel-tuned boost-mode and series-tuned buck-mode controllers. A load decoupling controller is used for the newly developed LCL tuned pick-up. Both the parallel-tuned boost-mode and series-tuned buck-mode controllers can operate in either a fast switching or a slow switching mode. The output power expressions for both controllers operating in either mode are identical. In this section, the operation of these circuits are mainly explained using a slow switching algorithm. The differences between the slow switching and fast switching controllers are further explained in Section 2.6.5.

2.6.2.1. Boost controller

The boost mode decoupling controller can be used with a series-tuned pick-up [134], but it is uncommon in practice, so the discussion here will focus on the boost controller with a parallel-tuned pick-up.

The basic structure of a parallel-tuned boost controller is shown in Figure 2-17 [16, 35]. As mentioned earlier, the magnitude of the output current of a parallel-tuned network is the short-circuit current $I_{sc}$ of the pick-up coil. This output current is converted into DC through a bridge rectifier. A large DC inductor is used to ensure a continuous current flow through the rectifier so that power can be continuously delivered from the AC side to the DC load and to maintain good efficiency [135]. However, the bridge rectifier is not necessarily zero voltage zero current switched (ZVZCS) – the ideal switching condition for any semiconductor device.

![Figure 2-17: Parallel-tuned pick-up with decoupling boost controller](image)

The simplest control algorithm of this circuit uses hysteresis control to turn on/off the switch $S$ to regulate the output voltage within a pre-defined range of variation, which normally is +/-5% of the nominal output DC voltage. While the switch is off, the rectified resonant output current flows through the DC inductor into the load $R_{DC}$ and the capacitor $C_{DC}$. This results in an increase in the output voltage $V_{DC}$. When the output voltage reaches the pre-set threshold, switch $S$ is turned on. The resonant output current now flows through
the rectifier, LDC and the control switch S, which together appears as a short circuit on the AC side. As the parallel tuned circuit forms a current sourced output, a short circuit collapses its resonance so that QV, and consequently Q2, become 0. Hence, the pick-up is effectively decoupled from the primary track. As the DC blocking diode prevents the DC capacitor being shorted by the control switch, CDC continually discharges through the load. When the set lower threshold of the output voltage is reached, the switch is turned off. The parallel tuned network starts resonating again in order to deliver power and the control cycle repeats. By collapsing the resonance at a duty cycle D, the average value of Q2 is regulated and given by (assuming Q1=1):

\[ Q_2 = (1-D) \frac{\pi}{2\sqrt{2}} \frac{V_{DC}}{|V_\infty|} \]  

(2.27)

The output power of a parallel-tuned boost-mode controller at steady state is then expressed by [115]:

\[ P_{Parallel} = Q_2 \omega \frac{M_2^2}{L_2} I_1^2 \]

\[ = (1-D) \frac{\pi}{2\sqrt{2}} V_{DC} \frac{M I_1}{L_2} \]  

(2.28)

2.6.2.2. Buck controller

The series-tuned buck controller, shown in Figure 2-18, works on a similar principle to the parallel-tuned boost-mode controller; by controlling the average value of the apparent Q2 by effectively decoupling the pick-up from the track [136, 137]. Compared to the parallel-tuned boost controller, the series-tuned buck controller has the advantage that the bridge rectifier is soft switched.

![Figure 2-18: Series-tuned pick-up with decoupling buck controller](image)

As the series-tuned network forms a voltage sourced output, the control switch S is in series with the resonant network and the rectifier. When the switch is turned on, the pick-up
resonant circuit operates with the designed loaded $Q_2$, which depends on the loading resistance $R_{DC}$ with $V_{DC}$ being controlled to be constant. When the switch is turned off, the series-tuned network sees an open-circuit, hence the resonance collapses so $Q_1$ and consequently $Q_2$ become 0. Therefore, the average value of $Q_2$ with controller is given by:

$$Q_2 = D \frac{\pi}{2\sqrt{2}} \frac{I_{DC}}{|I_m|}$$  \hspace{1cm} (2.29)

The average power at steady state is given by [115]:

$$P_{Series} = Q_2 \omega \frac{M^2}{L_2} I_1^2$$

$$= D \frac{\pi}{2\sqrt{2}} I_{DC} \omega M I_1$$  \hspace{1cm} (2.30)

$$= D \frac{\pi}{2\sqrt{2}} \frac{V_{DC}}{R_{DC}} \omega M I_1$$

2.6.2.3. Load Decoupling Switch Control

To regulate the output power of a series-parallel LCL tuned pick-up, a load decoupling switch is used at the output of the rectifier as shown in Figure 2-19 [116, 122, 123]. This structure and its operation is similar to the boost-mode controller of the parallel tuned pick-up. The major difference is that there are two resonant current paths in an LCL network as illustrated in Figure 2-19. With the switch $S$ turned off, the resonant current in the output inductor $L_3$ flows through the bridge rectifier, the DC blocking diode and into the load $R_{DC}$ and the DC capacitor. During this period, the pick-up inductor $L_2$ is resonant with its parallel tuning capacitor $C_2$ to source power from the primary track and to sustain the output resonance in order to deliver power to the load. When the switch $S$ is on, the output inductor $L_3$ is still resonant with $C_2$ but its resonant current circulates through the rectifier and the switch to stop power being transferred. Thus, the load is decoupled from the output resonant network and this causes the output resonant voltage of the LCL network to drop to zero (so $Q_2$ becomes 0). This makes the power flow control easy to implement. Another advantage of this pick-up configuration is that the diode bridge rectifier is zero current (ZC) switched, which is very attractive in high power applications [116, 122, 123].

The controlled average total apparent $Q_2$ is expressed:
\[ Q_2 = (1 - D) \frac{2\sqrt{2}}{\pi} \frac{V_{DC}}{V_{\infty}} \left| \frac{I_{L3}}{I_{\infty}} \right| \]

\[ = (1 - D) \frac{2\sqrt{2}}{\pi} \frac{V_{DC}}{I_{\infty}} X \] (2.31)

The steady state average output power is then given by:

\[ P_{LCL} = Q_2 \omega \frac{M^2}{L_2} I_i^2 \]

\[ = (1 - D) \frac{2\sqrt{2}}{\pi} \frac{V_{DC}}{X} \omega M I_i \] (2.32)

Figure 2-19: An LCL tuned pick-up with load decoupling switch

2.6.3. Dynamic Tuning-Detuning Controller

Another common method of regulating the pick-up output power is to actively tune/detune the pick-up resonant circuit, thereby controlling the available power from the resonant network to the load [138, 139]. If less power is required, the pick-up resonant network is detuned to limit the power transfer. The biggest disadvantage of this topology is that a mistuned pick-up is less efficient than one tuned correctly, and the reflected reactive load from the pick-up to the primary track will increase as the pick-up becomes mistuned. Therefore, this technique is normally only used in low power applications.

2.6.4. AC Processing Controller

A recently developed AC processing controller directly regulates the power in AC form [140-142]. It uses an AC switch formed by a back to back common source/emitter switch configuration. For a parallel-tuned pick-up the AC switch is placed in parallel to the resonant network and the load while for series-tuned pick-up, the AC switch is in series with the resonant network and the load as shown in Figure 2-20 (a) and (b) respectively. The output power is controlled by varying the delay angle of the AC switch which operates synchronously with the primary operating frequency \( \omega \). For the parallel-tuned pick-up, the
AC switch is used to clamp the voltage across the pick-up to zero for a set interval when the switches are on, reducing the overall output voltage. For the series-tuned pick-up, zero current switching is used to regulate the output current. As both control strategies achieve soft switching, they typically operate at high efficiencies, over 90%.

![Figure 2-20: AC processing controller (a) Parallel-tuned, (b) Series-tuned](image)

### 2.6.5. Secondary Side Controller Switching Speed

The switching speed of a pick-up controller is an important design decision and refers to the rate at which the controller switch is activated. It is categorised into two groups, slow switching and fast switching. The characteristic of a slow switching controller is that the pick-up resonant network \( Q_2 \) will oscillate between its maximum value and zero (resonance collapse) to give an average output power at the desired level. A typical slow switching waveform is illustrated in Figure 2-21. In this figure, it can be seen that the resonant voltage collapses to near zero with the gate drive voltage \( V_{switch} \) being held high and the resonant voltage increases while \( V_{switch} \) is low. The slow switching speed is normally around 100-1000Hz.

Fast switching controllers operate with an essentially constant apparent \( Q_2 \) for a given load, as the resonant voltage never collapses while a load is present. This is achieved by switching with a high frequency PWM waveform where the duty cycle is determined by the required output power. Both the boost-mode and the buck-mode controller can operate in fast switching mode while the load decoupling switch controller for the LCL pick-up can only operate with slow switching mode.

The fast switching controller has the advantages that it reduces the output voltage ripple and the transient load presented to the primary but the controller design is more complicated than a slow switching controller [143, 144]. A more in depth discussion of the strengths and weakness of slow and fast switching controllers with regards to battery charging applications is presented in Chapter 5.
2.7. Conclusion

The two main groups of IPT systems are the distributed system and the lumped system. Both groups share the same basic structure consisting of three main blocks; the primary power supply, the secondary pick-up power controller and the coupled magnetic structures of the primary and the secondary. This chapter examined the current state of the art in each of these blocks, with a focus on those elements which could be applied to the development of new techniques for hands-free EV battery charging systems.

A number of magnetic structures for EV battery charging applications were examined followed by an explanation of the primary/secondary tuning topologies. In industry, the most common power supply topology is the voltage-sourced inverter with an LCL compensation network as it is easy to drive, and control of the track current is simple. The two most common primary side control techniques were briefly explained together with their disadvantages for using in a hands-free EV battery charging system. A variety of pick-up tuning and control topologies were discussed with an emphasis on the decoupling controller and the series-parallel LCL tuning topology as its advantages are attractive for high power applications.
Chapter 3.

Introduction to Polarised Single-Sided Flux Lumped-Coil Systems

3.1. Introduction

In an IPT system for stationary EV battery charging applications, the magnetic structure is often a lumped coil system, which may be further categorised into either a closely [76] or loosely coupled system [22, 23, 25]. As presented in the previous chapter, early closely coupled systems for EV battery charging were often of a form that required a user to insert a charging paddle into a slot on the vehicle. A loosely coupled system however, may be installed underneath a vehicle for automatic charging without any user intervention. For a loosely coupled system (hereafter referred to as a pair of charging pads) to be practically operational underneath the vehicle, the operational gap between the pads will need to be within (typically) 150 - 250mm for vehicle clearance, and power levels of 2-7kW are required, for an overnight charge. In addition, such charging pads need to transfer full power with sufficient horizontal tolerance for practical parking constraints. This horizontal tolerance over which full power can be delivered needs to be as large as possible but a range of between ±100 to ±200mm is normally acceptable. The performance of the charging pad’s magnetics is crucial for the stationary IPT EV battery charging system to be practically viable. In this chapter, the performance of a common lumped coil structure, the circular charging pad, is firstly discussed. Secondly, a polarised magnetic structure, named a flat bar charging pad, with an improved coupling performance over a large air gap is proposed to be used for EV battery charging applications [145]. Lastly, based on the advantages of both the circular pad and the flat bar pad, the concept of a novel polarised single-sided magnetic flux structure called a Double D pad [146-148] is presented with practical prototype measurements.
3.2. Coupling Coefficient for Lumped Coil Systems

The analytical expression describing the output power of an IPT pick-up, discussed in Chapter 2, is shown in (3.1). As mentioned in Chapter 2, this equation is expressed in terms of the ratio of the mutual inductance $M$ over the pick-up inductance $L_2$ instead of the conventional magnetic coupling coefficient $k$.

In a distributed track system, a pick-up only couples a fraction of the magnetic flux generated by the primary track. Thus, $L_1$ is relatively large compared to $M$ which results in a very small or nearly zero $k$ [51]. Therefore, it is impractical to use $k$ for analysing the magnetic structure of a distributed IPT system. But the mutual inductance interacts with both the primary and the secondary magnetic structure. This makes a physical understanding of the system difficult due to the inter-relationships between parameters.

$$P_{\text{out}} = \omega \frac{M^2}{L_2} I_1^2 Q_2$$  \hspace{1cm} (3.1)

Here $Q_2$ is the loaded quality factor of the pick-up coil during operation.

However in a lumped coil system as used in battery charging applications the mutual inductance between a single pick-up and the primary track is no longer small compared to the primary inductance so the coupling coefficient $k$ now has a sensible value and is a useful measure for describing the magnetic performance of an IPT system. Thus it is helpful to rearrange the equation to be in terms of the conventional coupling coefficient $k$ and primary parameters for ease of comparison. Using the definitions of $M$, and coupling factor $k$, (3.1) now becomes:

$$P_{\text{out}} = \omega \frac{M^2}{L_1 L_2} L_1 I_1^2 Q_2 = k^2 V_A I_1 Q_2 = k^2 Q_2$$  \hspace{1cm} (3.2)

From (3.2), it can be seen that the coupled power to the pick-up pad is dependent on the input VA of the primary charging pad ($V_A$), the magnetic coupling factor $k$ and the loaded quality factor $Q_2$ of the pick-up resonant network. Increasing either the VA of the primary pad or the quality factor of the secondary resonant network will increase the standing losses in the charging pads. Increasing the coupling coefficient $k$ is an elegant solution to improve power transfer as it involves improving the magnetic circuit. Thus as in
transformers, the coupling coefficient $k$ provides a useful measure for directly comparing magnetic structures of different charging pad designs. It is desirable to have a high coupling coefficient in the range of 0.5 to 0.15 within the desired power transfer zone which is normally with an air gap of 150 – 250mm and ±100 to 200mm lateral tolerance, to ensure the overall feasibility and efficiency of the complete system.

It should be noted that the uncompensated VA output of the coupled secondary pad (VA$_2$) is directly related to the input VA of the primary pad ($kV_1$)*($kI_1$) = $V_{oc}I_{sc}$. If a different $V_1$ to $I_1$ ratio is needed, this can simply be done by changing the number of turns in the coils to achieve the desirable primary inductance. Thus a different ratio of $V_{oc}$ and $I_{sc}$ can be achieved by altering the turns ratio on the secondary relative to the primary. Changing both the number of turns and the turns ratio in the coil has a minimal effect on the coupled magnetic structure of the lumped coil system as long as the coil coverage area is maintained relatively constant [70, 71].

3.3. Characteristics of Circular Charging Pads with Aluminium Casing

3.3.1. Flux Path and Coupling Limits of Circular Charging Pads

As outlined in Chapter 2, the most common lumped coil design for EV battery charging applications is the circular charging pad design, shown conceptually in Figure 3-1. This design is derived from traditional pot cores and is well researched and optimised [49, 50, 70, 71, 149]. The design comprises an aluminium ring and backing plate containing twisted multiple strand windings or spiral windings on top of either a shaped solid ferrite disc or distributed ferrite strips as shown here. Ferrite strips are lighter, lower cost and less likely to fracture so they are preferred. The aluminium case design is used to limit the leakage magnetic field outside the circular charging pad cylindrical surface area as described in [70], although if field leakage is not as stringent the aluminium ring is not required. The aluminium back plate however is still essential to avoid nearby materials being heated (increasing loss) or if ferrimagnetic material, in nature, affects the inductances, and tuning of the pad as discussed further in Section 3.3.2.
The magnetic flux, generated by a track current, in the ferrite has flux lines that start on one end of the ferrite bar and propagate to the other end of the bar in a path containing the coil that may be thought of as a semi-elliptical shape. The flux lines leave the ferrite at right angles and propagate to the other end of the bar, entering it at right angles. This flux line profile is shown conceptually in Figure 3-1 while a simulated 2-D cross-section flux line profile using a Finite Element Method (FEM) modelling package JMAG is shown in Figure 3-2. Here the large contours correspond to low flux density (low power) areas where as the inner semi-circular contours are where sensible power transfer can occur.

Figure 3-2: A 2-D JMAG simulation of a circular charging pad*

The height of the flux path determines the coupling between the charging pads. In this 2-D cross-sectional diagram it can be seen that the height of the strong flux lines is roughly

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* The 2-D JMAG simulation plot was kindly provided by Mr. Mickel Budhia
proportional to \( \frac{1}{2} \) of the length of the ferrite bar and that is roughly \( \frac{1}{4} \) of a circular charging pad diameter. Thus the practical height of the flux path for power transfer application is low relative to the size of the charging pad. The parameters, practical inductance and coupling measurements of a 700mm diameter circular charging pad, constructed according to [71] and shown in Figure 3-3, are given in Table 3-1 and Figure 3-4 respectively. The charging pad inductance measurements in this chapter are collected by using an Agilent LCR meter Model: E4980A. The primary pad inductance is measured with the secondary pad open-circuited and short-circuited, and this process is repeated for the secondary pad. Using these measurements the mutual inductance between the charging pads, and hence the coupling coefficient \( k \), can easily be calculated [150].

![Figure 3-3: Photo of a constructed 700mm diameter circular charging pad.](image)

<table>
<thead>
<tr>
<th>A: Ferrite width</th>
<th>28mm</th>
<th>G: Pad thickness</th>
<th>30mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>B: Ferrite length</td>
<td>279mm</td>
<td>Coil</td>
<td>12 turns</td>
</tr>
<tr>
<td>C: Ferrite thickness</td>
<td>16mm</td>
<td>Wire diameter</td>
<td>4mm</td>
</tr>
<tr>
<td>D: Inner coil diameter</td>
<td>320mm</td>
<td>Number of strands</td>
<td>2</td>
</tr>
<tr>
<td>E: Outer coil diameter</td>
<td>520mm</td>
<td>Number of Ferrite Strips</td>
<td>12</td>
</tr>
<tr>
<td>F: Pad diameter</td>
<td>700mm</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 3-1: Parameters of a 700mm circular charging pad
As discussed earlier, for a practical EV battery charging system the desirable coupling coefficient $k$ is between 0.5 to 0.15 for an air gap between 150 to 250mm respectively. With the 700mm circular pad performance shown in Figure 3-4, the constructed charging pad has a $k$ of 0.2 at a vertical spacing offset of 175mm. This allows the IPT system to deliver 2 kW to the pick-up with a track current of 30A RMS at 20 kHz with a pick-up resonant network $Q_2$ of 3.33 - which is an excellent practical value. However, a $k$ of 0.2 at 175mm air gap is relatively low, and it is preferable that such $k$ could be achieved at 200mm air gap. Thus, to improve the coupling of the circular pad at the desirable air gap of 200mm the size of the pad will necessarily have to increase in proportion to the desired improvement in height [70, 71]. In this case, in order to achieve a $k$ of 0.2 at an air gap of 200mm instead of 175mm, the circular charging pad diameter will need to increase by roughly 100mm. This increase is 4
times that of the increase in air gap. As mentioned earlier, this pad design with its winding has already been computer optimised so these figures for a circular pad with the aluminium shielding as designed are practically the best possible [71]. Such large increases in diameter make the system impractical with large air gaps and high power levels where shorter charge times are highly desired. If a 10kW pad is required for a \( k \) of 0.3 at 200mm height the circular pad will be very large. In practice significantly higher coupling coefficients are required than can be practically achieved with the circular pad topology to ensure such an EV charging system remains efficient and cost effective.

### 3.3.2. Single Sided Flux Generation of Circular Charging Pads

Despite the practical limitations on its physical size, the circular pad structure has a very desirable characteristic that it only generates magnetic flux in the direction of the pick-up pad. This is because the coils sit above strips of ferrite, or a ferrite disc, that channels most of the flux through the ferrite under the coil. This leaves only a tiny amount of leakage flux underneath or out the sides of the ferrite. The pad efficiency can be quantified easily by its native quality factor, \( Q_L \). By definition the \( Q_L \) of an inductor is the ratio of its impedance to its AC resistance at the operating frequency \( (Q_L = \omega L / R_{AC}) \). This \( Q_L \) is a measure of the ratio of the peak stored (resonant) energy in the inductor to the standing loss in the inductor itself per radian. With the majority of the generated flux in the cylindrical space between the charging pads, the native quality factor \( Q_L \) of the circular charging pad is hardly affected when a shielding aluminium plate is added to the back and to the side of the pad to prevent spurious eddy current losses in surrounding metallic materials. This is especially important for an EV battery charging system as the floor pan of the EV is typically made out of steel. Without this shielding mechanism, eddy current losses in nearby metallic objects due to leakage flux could be problematic.

The measured \( Q_L \) of the 700mm charging pad using the LCR meter is \( \sim 220 \) at 20kHz. The measured inductance of the circular charging pad at the desired air gap 200mm (which is midway between 150 – 250mm) is \( \sim 132 \mu \text{H} \). With a primary power supply generating a track current \( I_1 \) of 30A RMS at 20kHz, the total standing loss (which includes the copper loss, the iron loss and the eddy current loss in the aluminium) in the charging pad is 68 W which is only 3.4% of the desired 2 kW output.
3.4. Development of a flat bar charging pad

As discussed in Section 3.3.1, the height of majority of the flux generated is proportional to half of the length of the ferrite bars underneath the energised coil. Ideally it is therefore possible to improve the coupling for a given pad size if a pad can be designed such that the flux lines exit and enter the pad at the extremes, using the whole length of the pad as shown conceptually in Figure 3-5. Given such flux paths are roughly semi-circular, this should approximately double the height at which a secondary pad can couple power over the circular design. With this preferred geometrical characteristic, a conceptual system, called a “flat bar” charging pad is examined, where both the primary and the secondary charging pad’ structures are identical.

![Figure 3-5: Ideal arrangement of coils and ferrite for generating flux lines enter and exit at the extremes of the pad](image)

3.4.1. Structure of a flat bar type geometry charging pad

The conceptual structure of the proposed flat bar charging pad is shown in Figure 3-6 [145]. The structure is composed of two main elements to produce a much enhanced magnetic coupling performance. These are the two separated flux transmitter/receiver pole areas, and a “flux pipe” that is used to connect the flux transceivers to each other [145].
Figure 3-6: Conceptual diagram of flat bar type charging pad

The flux pipe has a core of ferrite and a continuous winding to generate magnetic flux in the core and to channel it from one flux transceiver to the other. A back-plate of aluminium is used to prevent any flux leaking from the core. Above the core there is a separate aluminium plate to further contain the magnetic flux being channelled through the flux pipe, and not leaking through the winding coils. This plate cannot be electrically connected to the backing plate or the combination would constitute a short circuited turn. With this structure, the path of the generated magnetic flux in the ferrite core has the near full length of the charging pad. The flux paths from a flat bar type charging pad are shown conceptually in Figure 3-7 and a 2-D flux line profile simulated using JMAG is shown in Figure 3-8. As shown in the 2-D simulation, the flux line profile is more evenly distributed towards a higher height compared with the circular pad flux profile shown earlier in Figure 3-2.

As before the flux lines are approximately semi-elliptical / semi-circular but the height of the generated flux profile is a lot higher than a circular pad for the equivalent physical size. Therefore by comparison to a circular pad this charging pad magnetic structure can operate over a much larger separation. One characteristic of the flat bar charging pad is that the magnetic structure is polarised. This means that if the pick-up pad is rotated 90 degrees relative to the primary pad the coupling between the pads is minimal. However, this feature is relatively unimportant for EV battery charging applications because in practice the vehicle is normally parked within 10 -15 degrees of the designated parking lot direction.
To examine the flat bar charging type concept, a small scale prototype was constructed as shown in Figure 3-9 (a). Its performance is compared as shown in Figure 3-9 (b) with a similar size 420mm circular pad designed according to [70, 71] as the earlier presented 700mm circular pad discussed in Section 3.3.1. The parameters of both pads are given in Table 3-2. The small scale flat bar pad prototype has a coupling coefficient of 0.42 to 0.15 between an air gap of 50 to 140mm. This coupling coefficient is consistently higher than the 420mm circular pad by $\approx 0.08$ as indicated in Figure 3-9 (b). Although the flat bar prototype uses 1/3 more ferrite than the 420mm circular pad, it shows a much improved performance over the equivalent size circular pad.

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7 The 2-D JMAG simulation plot was kindly provided by Mr. Mickel Budhia
Figure 3-9: (a) Photo of a 420mm flat bar charging pad prototype. (b) Coupling coefficient comparison between 420mm flat bar charging pad and 420mm circular charging pad

<table>
<thead>
<tr>
<th>Parameters of a 420mm flat bar charging pad</th>
<th>Parameters of a 420mm circular charging pad</th>
</tr>
</thead>
<tbody>
<tr>
<td>No. of turns</td>
<td>10</td>
</tr>
<tr>
<td>No. of ferrite bars</td>
<td>12</td>
</tr>
<tr>
<td>Ferrite bar size</td>
<td>118 x 30 x 10mm</td>
</tr>
<tr>
<td>Casing size</td>
<td>420 x 420mm</td>
</tr>
</tbody>
</table>

Table 3-2: Parameters of a flat bar charging pad prototype and a 420mm circular charging pad

3.4.2. Design of a flat bar charging pad

In order for the charging pad magnetic structure to be practical for EV battery charging application, it requires not only a good coupling performance over large air gaps but also a good lateral tolerance to misalignment in order to enable a level of freedom in parking the vehicle. Hence, the transceiver pole areas are not just an extension of the flux pipe, but are deliberately widened in order to increase the tolerance in the y direction. This is explained later in this section. The tolerance limitation in the x direction of the flat bar charging pad is similar to a flat pick-up on a single-phase single conductor track as they both have similar flux line profiles in the x direction (as conceptually demonstrated in Figure 3-10). The x direction tolerance can be increased by lengthening the pole area, but this effect is limited without lengthening the flux pipe coil to channel the coupled magnetic flux through the flux pipe coil [151].
Tolerance in the y direction can be increased by adding ferrite bars in the y direction (referred to as a ferrite wing) to widen the pole areas. Four pairs of flat bar pad experimental prototypes with different pole area structures were implemented to investigate the effect of adding ferrite wings to the y direction. The structures of the four constructed pole areas are shown conceptually in Figure 3-11. Only half a charging pad is shown for the purposes of a clear picture. The pole areas for all four concepts were constructed using two ferrite flat bars of 113x30x10mm and two flat bars of 56x30x10mm. As all four structures used identical volumes of ferrite, increases in coupling in the y direction were the result of a better ferrite bar arrangement. The flux pipe ferrite structure for all four pads was fixed to be 300mm long and 120mm wide. The flux pipe coil was wound with 3mm diameter wire with a gap of one wire diameter in between the turns, and has 46 turns in total. This follows the suggestions in [151] that the pad coil needs to be evenly distributed with a maximum one wire diameter in between turns to achieve good flux linkage from one end of the charging pad to the other end. The inductance and coupling performance for all four pole area structures were measured using an LCR meter with both primary and secondary pads aligned and with a y direction offset of 150mm between them. Both set-ups were tested with a vertical air gap of 145mm. The measured pad inductances and coupling performances are given in Table 3-3.
Figure 3-11: Conceptual diagrams of four implemented flat bar pad pole area structures.

Table 3-3: Measured inductance and coupling coefficient of flat bar pad with different pole area structures shown in Figure 3-11 with vertical offset of 145mm and with y direction offset of 0 and 150mm.

<table>
<thead>
<tr>
<th>Pole area arrangement</th>
<th>$\delta y = 0$</th>
<th>$\delta y = 150$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Figure 3-11 (a)</td>
<td>822.19μH</td>
<td>818.49μH</td>
</tr>
<tr>
<td>Figure 3-11 (b)</td>
<td>847.36μH</td>
<td>838.9μH</td>
</tr>
<tr>
<td>Figure 3-11 (c)</td>
<td>932.3μH</td>
<td>927.88μH</td>
</tr>
<tr>
<td>Figure 3-11 (d)</td>
<td>873.11μH</td>
<td>879.14μH</td>
</tr>
</tbody>
</table>

With the same amount of ferrite being used, the charging pads with wing design (Figure 3-11 (c) and (d)) were found to have a lower coupling performance ($k$ of 0.163 and 0.162 respectively) compared with the pads with no ferrite wings on the sides ($k$ of 0.174 for Figure 3-11 (a) and 0.178 for (b)) providing the primary and the secondary pads were aligned in both the x and y directions. When a misalignment of 150mm in the y direction between the
pads was deliberately introduced, the charging pads with ferrite wings had a coupling factor of 0.132 and 0.126 for Figure 3-11 (c) and (d) respectively. In comparison, the pad without the wings (Figure 3-11 (a) and (b)), was found to have a coupling factor of 0.119 and 0.123 respectively; this demonstrates that the structure using ferrite wings provides a better performance for tolerating misalignment in the y direction, which is a desirable feature for EV battery charging applications.

Up to this point, each flux pipe prototype was wound with a densely packed continuous single strand flat toroidal winding structure covering its full length. In consequence the number of turns required to cover the full length of the flux pipe was high, resulting in a very high inductance which is impractical to drive. For example, if a 20kHz, 15A RMS power supply is used to drive a flat bar pad with the inductance value given in Table 3-3, the voltage across the pad would be as high as 1.7kV. This high voltage stress will make the charging pad component selection very difficult and uses voltage levels which are undesirable for domestic household use for safety reasons.

A common technique for lowering the pad inductance is to use a multiple parallel stranded winding structure as shown conceptually in Figure 3-12 (a) [70]. This will form a pad with windings which are electrically in parallel but magnetically in series and which in operation is identical to that of the earlier flux pipes. Another option is to use two mutually coupled windings at both ends of the flux pipe to approximate a continuous winding structure and to keep the effective length of the flux pipe the same as shown conceptually in Figure 3-12 (b). The mutual coupling between the two separated coils is able to channel the magnetic flux between the pole areas at one end of the pad to the other end. If these two coils are designed to have a good mutual coupling between them (here referred to as the intra-coil coupling) the coupling performance between the primary and the secondary pads (here referred to as the inter-pad coupling) will be approximately the same. The amp-turns determines the flux density in the charging pad. In order to keep the same amp-turns in the primary pad, for the two parallel winding structure of Figure 3-12 (b), the primary current from the supply will need to be doubled. This is similar to the multiple parallel strands winding technique discussed in [70].
A pair of flat bar pads, conceptually shown in Figure 3-13, with the pole area structure given in Figure 3-11 (b) were wound with two mutually coupled coils at both end of the flux pipe (300mm). The measured intra-coil coupling factor was found to be 0.39. Its measurements were then compared with the single strand continuous winding structure presented previously. These two mutually coupled coils were constructed using 23 turns each to keep the volume of wire used in both cases consistent. The measured pad inductances and inter-pad coupling performances of both designs are shown in Table 3-4.

<table>
<thead>
<tr>
<th>Winding structure</th>
<th>$\delta y = 0\text{mm}$</th>
<th>$k$</th>
<th>$\delta y = 150\text{mm}$</th>
<th>$k$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Single 46 turns continuous winding</td>
<td>847.36$\mu$H</td>
<td>0.1784</td>
<td>838.9$\mu$H</td>
<td>0.123</td>
</tr>
<tr>
<td>Two 23 turns winding connected in parallel</td>
<td>230.667$\mu$H</td>
<td>0.1663</td>
<td>224.88$\mu$H</td>
<td>0.12</td>
</tr>
</tbody>
</table>

Table 3-4: Measured inductance and coupling coefficient of flat bar pad structure shown in Figure 3-13 with two different winding structures. These are single 46 turns of continuous winding and two 23 turns winding connected in parallel with an intra-coil coupling factor of 0.39. These measurements are collected with a vertical offset of 145mm and with a y direction offset of 0 and 150mm.
With the measured results presented above, it can be seen that the inter-pad coupling performance of the pad with two mutually coupled windings is slightly reduced by 7% with both pads aligned in both x and y directions and is reduced by only 3.5% with an offset of 150mm in the y direction. Although the primary current $I_1$ is doubled to keep the same $VA_1$ in the primary pad, the voltage stress across the pad is reduced by nearly two times compared with a single strand continuous winding due to the reduction in inductance. Therefore, the slight reduction of inter-pad coupling is a good compromise to ease the practical restraints of the charging pad.

### 3.4.3. Practical Results and Discussion of a Flat Bar Charging Pad Prototype

The conceptual diagram and experimental set up of a practical sized prototype of the proposed flat bar charging pad is shown in Figure 3-14. The parameters of the pad are given in Table 3-5. This prototype is made up of 24 ferrite bars, 118 x 30 x 10mm and the width of the flux pipe is 120mm, which is made up of 4 ferrite bars to keep the flux density below saturation. These ferrite bars have been placed in such a manner that the joints between each of the ferrite bars in the direction of the flux in the midsection are not aligned. This prevents concentrated flux in a particular point when all the ferrite gaps are aligned together and results in a more evenly distributed flux profile in the cross-sectional area of the flux pipe. The flux pipe winding is designed with two 15 turn windings in parallel as shown in Figure 3-14. The measured intra-coil coupling factor for a single pad is 0.41. Similar to the circular pad design, an aluminium back plate and end plate are added (but not shown in Figure 3-14) to contain leakage flux to be within the space between the coupled pads.

<table>
<thead>
<tr>
<th>A: Ferrite bar width</th>
<th>30mm</th>
<th>Pad thickness</th>
<th>30mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>B: Ferrite bar length</td>
<td>118mm</td>
<td>Coil</td>
<td>15 turns each</td>
</tr>
<tr>
<td>C: Flux pipe length</td>
<td>300mm</td>
<td>Coil diameter</td>
<td>4mm</td>
</tr>
<tr>
<td>D: Flux pipe width</td>
<td>120mm</td>
<td>Intra-coil coupling factor</td>
<td>0.41</td>
</tr>
<tr>
<td>E: Pad length</td>
<td>590mm</td>
<td>Number of Ferrite bar</td>
<td>24 bars</td>
</tr>
<tr>
<td>F: Pad width</td>
<td>356mm</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 3-5: Parameter of a 590 x 356mm flat bar charging pad prototype
The self-inductance, mutual inductance, and coupling coefficient of the prototype with variations in vertical air gap from 50mm to 190mm have been measured with an LCR meter and the results are shown in Figure 3-15. A comparison of the coupling coefficient of the flat bar charging pad prototype and the 700mm circular pad with aluminium case presented earlier is given in Figure 3-16. The trend of the results shown in the graph indicates that the coupling performance of these two lumped coil topologies have similar performance for the given measurement within the required range of vertical offset of 150mm to 250mm. The constructed prototype flat bar type charging pad has a $k$ of 0.2 at a vertical offset of 180mm. Similar to the circular pad example presented earlier, if a $k$ of 0.2 at an air gap of 200mm is desired, the flat bar type charging pad size would need to be increased in proportion to the increase in height but only by a factor of 2. In this case, the physical size of the flat bar charging pad would need to be increased by 40mm which will result in a 630mm long charging pad.

A comparison of various key parameters between the constructed flat bar and 700mm circular charging pads is given in Table 3-6. The flat bar charging pad prototype is physically smaller (590mm compared to 700mm) and the amount of ferrite used in the prototype is less ($0.85 \times 10^{-3} \text{m}^3$) compared with the circular charging pad ($1.5 \times 10^{-3} \text{m}^3$). This is because the proposed flat bar pad topology effectively uses all the ferrite to form one single flux pipe in one direction comparing to the circular pad which forms a radial structure. The measurements show that this newly proposed flat bar pad has a coupling performance which is suitable for the larger air gaps required for the EV battery charging systems [72].
Figure 3-15: Measurement with LCR meter of a 590mm flat bar charging pad. (a): Primary inductance with secondary pad open-circuited and short-circuited (b): Mutual inductance and coupling coefficient

Figure 3-16: Coupling comparison between the 700mm Circular pad and the 590mm Flat bar type pad
<table>
<thead>
<tr>
<th></th>
<th>Circular Pad</th>
<th>Flat Bar Type Pad</th>
</tr>
</thead>
<tbody>
<tr>
<td>Physical size (mm)</td>
<td>700mm diameter</td>
<td>590 x 356mm</td>
</tr>
<tr>
<td>$k$ range between 150mm to 190mm</td>
<td>0.25 – 0.16</td>
<td>0.245 – 0.187</td>
</tr>
<tr>
<td>Ferrite volume (m$^3$)</td>
<td>$1.5 \times 10^{-3}$</td>
<td>$0.85 \times 10^{-3}$</td>
</tr>
<tr>
<td>Native quality factor at 20 kHz ($Q_L$)</td>
<td>220</td>
<td>86</td>
</tr>
<tr>
<td>Length of Litz wire (m)</td>
<td>33</td>
<td>8.6</td>
</tr>
</tbody>
</table>

**Table 3-6: Table comparison of 700mm Circular and 590mm Flat bar pad**

While this flat bar charging pad appears to be significantly better being a physically smaller and more resource-effective pad design than the traditional circular charging pad with aluminium case for a similar magnetic coupling performance at the desired 200mm vertical offset, as noted in Table 3-6, its native quality factor $Q_L$ is much lower. This can be explained by examining the structure of the flux pipe. As the flux pipe winding is a type of low profile solenoid structure, it naturally generates flux on both the top and the bottom of the pad as shown in Figure 3-8. As such flux is generated between the flux pipe and the aluminium plate. In the prototype design, the aluminium back plate and end plate are used to contain any leakage flux within the space between the primary and secondary pads and to prevent unaccounted for inductance due to the interaction between the leakage flux and nearby metallic objects. However, in order to achieve this there are excessive eddy current losses generated in the aluminium casing causing its $Q_L$ to be low (2.5 times less than that a 700mm circular charging pad.)

As both pads have a similar coupling performance over the interested vertical offset, the VA generation will be similar, and consequently the standing loss in the flat bar type prototype will be 2.5 times higher than the 700mm circular pad for the same primary current, operating at the same power level. With this reduction in efficiency the proposed pad topology is impractical and uncompetitive against a circular charging pad from an efficiency perspective. However, despite the poor $Q_L$ of the proposed flat bar type charging pad, the flux pipe concept demonstrated promising coupling performance which is very desirable for the EV battery charging application. A new improved design which takes advantage of this without the loss problems is detailed in the following section.
3.5. Development of a Polarised Single Sided Flux Charging Pad

Based on the structural advantage of the flux pipe concept in the flat bar pad, a new polarised single-sided charging pad, named the Double D pad, is proposed here. This design removes the generated flux at the bottom of the charging pad rather than using an aluminium backing plate to shield it. The return paths of each of the flux pipe coils are moved up to the top surface of the ferrite which acts as the flux pipe as demonstrated conceptually in Figure 3-17. This structure retains the magnetic potential of the flux pipe coil in its middle. In consequence, minimal flux is generated at the back of the charging pad similar to a circular pad in this regard, and any aluminium which is placed underneath is only required to minimise leakage (rather than main) flux.

![Figure 3-17: Conceptual diagram of the transition of coil position from a flat bar pad to a Double D pad](image)

3.5.1. Structure of a polarised Double D Pad

Based on the concept discussed above, the structure of the proposed Double D charging pad is shown in Figure 3-18 with its conceptual flux profile [146-148]. It has two flat Archimedean spiral winding coils which sit in a co-planar relationship in close proximity to each other on top of a flat ferrite core. The ferrite core and the associated coil winding form the flux pipe mechanism as indicated by the dashed rectangle in Figure 3-18 [146-148]. For a better inter-pad coupling performance in both vertical and lateral offset, the width and length of the flux pipe coil should be as wide and as long as possible for any given physical size constraint. Conversely, the remaining length of the coil should be minimised to be resource effective and efficient. By doing so, it forms two co-planar windings that are mirror imaged giving rise to the Double D structure as named.
There is no straight path through the flux pipe that passes through the coils. Instead, the arrangement of the Double D coils means that the majority of the flux entering the pad through one of the pole areas propagates through the effective section of coils into the flux pipe core. It then propagates along the core and exits the pad through the other pole area, completing its path back through air to the first pole area (in the form of a complete curved flux path) as shown conceptually in Figure 3-18. The flux path so formed is essentially completely above a top surface of the pad and extends into a space beyond the top surface. The arrangement of the Double D coils sitting above the ferrite backing means that there is essentially no flux extending out the rear face of the pad.

The flux pipe coil arrangement shown in the conceptual diagram is two co-planar windings in close proximity and with good intra-coil coupling connected in parallel. Similar to the flat bar pad structure as explained in Section 3.4.2, the flux pipe coil could be formed by either connecting these two windings in parallel or connecting them in series as one single continuous winding but using multiple parallel strands to lower its inductance, e.g.: bi-filar or tri-filar windings. Both structures can be used to achieve the same effect as the magnetic circuit, given all of these winding structures are identical - they are electrically in parallel and magnetically in series.
A 2-D JMAG simulation of the proposed Double D pad is shown in Figure 3-19. This simulation result which shows the conceptual magnetic flux generated on the top surface of the pad, has a nice vertical distribution profile as expected which is similar to the flat bar type charging pad. However, the key difference is that the magnetic flux generated towards the rear surface of the Double D pad is minimal. There is some leakage flux generated at the side of the pad due to the return path of the coils. This leakage flux does contribute slightly to additional loss in the pad but the native quality factor $Q_L$ measurements showed that this loss is insignificant for a 2kW EV battery charging system. The $Q_L$ measurements of the pad will be presented and discussed further in Section 3.5.2.

![Figure 3-19: A 2-D JMAG simulation of a Double D flux profile](image)

3.5.2. Design of a practical Double D Pad prototype

The structure of Double D pad is fundamentally similar to the flat bar pad as they share key features to enhance their coupling performance. Therefore, the techniques applied in the flat bar pad could also be used here for designing the Double D pad. As discussed earlier, the part of the coil which forms the flux pipe may be spread or distributed across the surface of the ferrite core to ensure that the primary flux in the centre of the pad is constrained within the core. It also extends the effective length of the flux pipe between the two pole areas so that a higher fundamental flux profile can be achieved. As mentioned in Section 3.4.2, it is preferable to keep the windings evenly spaced and if gaps between turns are required they should be less than one wire diameter to keep flux leakage losses through the windings to a minimum [151]. If the coils in the pad are constructed using two parallel windings, they need to have a good intra-coil coupling similar to the flat bar pad. It was

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*The 2-D JMAG simulation plot was kindly provided by Mr. Mickel Budhia.*
found from coupling measurements that these two coils need to be placed no more than a few wire diameters apart to maintain a good intra-coil coupling factor for a practical coupling performance between the primary and secondary pad with a desirable air gap. Otherwise, these two coils do not have enough flux linkages between them and appear as two individual magnetic circuits.

Unlike the flat bar pad structure, the pole areas in the Double D pad structure are enclosed by each coils return path as demonstrated in Figure 3-18. To widen the pole areas for a better lateral tolerance in the y-direction, the width of the flux pipe must necessarily be increased in the process. Similar to the circular pad design presented in [149], a customised ferrite plate, with enough volume of ferrite to keep it below saturation, could be used for the flux pipe core to minimise the weight of the pad. Alternatively, a similar approach presented in [70] for optimising circular pad design could be adopted here. It uses readily available commercial ferrite bars to make up several distributed ferrite strips for the flux pipe as shown in the photograph of a Double D charging pad prototype of Figure 3-20. These ferrite strips are aligned in the direction of the flux path in the core. With this approach, the ferrite core in the pad is a lot stronger and more durable - hence the charging pad is more robust for practical use.

The parameters of the pad shown in Figure 3-20 are given in Table 3-7. In this design, the flux pipe coil arrangement uses two coils wound in parallel. The effective flux pipe section of the coil winding is distributed with a 2.1mm gap in-between each turn to increase its length. The flux pipe ferrite core is made up of 4 distributed strips of ferrite bars. The amount of ferrite used is calculated to keep the average flux density in the core below saturation [147].
Figure 3-20: Top view photo of a 760 x 400mm Double D pad prototype

Table 3-7: Parameters of 760 x 400mm Double D charging pad

<table>
<thead>
<tr>
<th>A: Ferrite bar width</th>
<th>28mm</th>
<th>Pad thickness</th>
<th>45mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>B: Flux pipe width</td>
<td>211mm</td>
<td>Ferrite bar length</td>
<td>93mm</td>
</tr>
<tr>
<td>C: Flux pipe length</td>
<td>241.9mm</td>
<td>Ferrite bar thickness</td>
<td>16mm</td>
</tr>
<tr>
<td>D: Flux pipe length plus pole areas</td>
<td>558mm</td>
<td>Number of ferrite bar</td>
<td>24</td>
</tr>
<tr>
<td>E: Pad length</td>
<td>760mm</td>
<td>Coils</td>
<td>20 turns each</td>
</tr>
<tr>
<td>F: Pad width</td>
<td>400mm</td>
<td>Wire diameter</td>
<td>4mm</td>
</tr>
</tbody>
</table>

With the configuration of the coil winding in the constructed prototype, the intra-coil coupling is 0.4, which is similar to the flat bar type prototype presented in an earlier section. The flux pipe parallel winding structure for the Double D pad design could not be done in a similar fashion to the flat bar type charging pad prototype presented earlier that uses concentrated winding at the ends of the flux pipe with a substantial gap in-between. It was found by experiments measuring both the intra-coil and inter-pad coupling performance that when the two parallel winding coils are not in close proximity, the magnetic circuit of each individual coil formed with the ferrite underneath does not interact with each other. This results in a poor intra-coil coupling and a lower inter-pad coupling coefficient for the extant air gap.
As mentioned earlier, one key characteristic of the Double D charging pad design is that it generates a single-sided flux path in the direction of the pick-up pad. This single sided flux nature of the pad combined with aluminium shielding results in extremely little leakage flux beyond the rear face of the charging pad and a small amount of leakage out at the end of the pad. This makes the Double D pad very efficient and insensitive to nearby metallic objects. Measurements of the $Q_L$ and inductance ($L_1$) of the Double D pad prototype with various aluminium shielding are given in Table 3-8. These measurements show that the Double D pad has an excellent $Q_L$ value typically above 300 with aluminium backing plates and side plates. With a shielding end plate, which intercepts the leakage flux, the $Q_L$ drops slightly by 11% which represents an increase of 11% in the pad standing loss, and the inductance drops by 2.8%. The standing loss calculations of the pad driven by a 30A RMS power supply at 20kHz with various options of shielding are shown in Table 3-8. With both backing plate and end plate, the increased loss is 2.2W above that measured when only the backing plate is used, while the additional loss is 5.6W if aluminium backing, side and end plates are all used. For a 2kW system, this extra loss reduces the overall system efficiency by less than 0.3%. These figures indicate that the Double D pad structure has excellent inter-pad efficiency and is insensitive to nearby metallic objects. Both of these are highly desirable features for any IPT pad system including EV battery charging [70, 72, 147, 148].

<table>
<thead>
<tr>
<th>Shielding Options</th>
<th>$Q_L$</th>
<th>$L_1$ (μH)</th>
<th>Standing loss (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminium back plate</td>
<td>330</td>
<td>147</td>
<td>50</td>
</tr>
<tr>
<td>Aluminium back and side plate</td>
<td>326</td>
<td>145</td>
<td>50.3</td>
</tr>
<tr>
<td>Aluminium back and end plate</td>
<td>312</td>
<td>144</td>
<td>52.2</td>
</tr>
<tr>
<td>Aluminium back, side and end plate</td>
<td>291</td>
<td>143</td>
<td>55.6</td>
</tr>
</tbody>
</table>

Table 3-8: Measurements of $Q_L$ and $L_1$ for various option of aluminium shielding plate and its associated standing loss with a primary power supply outputting 30A RMS at 20kHz.

3.5.3. Practical Measurements on the Double D Charging Pad

The vertical and horizontal characteristics of the 760 x 400mm Double D charging pad were determined by measurements of its self-inductance, mutual inductance and the coupling coefficient. The vertical measurements, shown in Figure 3-21, were taken with variations in the air gap from 110 to 300mm with both pads are aligned. Horizontal tolerance measurements were taken with the pick-up pad sitting at 200mm above the primary pad and moving in the x axis from 0 to 120mm at 40mm steps and in the y axis from 0 to 160mm at
40mm steps where the x axis defined to be across the length E of the Double D pad as shown in Figure 3-20. These measurements against the horizontal offset variation are shown in Figure 3-22 and Figure 3-23.

Within the interested vertical offset range of 150 to 250mm, the vertical offset measurements indicate that the Double D pad achieves a range of coupling coefficients between 0.335 to 0.16. With the pick-up pad sitting at 200mm above the primary pad, moving the pick-up pad from being perfectly aligned towards an offset of 120mm in both the x and y directions shows that the coupling factor varies from 0.237 to 0.145. These measurements indicate that with a 30A RMS 20kHz primary power supply, the loaded secondary quality factor $Q_2$ of the pick-up will be below 6 when delivering 2kW with a power transfer zone of ±120mm in both x and y directions with an air gap of 150 to 200mm. A pick-up operating with a loaded $Q_2$ of 6 is practical at 20kHz for such IPT systems.
Figure 3-21: Measurements of 760 x 400 Double D pad versus vertical offset between 110 to 300mm (a) Pad open-circuited and short-circuited inductance. (b) Mutual inductance and coupling coefficient.
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Figure 3.22: Double D pad open-circuited and short-circuited inductance at z axis of 200mm versus y axis offset at x axis offset of (a) 0, (b) 40, (c) 80, and (d) 120mm.

Figure 3.23: Double D pad mutual inductance M and coupling coefficient k at z axis of 200mm versus y axis offset at x axis offset of (a) 0, (b) 40, (c) 80, and (d) 120mm.
3.5.4. Discussion and comparison between Double D pad and Circular pad

The coupling coefficient for a 760 x 400mm Double D pad versus the 700mm diameter circular pad presented earlier are shown in Figure 3-24 and Figure 3-25 for vertical and horizontal offset tolerances respectively. Even though the surface area of the constructed circular and Double D pads are similar, the comparison shows that the Double D pad has a superior coupling performance to both vertical and horizontal offset tolerances. The parameters of the constructed circular pad and the Double D pad are given in Table 3-9.

<table>
<thead>
<tr>
<th>Physical size (mm)</th>
<th>Circular Pad</th>
<th>Double D Pad</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>700mm diameter</td>
<td>760 x 400mm</td>
</tr>
<tr>
<td>$k$ range between 150mm to 250mm in z direction</td>
<td>0.25 – 0.11</td>
<td>0.335 – 0.16</td>
</tr>
<tr>
<td>$k$ range between (0,0)mm to (120,120)mm with $z = 200mm$</td>
<td>0.145 – 0.07</td>
<td>0.231 – 0.143</td>
</tr>
<tr>
<td>Ferrite volume ($m^3$)</td>
<td>$1.5 \times 10^{-3}$</td>
<td>$1 \times 10^{-3}$</td>
</tr>
<tr>
<td>Native quality factor at 20 kHz ($Q_L$)</td>
<td>220</td>
<td>291</td>
</tr>
<tr>
<td>Length of Litz wire (m)</td>
<td>33</td>
<td>35.1</td>
</tr>
</tbody>
</table>

Table 3-9: Table comparison of 700mm Circular and 760mm Double D pad

Figure 3-24: Coupling coefficient comparison between 700mm Circular pad and 760x400mm Double D pad versus vertical offset
From the comparison table and the previous discussion on coupling performance, it can be seen that the Double D pad topology has a superior coupling performance to an equivalent size circular pad, while using 33% less ferrite. Also, unlike the flat bar type pad which has a low $Q_L$ value, the Double D prototype has a higher $Q_L$ measurement than the circular pad. In practice, when operating under the same VA conditions the Double D pad prototype will have 32% less pad loss than the circular pad. In order to achieve identical coupled VA in their respective pick-ups ($VA_2$), the circular pad will need to operate with a higher primary VA ($VA_1$) to compensate for the difference in $k$. This results in even higher loss.

The only weakness of the Double D charging pad is the ineffective use of Litz wire. The effective section of Litz wire which generates magnetic flux towards the direction of fundamental flux path is the part where it is perpendicular to the flux pipe. In this case, it is the section directly on top of the flux pipe ferrite strips. In order to wind multiple turns to generate the effective length of flux pipe, a return path of the Litz wire, which does not contribute to the desirable flux path, is unavoidably needed. By comparison all the Litz wire
in the circular charging pad structure is generating useful magnetic flux towards the inter-pad coupling flux path. Although the Double D pad structure does not use all its Litz wire as effectively; as shown in Table 3-9 the Double D pad prototype only uses 7% more Litz wire than its equivalent size circular pad but still has a 50% better coupling factor. This is because, similar to the flat bar pad, the flux pipe concept in the Double D pad structure uses the ferrite and its associated section of coils a lot more effectively to form a long single direction flux pipe comparing to the circular pad which has its ferrite distributed in a radial shape to form a much lower flux path profile.

3.6. Discussion

The presented Double D charging pad has a high single-sided flux pattern, tolerance to misalignment and insensitivity to extraneous metal. All these are key technological advances that are needed for IPT EV applications including stationary battery charging. With its structure and characteristics, it is similar to a Halbach array which uses permanent magnets to produce a single sided flux pattern. A conceptual magnetic flux path diagram of both a Halbach array [152] and the Double D charging pad are shown in Figure 3-26 (a) and (b) respectively.

![Figure 3-26: Conceptual magnetic flux path formation comparison between Halbach array and Double D charging pad](image-url)
The Halbach array shows great similarity in the generated magnetic flux path pattern but here with the Double D charging pad the magnetics are AC and a ferrite backing is needed. Nonetheless the Halbach array is acknowledged as a significant advance, and is in widespread use in motors, maglev trains, etc. The Double D charging pad is likewise very significant. It is clearly an improvement on previous lumped coil systems designed for IPT applications where large air gaps are required.

3.7. Conclusion

The use of a flux pipe in IPT lumped coil systems to enhance the coupling performance for large air gap applications is proposed here. Two novel IPT lumped coil systems, named the flat bar type charging pad and the Double D charging pad, are presented, which have been designed using the concept of a flux pipe. Both conceptual prototypes have been compared against an optimised circular charging pad design with aluminium case [71].

The flux pipe is shown to significantly enhance the coupling for pads which are separated with large air gaps. The flat bar type charging pad initially proposed was found to be impractical because it naturally generates magnetic flux on both sides of the pad which makes it sensitive to nearby metallic objects. When shielding is added to mitigate this effect, it adds too much loss to make the system practical. The single sided flux Double D pad, which conceptually has a similar structure to a Halbach Array, only generates flux in one direction, similar to the circular pad, which until now has been the preferred structure for EV charging system. Practical measurements have shown that the Double D pad is material efficient and low loss. It fundamentally has a better coupling than the circular pad. The prototype Double D pad has a 50% better coupling performance than a similar sized circular pad with an air gap of 200mm that has a horizontal tolerance of ±120mm in both x and y directions. Developments of this Double D concept are ongoing with two PhD candidates focusing on magnetic optimisation of this and other similar improved structures.
4.1. Introduction

For a hands-free loosely coupled IPT lumped coil system as used in Electric Vehicle (EV) battery charging application, the magnetic structure for both the primary and the secondary charging pads often includes ferrimagnetic material to enhance the magnetic coupling as described in Chapter 3. Ferrimagnetic material is also used in conjunction with aluminium plates for magnetic shielding purposes for complying with emissions regulations and guidelines such as ICNIRP [70].

The biggest advantage of IPT for EV battery charging applications is its tolerance to misalignment between the primary and secondary pads due to its loosely coupled nature. The physical movement of the ferrimagnetic material and coils in the pick-up relative to the primary pad necessarily introduces variations in the magnetic coupling between the pads and also introduces variation in the pad self-inductance as illustrated in Chapter 3. In this scenario, it is impossible for both the primary and the secondary charging pads to always be accurately tuned over a given range of movement within a specified power transfer zone without adopting self-tuning circuitry [138, 153, 154]. In a fixed-frequency primary-side current-controlled system, this places additional reactive load on the power supply, although it does not affect the power transfer capability to the pick-up providing the track current can be regulated at the desired magnitude. However, the mistuned resonant network necessarily introduces additional reactive load in the system and this reactive load increases the losses in the power supply. A general load model of various tuning networks designed for IPT systems was presented in [109] and it has been applied to distributed track applications in order to estimate the impact of the pick-up reflected impedance [66]. However, the analysis was done
on the basis of an ideally tuned pick-up with no ferrimagnetic material on the primary side. In a lumped coil system with ferrite in both primary and secondary, the pick-up mistuning effect must be included in order to estimate the reflected equivalent impedance of the pick-up.

This chapter presents the fundamental structure of a commonly used industrial IPT power supply, and pick-up tuning topologies for EV battery charging application. In a primary side current control system, the pick-up structure can be designed to be as simple as possible [132], allowing both primary and secondary mistuned resonant networks to be modelled by considering only the fundamental frequency component. Using the developed model, the impact on the power supply due to both mistuned primary and secondary resonant networks operating at rated load is investigated and a number of design considerations are discussed. A design methodology is proposed for both networks to minimise the reactive load in the primary power supply when operating at the rated load. At the end of this chapter, a tuning network for a 1.2kW system with a pair of 700mm circular charging pads is designed using the proposed strategy, and simulated results are presented.

4.2. Fundamental Structure of a Lumped Coil Battery Charging System

The most common industrial IPT power supply topology is the full bridge voltage-sourced inverter with a series-parallel LCL resonant network as outlined in Chapter 2. As this chapter is focusing on modelling the mistuned resonant networks for variable coupling applications and the reactive power flow between the primary and the secondary, the impact on the power supply switching is not analysed here and only the fundamental frequency is considered for simplicity purposes. A conceptual representation of the power supply with a coupled pick-up is illustrated in Figure 4-1 where the inverter bridge output is represented by its fundamental voltage component \( V_{B_1} \). In some cases a capacitor \( C_{L1} \), in series with the primary track inductor \( L_1 \), is used to partially compensate \( L_1 \) to have a total reactance of \( X \), the designed characteristic impedance of the primary LCL network. In a primary side current control system, minimal control is required in the pick-up so it consists of only the bridge rectifier and a DC filter. The track current is directly controlled by the fundamental component of the input inverter bridge voltage \( V_{B_1} \) as given in (4.1) [117].

\[
I_1 = \frac{V_{B_1}}{X} \tag{4.1}
\]
The three most common pick-up tuning topologies are series-tuned, parallel-tuned and LCL tuned topologies as described in Chapter 2. Considering only the fundamental frequency, the pick-up can be modelled as an LCR circuit using an equivalent AC resistive load to represent the DC load as shown in Figure 4-2 [106]. This equivalent AC resistive load for a series tuned or LCL tuned pick-up is given by [106]:

$$R_{AC} = \frac{8}{\pi^2} R_{DC}$$  \hspace{1cm} (4.2)

and for a parallel tuned pick-up is given by:

$$R_{AC} = \frac{\pi^2}{8} R_{DC}$$  \hspace{1cm} (4.3)

As the DC output is represented by its equivalent AC load, the reflected impedance of these three ideally tuned pick-up topologies, discussed in Chapter 2, can be directly applied here and are shown below: [106, 109]
Chapter 4  
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\[
Z_r = \begin{cases} \omega M^2 \frac{L_2}{R_{AC}} = \frac{\omega M^2}{L_2} Q_2 & \text{series tuned} \\ \frac{M^2}{L_2} (R_{AC} - j \omega L_2) = \frac{\omega M^2}{L_2} (Q_2 - j) & \text{parallel tuned} \\ \frac{\omega^2 M^2 R_{AC}}{X_2^2} = \frac{\omega M^2}{L_2} Q_2 & \text{LCL tuned} \end{cases} \tag{4.4}
\]

(4.4) illustrates that the reflected impedance of both the series tuned and the LCL tuned pick-up have the same characteristic. The reflected load is purely resistive and is directly proportional to the pick-up loaded \(Q_2\) when the pick-up coil is tuned in situ. As these two topologies share the same characteristic, in this chapter only the parallel-tuned and the series-tuned pick-ups are investigated for variations in magnetic coupling.

4.3. Load Modelling of a Mistuned Primary and Secondary Resonant Network

In order to investigate the reactive load in the primary power supply due to a combination of both a mistuned primary and secondary resonant network, the reflected impedance of both a mistuned series-tuned and a mistuned parallel-tuned pick-up are modelled. The model, in conjunction with measured track tuning variations, allows the reactive load in the power supply inverter bridge to be calculated. In this section the load model of a mistuned pick-up is presented, followed by the model of a mistuned primary resonant network.

4.3.1. Reflected Impedance of Mistuned Pick-up

4.3.1.1. Series tuned

The conceptual diagram of a mistuned series-tuned pick-up is shown in Figure 4-3. The terms \(L_{20}\) and \(C_{20}\) are the (designed) nominal value of the pick-up inductance and its tuning capacitance. The variation in the pick-up inductance is modelled using the term \(\Delta L_2\) which is defined by:

\[
\Delta L_2 = L_2 - L_{20} \tag{4.5}
\]

where \(L_2\) is the pick-up inductance at the current physical position.
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Figure 4-3: Conceptual diagram of a mistuned series-tuned pick-up. Note that both $L_{20}$ and $\Delta L_2$ are coupled with $L_1$.

The input impedance of the series-tuned pick-up $Z_{2s}$ is given by:

$$Z_{2s} = R_{AC} + j\omega L_{20} + \frac{1}{j\omega C_{20}} + j\omega\Delta L_2$$

$$= R_{AC} + j\omega\Delta L_2$$

(4.6)

The definition of the pick-up reflected impedance $Z_r$ presented in [106, 109] is given by:

$$Z_r = \frac{\omega^2 M^2}{Z_2}$$

(4.7)

Substituting (4.6) into (4.7), the mistuned series-tuned pick-up reflected impedance $Z_{rs}$ is then given by:

$$Z_{rs} = \frac{\omega^2 M^2 R_{AC}}{R_{AC}^2 + (\omega\Delta L_2)^2} - j\frac{\omega^2 M^2 (\omega\Delta L_2)}{R_{AC}^2 + (\omega\Delta L_2)^2}$$

(4.8)

From (4.8), it can be seen that $\Delta L_2$ causes a reflected reactance, which has the opposite polarity to $\Delta L_2$. The real and the imaginary parts of (4.8) share many common terms, and it is convenient to derive an expression for the ratio between the reactive and the real load:

$$\frac{\text{Im}(Z_r)}{\text{Re}(Z_r)} = -Q_{20}\gamma$$

(4.9)

where $\gamma$ is the per unit (pu) variation of $\Delta L_2$ with respect to the designed tuning inductance $L_{20}$ and is defined by:

$$\gamma = \frac{L_2 - L_{20}}{L_{20}} = \frac{\Delta L_2}{L_{20}}$$

(4.10)
and $Q_{20}$ is the nominal loaded quality factor of the pick-up when tuned at the designed operating position, defined by:

$$Q_{20} = \frac{\omega L_{20}}{R_{AC}}$$ (4.11)

In a battery charging application running at a constant output voltage and power for the majority of the time, the load $R_{AC}$ and hence $Q_{20}$ are normally maintained constant. Therefore, (4.9) is a very useful expression for estimating the reactive power as it only consists of the designed circuit $Q_{20}$ and the pick-up inductance variation and directly indicates the polarity of the reflected reactive power.

Using the reflected resistive load in (4.8), the track current ($I_{1s}$) required to deliver the desired output power can be simply calculated using $P = \text{Re}(Z_{rs})I_{1s}^2$ and is given by:

$$I_{1s} = \frac{1}{\omega M} \sqrt{\frac{P_{out} \left( R_{AC}^2 + (\omega \Delta L_2)^2 \right)}{R_{AC}}}$$ \hspace{1cm} (4.12)

$$= \frac{1}{\omega M} \sqrt{P_{out} R_{dc} \left( 1 + (Q_{20} \gamma)^2 \right)}$$

Using (4.12), the increase in track current $I_1$ as a result of the pick-up being mistuned can be expressed by:

$$\frac{I_{1s\_mistuned}}{I_{1s\_tuned}} = \sqrt{1 + (Q_{20} \gamma)^2}$$ (4.13)

4.3.1.2. Parallel tuned

The conceptual diagram of a mistuned parallel-tuned pick-up is shown in Figure 4-4. Here the variation in tuning due to variations in $L_2$ is represented using the term $\Delta C_2$ which is given by:

$$\Delta C_2 = \frac{1}{\omega^2} \left( 1 - \frac{1}{L_{20}} \right) = C_{20} - C_2$$ (4.14)

where $C_{20}$ is the nominal tuning capacitance with the designed value of $L_{20}$, and $C_2$ is the ideal tuning capacitance value of $L_2$. 

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Using a Norton transformation on the pick-up resonant network, the pick-up coil current, and hence the pick-up reflected impedance $Z_{rp}$, can be described by (Appendix A):

$$Z_{rp} = \frac{M^2/R_{AC}}{L_2^2 + L_2^2 (\omega \Delta C_2)^2} - j \frac{\omega M^2}{L_2^2} - j \frac{\omega M^2 \Delta C_2}{L_2^2 + L_2^2 (\omega \Delta C_2)^2}$$

This expression (4.15) indicates that the reflected impedance of a mistuned parallel-tuned pick-up, has two reactive components. The first one is the capacitive component (-$j\omega M^2/L_2$) which was described in Chapter 2 and in (4.4). This capacitive component is independent of the load but proportional to the magnetic coupling. It is normally included in the primary track inductance when the power supply operates with a parallel-tuned pick-up at setup for systems with constant coupling (such as monorail systems). Variations in this term (-$j\omega M^2/L_2$) due to changes in the magnetic structure will not be discussed here as this is regarded as a change in the primary track inductance due to physical movement of the charging pad, but will be discussed later in Section 4.5.1.2.

Similar to the series-tuned pick-up, the second reactive component is introduced by the variable capacitor $\Delta C_2$ and exhibits a polarity opposite to that of $\Delta C_2$. The ratio between the variable reflected reactive load, which excludes the (-$j\omega M^2/L_2$) term, and the resistive load is given in (4.16) and it is nearly identical to (4.9) for the series-tuned pick-up.

$$\frac{\text{Im}(Z_{rp})}{\text{Re}(Z_{rp})} = -Q_{20} \delta$$

Here $\delta$ is the variation of $\Delta C_2$ with respect to the designed tuning capacitance $C_{20}$ and is defined by:

$$\delta = \frac{\Delta C_2}{C_{20}} = \frac{\omega^2 L_{20} \Delta C_2}{C_{20}}$$

Figure 4-4: Conceptual diagram of a mistuned parallel-tuned pick-up
The nominal loaded quality factor \( Q_{20} \) of the pick-up when tuned at the designed operating position is defined by:

\[
Q_{20} = \frac{R_{AC}}{\omega L_{20}}
\]  

(4.18)

Using the resistive term of (4.15), the required track current for a mistuned parallel-tuned pick-up is then:

\[
I_{1p} = \frac{L_2}{M} \sqrt{PR_{AC} \left( \frac{1}{R_{AC}^2} + (\omega \Delta C_v)^2 \right)} = \frac{L_2}{M} \sqrt{\frac{P}{R_{AC}} \left( 1 + (Q_{20} \delta)^2 \right)}
\]  

(4.19)

Using (4.19), the increase in track current \( I_1 \) due to the pick-up being mistuned is given in (4.20), and as expected, is similar to (4.13):

\[
\frac{I_{1p, \text{mistuned}}}{I_{1p, \text{tuned}}} = \sqrt{1 + (Q_{20} \delta)^2}
\]  

(4.20)

### 4.3.2. Load Modelling of the Primary Resonant Network

The conceptual diagram of a voltage-sourced LCL resonant power supply is shown in Figure 4-5 with the pick-up equivalent reflected impedance \((\text{Re}(Z_r) + j\text{Im}(Z_r))\). As the reflected reactance is in series with the primary track inductance \(L_1\), it is convenient to interpret \(\text{Im}(Z_r)\) in terms of inductance. To define this pick-up equivalent reflected inductance \(L_r\), the operating frequency \(\omega\) is considered:

\[
L_r = \frac{\text{Im}(Z_r)}{\omega}
\]  

(4.21)

![Figure 4-5: Conceptual diagram of the primary LCL network load modelling](image)

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The measured primary track inductance $L_1$ within the given pick-up movement tolerance is separated into two components here: $L_{10}$ and $\Delta L_1$ ($L_1 = L_{10} + \Delta L_1$). $L_{10}$ is the nominally designed track tuning inductance, and its reactance combined with $C_L$ is the primary LCL network characteristic impedance $X$. $\Delta L_1$ represents the difference between the measured track inductance $L_1$ and the nominal track inductance $L_{10}$. However, the total inductance variation seen by the power supply $\Delta L_{\text{p}}$ is a combination of $\Delta L_1$ and the pick-up reflected inductance $L_r$ as illustrated in Figure 4-5 and (4.22). Thus it is difficult to estimate the overall inductance variation $\Delta L_{\text{p}}$ while choosing the tuning value $L_{10}$ based on measured $L_1$.

\[
\Delta L_{\text{p}} = L_1 - L_{10} + \frac{\text{Im}(Z_r)}{\omega} = \Delta L_1 + L_r = \Delta X_f/\omega \tag{4.22}
\]

Here $\Delta X_f$ is the total output reactive load of the LCL network.

Instead of choosing the tuning value $L_{10}$ based only on the measured value $L_1$, it is recommended to firstly combine the calculated pick-up reflected inductance $L_r$ together with $L_1$ to form one single inductive component called $L_{1\text{equiv}}$, which is the total equivalent track inductance seen by the tuning capacitor $C_L$ as illustrated in Figure 4-5 and in (4.23).

\[
L_{1\text{equiv}} = L_1 + L_r = L_{10} + \Delta L_{\text{p}} \tag{4.23}
\]

Then based on the value of $L_{1\text{equiv}}$, $L_{10}$ can now be designed to minimise the variation in the total track inductance and to achieve a desirable pattern of $\Delta L_{\text{p}}$ within the misalignment tolerance, in order to minimise the reactive load within the primary power supply. This minimisation process is explained further in Section 4.4.3.2. The LCL network output impedance $Z_f$ is expressed by:

\[
Z_f = \text{Re}(Z_r) + j\omega(\Delta L_1 + L_r) = \text{Re}(Z_r) + j\Delta X_f \tag{4.24}
\]

Using (4.24), the input impedance ($Z_{\text{in}}$) of the primary LCL network is given by: [155]
\[ Z_{in} = \frac{X^2}{Z_l} \]
\[ = \frac{X^2}{\text{Re}(Z_r) + j\Delta X_l} \]
\[ = \frac{X^2 \text{Re}(Z_r)}{\text{Re}(Z_r)^2 + \Delta X_l^2} - j \frac{X^2 \Delta X_l}{\text{Re}(Z_r)^2 + \Delta X_l^2} \]

(4.25)

The input displacement power factor (DPF_{LCL}) of the primary LCL resonant network is then given by:

\[ DPF_{LCL} = \frac{\text{Re}(Z_{in})}{\sqrt{\text{Re}(Z_{in})^2 + \text{Im}(Z_{in})^2}} \]
\[ = \frac{\text{Re}(Z_r)}{\sqrt{\text{Re}(Z_r)^2 + \Delta X_l^2}} \]

(4.26)

Using the \( V_{B,1} \) expression in (4.1) and (4.26), for a given power \( P_{out} \) the fundamental component of the inverter bridge current \( I_{B,1} \) is then given by:

\[ I_{B,1} = \frac{P_{out}}{DPF_{LCL} V_{B,1}} \]
\[ = \frac{P_{out}}{DPF_{LCL} I_X} \]

(4.27)

4.4. Design Considerations

4.4.1. Additional pick-up reactive power due to mistuning

The additional reactive load \( \text{Im}(Z_r) \) of both the series-tuned and parallel-tuned pick-ups, reflected back onto the track is proportional to the loaded \( Q_{20} \) and the tuning variation \( \gamma \) or \( \delta \) respectively. \( \text{Im}(Z_r) \) also represents an increase in the reactive load within the pick-up resonant network compared with an ideally tuned pick-up. Practical measurements of charging pad inductance (as presented in Chapter 3) indicate that the pad inductance has a variation of typically 2-7% depending on allowed (expected) misalignment. The pick-up loaded \( Q_2 \) for conventional distributed IPT systems is typically designed to be below 10 [16] and for a practical IPT battery charging system is normally kept below 6 [70]. With a \( Q_{20} \) value of 6, the additional reactive load is between 12-42% of the real power. If the magnetic
structure has a bigger inductance variation ($\delta$ or $\gamma > 0.15$), for the same $Q_{20}$ of 6, the additional reactive load would be 90% of the real power. This increases the stress in the pick-up resonant components and thus the component ratings need to be significantly higher than would be indicated by an ideally tuned design.

**4.4.2. Increase of track current with mistuned pick-up**

Both (4.13) and (4.20) illustrate that the track current needs to be increased to deliver the same output power (constant $Q_{20}$ and $R_{AC}$ for primary side current control) to a mistuned pick-up. However, increasing $I_1$ also increases the track conduction loss. The square of the required track current increase ($I_{1\_mistuned}/I_{1\_tuned}$)$^2$, which represents the increase in the conduction loss, as a function of the pick-up tuning variation $\gamma$ and $\delta$ with various values of $Q_{20}$ is shown in Figure 4-6. With a $Q_{20}$ of 3 and with a pick-up tuning variation of 7.5%, the increased conduction loss is 5% compared with a pick-up that is always tuned. With the same tuning variation but a $Q_{20}$ of 6, the conduction loss increase is 4 times higher (20%) as shown in Figure 4-6.

![Figure 4-6: Track current variation versus pick-up tuning variation for various values of $Q_{20}$](image)

As illustrated in (4.12) and (4.19), the required track current to achieve the same $Q_{20}$ is inversely proportional to the coupling condition. Thus, the increase in the track conduction loss will be more significant at operating positions with low coupling compared to operating positions with higher coupling as the required track current at a low coupling position is necessarily higher than that at a high coupling position to deliver the same amount of power. If a design decision is made to minimise the charging pad conduction loss without using self-tuning circuitry on the secondary side [138, 153, 154], the best practice for systems with high
Q_{20} (near 6) and high inductance variation ($\gamma$ or $\delta > 0.1$) is to ensure the pick-up is tuned at the operating position with the lowest coupling.

4.4.3. Reactive power flow of the LCL network

The input DPF of the primary LCL network is controlled by its output reactive load $\Delta X_l$ as illustrated in (4.26) where Re($Z_d$) is assumed to be constant at a fixed operating position during steady state. Therefore, choosing the primary tuning $\omega_{L10}$, which determines $\Delta X_l$, is key to determining the burden of reactive load on the inverter bridge within the specified pad power transfer zone. There are two considerations when choosing the primary tuning $\omega_{L10}$ and they are discussed following in this section.

4.4.3.1. Ensuring inductive load ($Z_{in}$) for inverter bridge voltage $V_{B,1}$

The LCL tuning network has an impedance converting characteristic as explained in Chapter 2 and illustrated in (4.25). Therefore by ensuring $\Delta X_l$ is either zero or capacitive within the misalignment tolerance, $Z_{in}$ is ensured to be either pure resistive or only slightly inductive. Thus, the DPF between $V_{B,1}$ and $I_{B,1}$ is either unity or slightly lagging which is normally preferred in inverter bridge design to prevent undesirable switching losses due to the diode reverse recovery currents in the switches [130].

4.4.3.2. Reactive power minimisation of the primary LCL network

In order to minimise the reactive load and to achieve the best possible input DPF of the primary LCL network, both the pick-up tuning $\omega_{L20}$ and the primary tuning $\omega_{L10}$ need to be chosen carefully to result in a desirable variation pattern of $\Delta X_l$. Using the expressions for the required track current in series and parallel-tuned pick-ups given in (4.12) and (4.19) respectively, the common definition of the LCL network output reactive power can be expressed by:

$$VAR = I_1^2 \Delta X_l$$

$$\propto \frac{\Delta X_l}{M^2}$$  \hspace{1cm} (4.28)

This indicates that the additional reactive load within the LCL network is proportional to $\Delta X_l$ and inversely proportional to $M^2$ which represents the relative coupling variation for a given charging pad design. Therefore, in order to minimise the reactive load, the variation in
\( \Delta X_f \) should be minimised at an operating position where the coupling \((M^2)\) is at its lowest, so that the overall reactive load seen by the inverter bridge is minimised.

### 4.4.4. Inverter bridge current

Using the track current expressions in (4.12) and (4.19), and the inverter bridge current \( I_{B,1} \) expression in (4.27), \( I_{B,1} \) can now be expressed in terms of the output power, the mutual coupling and the DPF, as shown in (4.29), indicating that \( I_{B,1} \) is directly proportional to the output power and the magnetic coupling.

\[
I_{B,1} \propto \frac{P_{out}}{DPF_{LCL}} \tag{4.29}
\]

For a variable coupling system using primary side current control, the inverter bridge current will necessarily have the same range of variation as the magnetic coupling in order to maintain the output power to be constant. This bridge current variation needs to be treated carefully and the power supply inverter design necessarily needs to be rated for the maximum possible coupling variation.

### 4.4.5. Design flow chart

Conventionally, the tuning network of a lumped coil system is designed based on the physically measured charging pad inductances at the same operating position, which then becomes the optimum position of operation. However, as discussed earlier, when designing tuning networks for lumped coil systems with a specified power transfer zone, there are a number of design issues that need to be considered in order to achieve a suitable result.

In practice, it is difficult to achieve all the considerations presented above in one tuning network. To assist the design process, a design flow chart which combines the considerations discussed earlier is presented in Figure 4-7. This flow chart presents a sequence of steps, beginning with the design of the secondary tuning \( L_{20} \), and then considering the primary tuning \( L_{10} \) in order to achieve the design requirements which include minimising the primary pad conduction loss, or achieving the best possible input DPF in the primary LCL network. The default design focus is to minimise the primary charging pad conduction loss and thus the pick-up tuning is designed accordingly. If minimising the reactive load in the primary resonant network is the priority, the primary and the secondary tuning is then adjusted to achieve this.
Figure 4-7: Design process for charging pad resonant network
4.5. Design Example

A tuning network of a 1.2kW EV battery charging system designed using the strategy in Figure 4-7 is presented in this section. The analytical results determined using the designed network are compared against systems with charging pad tuning networks designed at both the maximum and minimum coupling position to demonstrate the improvement in the input loading variation of the primary LCL network. The analytical results of the proposed design are also verified against SPICE simulations.

4.5.1. System parameters

4.5.1.1. Primary power supply

The conceptual structure of the 1.2kW battery charging system is shown in Figure 4-8 and its parameters are given in Table 4-1. The inverter bridge voltage $V_{B_1}$ has a voltage variation range of 0 to 225V RMS to perform primary side current control for regulating the power flow to the pick-up load ($R_{AC}$). As explained in Section 4.2, the AC load $R_{AC}$ has different values for series and parallel-tuned pick-ups in order to have the same equivalent DC output power and voltage.

![Figure 4-8: Structure of the designed 1.2kW battery charging system with equivalent AC resistor load at the pick-up output.](image)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{B_1}$ range</td>
<td>0 – 225V RMS</td>
</tr>
<tr>
<td>Frequency</td>
<td>20kHz</td>
</tr>
<tr>
<td>$C_B$</td>
<td>1.043$\mu$F</td>
</tr>
<tr>
<td>$L_B$</td>
<td>87$\mu$H</td>
</tr>
<tr>
<td>$P_{out}$</td>
<td>1.23kW</td>
</tr>
<tr>
<td>$R_{AC}$ (series)</td>
<td>26.34$\Omega$</td>
</tr>
<tr>
<td>$R_{AC}$ (parallel)</td>
<td>40$\Omega$</td>
</tr>
<tr>
<td>$M$</td>
<td>13.22$\Omega$</td>
</tr>
<tr>
<td>$I_{1\text{ max}}$</td>
<td>34A RMS</td>
</tr>
</tbody>
</table>

Table 4-1: Parameters of the 1.2kW battery charging system
4.5.1.2. Charging pad magnetic structure

The selected charging pad magnetic structure in this design example is the 700mm circular charging pad presented in Chapter 3. The operating air gap is between 100mm and 150mm with a lateral tolerance of ±100mm. This forms a rectangular boundary within a specified power transfer zone as illustrated in Figure 4-9. The position where the pick-up pad is at the tightest coupling position (closest to the primary pad) is labelled “A” and the position where the pick-up pad has the lowest coupling position is labelled “B” in the diagram. The proposed tuning strategy and conventional tuning methods are examined with the pick-up pad moving along the horizontal boundary (δx from 0 to 100mm) at the extreme vertical boundary (150mm and 100mm).

Figure 4-9: Conceptual diagram of circular pad tolerance of movement

The magnetic structures of both charging pads are identical. The primary pad adopts a bi-filar winding with 12 turns where the inductances of both windings are nearly identical as they share a nearly identical magnetic structure. The measured inductance for the inner coil is 127.49μH and the inductance for the outer coil is 128.79μH. Regarding the secondary pad, a different number of turns is used for the series and parallel-tuned pick-up. As explained in Chapter 2, the series-tuned pick-up acts as a voltage source and therefore boosts the current while the parallel-tuned pick-up acts as a current source and boosts its output voltage. In order for these two tuning topologies to have the same output DC voltage and power with the same magnetic structure, the number of turns on the pick-up pad needs to be designed for a suitable $V_{oc}$ and $I_{sc}$ ratio. The winding structure of each of the charging pads which achieves
the same equivalent DC output characteristics used in this design example is outlined in Table 4-2.

<table>
<thead>
<tr>
<th>Winding structure</th>
</tr>
</thead>
<tbody>
<tr>
<td>Primary charging pad</td>
</tr>
<tr>
<td>Bi-filar with 12 turns (2x12 turn)</td>
</tr>
<tr>
<td>Secondary charging pad: series-tuned pick-up</td>
</tr>
<tr>
<td>Single wire with 24 turns</td>
</tr>
<tr>
<td>Secondary charging pad: parallel-tuned pick-up</td>
</tr>
<tr>
<td>Bi-filar with 12 turns (2x12 turn, identical to the primary pad)</td>
</tr>
</tbody>
</table>

Table 4-2: Winding structures of the 700mm circular charging pad for both primary and secondary

Measured inductances of the circular charging pad at an air gap of 100mm and 150mm with x direction movement between 0 to 100mm are shown in Figure 4-10. The open-circuit bi-filar pad inductance measurements are for primary pad tuning when a series-tuned pick-up is used, while the short-circuited inductance measurements are for primary pad tuning when a parallel-tuned pick-up is used. This is because the short-circuited measurements include the pad self-inductance $L_1$ and pick-up reflected capacitive component ($-M^2/L_2$) as illustrated in (4.30).

$$L_{1sc} = L_1 - \frac{M^2}{L_2} \quad (4.30)$$

Since the secondary charging pad in a parallel-tuned pick-up has the same winding structure as the primary pad, the self-inductance of the secondary pad is identical to the measured primary self-inductance shown in Figure 4-10 (a) and (b). The measured secondary pad self-inductance with a single wire wound structure as used in the series-tuned topology is shown in Figure 4-10 (c) and (d). The calculated mutual inductance using the open-circuit and short-circuit measurements are shown in Figure 4-10 (e) and (f) [150]. This calculated mutual inductance is referred back to the primary side. When using this calculated mutual inductance value with a series-tuned pick-up, a turns ratio of 2 needs to be taken into account.

These measurements indicate that the charging pad self-inductance has a variation of 7% from the operating position with maximum coupling (labelled A on the graph) to the minimum coupling position (labelled B on the graph). This variation occurs due to the relative physical movement between the ferrimagnetic materials and the coils in the charging pad. The coupling variation between the two extreme points A and B is about a factor of two as shown in Figure 4-10 (e) and (f) where the mutual inductance varies from 55μH to 23μH.
4.5.2. Performances of the various tuning options

In this design example, the maximum variation in inductance is below 7% and the value of the loaded $Q_2$ is below 3, which is given in the next section. With such low inductance variations and a small $Q_2$, the required increase in track current to deliver the rated power with a mistuned pick-up is only in the range of 2 - 3% as illustrated in Figure 4-6. Therefore, the pick-up is not necessarily required to be tuned at the minimum coupling.
position B, so the tuning network design focus is then to minimise the input load variation of the primary LCL network. The performance of the system, designed using the proposed strategy, is compared against systems with both charging pads designed at the maximum coupling position, which is referred to as “AA”, and at the minimum coupling position, which is referred to as “BB”.

4.5.2.1. Series tuned pick-up

The parameters of the tuning network designed at operating position AA, BB and designed using the proposed methodology of Section 4.4.5 are given in Table 4-3. In the design using the proposed methodology, the nominal tuning value of the pick-up pad L\textsubscript{20} is its self-inductance at the minimum coupling position B and the nominal tuning value of the primary pad is a calculated value of 128\(\mu\)H as determined from Figure 4-11 (d) and explained later. The analytical results of these three tuning network designs are shown in Figure 4-11 and the proposed design is verified with SPICE simulations which are shown in Figure 4-12.

<table>
<thead>
<tr>
<th></th>
<th>L\textsubscript{10}:</th>
<th>C\textsubscript{L1}:</th>
<th>L\textsubscript{20}:</th>
<th>Q\textsubscript{20}:</th>
</tr>
</thead>
<tbody>
<tr>
<td>AA:</td>
<td>134.8(\mu)H</td>
<td>2.14(\mu)F</td>
<td>571.31(\mu)H</td>
<td>2.73 Max. (\gamma): -0.0607</td>
</tr>
<tr>
<td>BB:</td>
<td>126.57(\mu)H</td>
<td>2.963(\mu)F</td>
<td>536.61(\mu)H</td>
<td>0.0647</td>
</tr>
<tr>
<td>New tuning design:</td>
<td>128(\mu)H</td>
<td>2.778(\mu)F</td>
<td>536.61(\mu)H</td>
<td>0.0647</td>
</tr>
</tbody>
</table>

Table 4-3: Tuning network parameters of the various tuning options for series-tuned pick-up

In the calculated pick-up reflected equivalent inductance L\textsubscript{r} shown in Figure 4-11 (a) and (b), tuning option AA has the smallest variation compared with tuning option BB and the new tuning design option. But in the L\textsubscript{1eqv} graph shown in Figure 4-11 (c) and (d), tuning option AA has the biggest primary inductance variation seen by the tuning capacitor C\textsubscript{L1} compared with the other options. Since the biggest \(\Delta X\textsubscript{r}\) for tuning option AA occurs at the minimum coupling position, it has the biggest reactive load variation at the input of the primary LCL network as shown in Figure 4-11 (e) to (h).

Figure 4-11 (d) demonstrates that the biggest value of L\textsubscript{1eqv} in the new tuning design is 128\(\mu\)H and hence that is the chosen L\textsubscript{10} value. Although option BB demonstrates the best DPF performance compared with the new tuning as shown in Figure 4-11 (f), option BB also results in a capacitive load at the input of the LCL network for the pick-up pad moving in the
x direction at the 150mm air gap. Therefore, the new tuning design has minimised the primary DPF variation while keeping the load on the inverter bridge inductive. The SPICE simulation results of the new tuning design, given in Figure 4-12, show very good agreement with the analytical results.
Chapter 4  Design Considerations for Variable Coupling Lumped Coil Systems

Figure 4-11: Analytical results of the primary LCL network with 700mm circular pad and series-tuned pick-up tuned at position AA, BB and the new tuning design versus $\delta x$ at $z$ of 100mm and 150mm
4.5.2.2. Parallel tuned pick-up

The parameters of the tuning network designed at operating position AA, BB and the new tuning design are given in Table 4-4. In the new tuning design the nominal tuning value of the pick-up pad L_{20} is its self-inductance at the maximum coupling position A and the nominal tuning value of the primary pad is calculated to be 123.8μH as determined from Figure 4-13 (d) and explained later. The analytical results of these three tuning network
designs are shown in Figure 4-13 and the proposed design is verified against SPICE simulations shown in Figure 4-14.

<table>
<thead>
<tr>
<th></th>
<th>L10:</th>
<th>C_{L1}:</th>
<th>L_{20}:</th>
<th>Q_{20}:</th>
<th>Max. δ:</th>
</tr>
</thead>
<tbody>
<tr>
<td>AA:</td>
<td>111.81μH</td>
<td>9.58μF</td>
<td>134.8μH</td>
<td>2.36</td>
<td>-0.065</td>
</tr>
<tr>
<td>BB:</td>
<td>122.17μH</td>
<td>3.73μF</td>
<td>126.57μH</td>
<td>2.515</td>
<td>0.061</td>
</tr>
<tr>
<td>New tuning design:</td>
<td>123.8μH</td>
<td>3.4μF</td>
<td>134.8μH</td>
<td>2.36</td>
<td>-0.065</td>
</tr>
</tbody>
</table>

Table 4-4: Tuning network parameters of the various tuning options for parallel-tuned pick-up

In the calculated pick-up reflected equivalent inductance (L_r) and the primary L_{1eqv} inductance shown in Figure 4-13 (a) - (b) and (c) – (d) respectively, tuning option AA and the new tuning design have the smallest variation compared with tuning option BB. In the L_{1eqv} graph shown in Figure 4-13 (d), the calculated L_{1eqv} graph has a maximum value of 123.8μH and hence that is the chosen L_{10} value for the design.

Although tuning option AA has the same L_{1eqv} variation as the new tuning design, Figure 4-13 (e) to (f) indicate that option AA has the bigger reactive load variation and is capacitive. The reason for this is that the primary is tuned at position A so the biggest variation of ΔX_l occurs at the minimum coupling position B which then results in the biggest additional reactive load compared with others. The analytical results in Figure 4-13 (e) to (f) also demonstrate that the new tuning design is inductive and has the least input DPF variation, which is between unity and 0.976. The SPICE simulation results of the new tuning design are shown in Figure 4-14 and again demonstrate very good agreement with the analytical results.
Chapter 4  Design Considerations for Variable Coupling Lumped Coil Systems

Figure 4-13: Analytical results of the primary LCL network with 700mm circular pad and parallel-tuned pick-up tuned at position AA, BB and the new tuning value versus δx at z of 100mm and 150mm
4.5.3. Inverter bridge current variation

As discussed in Section 4.4.4, the inverter bridge current is directly proportional to the coupling of the magnetic structure. In this design example, the mutual inductance of the charging pad varies from 23\( \mu \)H to 55\( \mu \)H. This implies that the inverter bridge current will also have same degree of variation.
The SPICE simulation results for the inverter bridge voltage and current operating under the rated load of 1.2kW with both series and parallel-tuned pick-ups are shown in Figure 4-15 (a) to (d). The tuning networks are designed using the proposed strategy of Section 4.4.5 so the reactive component in the bridge current has been minimised as much as possible. The simulation results show that the inverter bridge voltage varies from 90 to 200V RMS as the coupling changes while the inverter bridge current varies from 15 to 6A RMS. In order to lower the LCL network input voltage from 200V RMS to 90V RMS, the inverter bridge will necessarily be working over a wide variation of phase shifts $\theta$ (explained in Chapter 2) to achieve a wide $V_{B,1}$ variation. While the inverter bridge operates with small phase shifts to lower $V_{B,1}$, the current $I_{B,1}$ will be at its highest value, as illustrated in Figure 4-15. For higher power systems, in the range of 7kW, this wide variation of $I_{B,1}$ complicates the inverter bridge design and makes the semi-conductor switch selection difficult.

![Figure 4-15: SPICE simulation results of the fundamental component of the primary inverter bridge voltage $V_{B,1}$ and current $I_{B,1}$ with 700mm circular pad with series-tuned and parallel-tuned pick-up using the Developed tuning design versus $\delta x$ at z of 100mm and 150mm](image-url)
4.5.4. Alternative magnetic structure

The developed tuning network design strategy can be used on any magnetic structure. The key parameter, as mentioned in the circular pad example, is the boundary of movement or tolerance within the specified power transfer zone. The circular pad has rotational symmetry therefore the design example requires only lateral direction of movement with vertical movement to determine the tuning design. For the Double D charging pad presented in Chapter 3, the defined boundary of the power transfer zone is a rectangular prism as illustrated in Figure 4-16. This is due to its polarised structure. Therefore, in order to design the tuning network for systems using Double D pads, the required inductance measurements are along the boundary of the two square plans at the extremes of vertical offset. These two plans are A-B-C-D and E-F-G-H as indicated in Figure 4-16.

Preliminary work of applying the new tuning design strategy on the Double D pad, presented in Chapter 3, has been done and the design process is found to be identical to the circular pad design example. A significantly larger number of graphs are however required to performing the tuning design of the Double D pad due to its movement tolerance being a rectangular prism shown in Figure 4-16. However, presenting these graphs do not provide any new insight for understanding the developed tuning design methodology, thus this is not in the scope of this thesis.

![Figure 4-16: Conceptual diagram of Double D pad tolerance of movement](image-url)
4.6. Conclusion

In this chapter, the impact of mistuned primary and secondary charging pads on a primary power supply used to drive these lumped coils under variable coupling applications has been analysed. The pick-up tuning options investigated are the conventional series-tuned and parallel-tuned topologies, while the primary tuning network is a series-parallel LCL network with voltage source input. A number of design considerations needing to be addressed in the design of such systems have been presented. These are: overrating components for additional reactive load, increased track conduction loss, primary tuning design, and inverter bridge current variations.

A design methodology for minimising the reactive power at the input of the primary LCL network was presented. A 1.2kW design example using a pair of 700mm circular charging pads with variable coupling using the proposed tuning design strategy was presented. The proposed design was compared against two standard tuning options; designing the charging pad tuning at either the maximum or the minimum coupling position. The comparison demonstrated the effectiveness of the proposed design methodology by ensuring the reactive load on the primary LCL network was inductive, while maintaining the DPF to be near unity. The developed analysis was verified using SPICE simulations.
Chapter 5.

A Practical Pick-up Design Topology for Variable Coupling Lumped Coil System

5.1. Introduction

In the previous chapter the loading scenario on the power supply inverter bridge for a lumped coil system, with a single pick-up and a single power supply, under conditions of variable coupling was discussed. The power supply controls the track current on the primary side to compensate for variations in magnetic coupling and to control the pick-up output power. It was suggested that the increased burden on the variable current primary side operating at rated load with a wide range of magnetic coupling variations would complicate the power supply design and make the semi-conductor switch selections difficult. In practice, a controller on the secondary side is always required for pick-up protection purposes. Thus, it is convenient to use the same controller to perform partial power regulation on the secondary to ease the control required in the primary power supply. A pick-up controller that can dynamically regulate its output power with a variable coupled input power \( V_{dc}/V_{ac} \) is therefore needed for this situation. This chapter details the development of such a controller.

This chapter begins by examining the most common practical pick-up topologies and then discusses their strengths and weaknesses for use in variable coupling applications. Following this a multi-path pick-up which uses a combination of a series-tuned pick-up together with a slow switched LCL network with rectifier output is proposed in order to achieve good power factor (PF) over the practical operating ranges considered. The design approach of traditional LCL network with DC output rectifier (as detailed in [116, 123]) assumes an application with fixed magnetic coupling. Consequently a new design approach is described to accommodate the expected significant variations in coupling. A suitable design is then developed for a 2kW prototype system enabling experimental evaluation which
validates the presented design approach and demonstrates its performance over the intended operating range.

5.2. Concept and Design Methodology of Multi-Path Pick-up

5.2.1. Suitable pick-up controllers for EV battery charging application

The three most common practical pick-up control topologies, as discussed in Chapter 2, that are suitable for the desirable power level (1-2 kW) for domestic EV battery charging applications, are given in Table 5-1. These are the parallel LC tuned pick-up with boost switch mode converter, the series LC tuned pick-up with buck switch mode converter and the series-parallel LCL tuned pick-up with load decoupling controller.

<table>
<thead>
<tr>
<th>Pick-up Tuning</th>
<th>Controller</th>
<th>Switching Speed</th>
</tr>
</thead>
<tbody>
<tr>
<td>Series LC</td>
<td>Buck converter</td>
<td>Fast/Slow</td>
</tr>
<tr>
<td>Parallel LC</td>
<td>Boost converter</td>
<td>Fast/Slow</td>
</tr>
<tr>
<td>Series-parallel LCL</td>
<td>Load decoupling control</td>
<td>Slow</td>
</tr>
</tbody>
</table>

Table 5-1: Summarised table of most common practical pick-up controllers for outputting at 1-2kW

Both the parallel and the series LC tuned pick-ups with their switch mode converters can operate in either fast or slow switching modes. The series-parallel LCL tuned pick-up is limited to a slow switching load decoupling topology. The equations describing the power flow of these controllers were presented in Chapter 2. In a lumped coil system where the primary charging pad is identical with the secondary, the power flow of each system can be expressed in terms of the magnetic coupling coefficient and the charging pad inductance as shown in (5.1), (5.2) and (5.3) for series LC tuned, parallel LC tuned and series-parallel LCL tuned pick-up respectively.

\[
P_{\text{Series}} = k\omega L_i I_1 \frac{\pi}{2\sqrt{2}} \frac{V_{\text{DC}}}{R_{\text{DC}}} D \tag{5.1}
\]

\[
P_{\text{Parallel}} = kI_1 \frac{\pi}{2\sqrt{2}} V_{\text{DC}} (1 - D) \tag{5.2}
\]

\[
P_{\text{LCL}} = k\omega L_i I_1 \frac{2\sqrt{2}}{\pi} V_{\text{DC}} \frac{1}{X} (1 - D) \tag{5.3}
\]
Here X is the characteristic impedance of the LCL network, k is the magnetic coupling coefficient, and D is the switch duty cycle ratio.

When the pick-up controller is operating in a slow switching mode, the duty cycle D is either 0 or 1. In a variable coupling lumped coil system with one power supply and one pick-up, the adjustments to the primary track current I_1 are limited and cannot fully compensate for the increase in coupling factor k between the charging pads so that the instantaneous power drawn by the slow switching pick-up may overload the primary power supply. Therefore both the LC parallel-tuned and LC series-tuned pick-ups need to operate in a fast switching mode in order to dynamically regulate their output power in places where the variation in magnetic coupling cannot be compensated for solely on the primary power supply side.

However a slow switching scheme has a number of advantages including lower switching losses, simple controller design, and less stringent requirements for the semiconductor switch selection. A controller topology which adopts the advantages of a slow switching type, but also has the capability to adjust its output power dynamically would provide an alternative option which could be more favourable. A pick-up controller, named the multi-path pick-up, is proposed and presented in this chapter for a 2kW battery charging system with a pair of identical 420mm diameter circular charging pads and a unity power factor (UPF) single phase input power supply with minimal DC energy storage as described in [120, 156]. With the power supply having minimal DC energy storage, the track current is amplitude modulated (AM) and has an envelope at twice the utility frequency. For simplification the primary power supply is set to operate with constant track current while the pick-up controller is investigated with variable magnetic coupling between the primary charging pad and the secondary pick-up. However, for a complete practical system design, the variation of track current from the primary side control in conjunction with the variation in magnetic coupling will need to be considered together in the design of the pick-up controller but this is outside the scope of this thesis.

5.2.2. Basic structure of a multi-path pick-up controller

The basic structure of the proposed multi-path pick-up is shown conceptually in Figure 5-1 [36]. The pick-up inductor L_2 is fully series compensated by C_2 at the primary operating frequency. This way the voltage level that appears across the AC voltage bus (V_in) is essentially the same as the self induced open circuit voltage V_oc in L_2. Note that with a fixed primary track current, the pick-up mistuning due to variations in mutual inductance
between the primary and the secondary has an insignificant effect on the voltage level after the series tuning capacitor \( C_2 \). In places where a variable track current power supply is used, \( I_1 \) needs to be controlled according to the demand from the pick-up controller for a desirable voltage level of \( V_{\text{in}} \).

This internal AC voltage busbar connects to a number of LCL networks in parallel, with each active LCL network carrying a portion of the pick-up coil current \( I_2 \). A circuit diagram of the LCL network with rectifier and DC filter capacitor at the output is shown in Figure 5-2. As described in Chapter 2, the essential characteristic of an impedance converting LCL network is to convert a voltage source input into a current source output. Each LCL network has its own bridge rectifier and these bridge rectifiers are connected in parallel to deliver power to the load. With their current source outputs, each LCL network can be decoupled individually from the load using a shunt decoupling switch and will not affect the other networks [36]. Using the LCL networks with rectifiers in this multi-path pick-up application does however require adjustments to the decoupling switch and rectifier so that it can be practically implemented. These adjustments are explained in detail in Section 5.2.3.

![Figure 5-1: A conceptual diagram of a series-tuned pick-up with multi-path pick-up controller](image)

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Each LCL network can be made to be identical and designed to take a portion of the total output power, but because they operate independently from each other they can be switched on and off individually. Thus by controlling the number of active LCL networks based on the expected output power and a measurement of the available input power of the pick-up (as indicated by the AC bus voltage), the power transferred to the secondary can be regulated through finite steps. Although the total number of passive components required by the multi-path pick-up is greater than that used in a conventional IPT pick-up controller; each LCL network is only carrying a portion of the designed output power, so that the components are relatively small and the total installed VA of the entire circuit is the same.

5.2.3. Structures of a single LCL network channel

As mentioned in Chapter 2, using the LCL network in pick-up applications requires a full bridge rectifier and a slow switching load decoupling switch to regulate the power flow. To use the LCL network in the proposed multi-path pick-up structure, the ground potential between each active and inactive network needs to be kept the same as they have a common input voltage bus. Thus, a common load decoupling switch with a full bridge rectifier cannot be used here to ensure each network can be controlled individually. A common voltage doubler circuit, shown in Figure 5-3 (a), could be used for the rectifier stage as it has common ground for both input and output stage. As demonstrated in Figure 5-3 (b) and (c), the voltage doubler rectifier uses the pre-existing capacitor $C_4$ to store the DC offset ($1/2$ of $V_{DC}$) for the voltage doubling effect. With the switch $S$ turned on, the LCL network is decoupled from the output but the ground potential between the input voltage ($V_{in}$), the inactive network, and the output voltage ($V_{DC}$) is still the same as shown in Figure 5-3 (d). While the switch is on, the resonant current still circulates in $C_3$, $L_4$ and $C_5$. 

---

**Figure 5-2: A simplified circuit diagram of a single path LCL network and rectifier**
Figure 5-3: (a) A schematic diagram of the voltage doubler / current divider for a single LCL network. (b): Current $I_{L4}$ flows through $D_1$ to charge up $C_4$ at $\frac{1}{2} V_{DC}$. (c): $I_{L4}$ flows through $D_2$ to output voltage $V_{DC}$. (d) with switch $S$ turned on, LCL network is completed with capacitor $C_5$ and decoupled from output $V_{DC}$.

With this configuration the LCL network can be easily controlled with the same shunt decoupling switch. However, attention must be paid to the effect of a decoupled network on the pick-up. Ideally the network must draw zero reactive power on full load and also when decoupled. As explained in [116, 123], the combination of $L_4$ and $C_4$ is used to force continuous conduction and to compensate for the extra inductive loading caused by the rectifier. Thus, with the decoupling switch turned on, a capacitive load occurs on the output of the LCL network, which results in extra reactive load at the input of the LCL network. In order to maintain unity power factor at the input of each LCL network, an extra capacitor $C_5$, shown in Figure 5-3 (a) and (d), is used in series with the decoupling switch $S$ to form an ideally tuned LCL network when deactivated.

Alternatively a series AC switch control, shown in Figure 5-4 (a), could also be used to decouple the LCL network from the internal AC busbar. With the LCL network being decoupled by the series AC switch, the network is completely deactivated so there is no resonant current in capacitor $C_3$ and $L_4$. Compared with a normal shunt switch topology, the copper and iron losses in this LCL network are minimised when inactive. The series AC switch of this LCL network could be connected to a symmetrical voltage doubler rectifier, as shown in Figure 5-4 (b), or a full bridge rectifier since any inactive LCL network is decoupled from the internal busbar and other LCL networks.
Chapter 5  A Practical Pick-up Design Topology for Variable Coupling Lumped Coil System

Figure 5-4: (a) A LCL network with front end series AC switch and with a full bridge rectifier. (b) A symmetrical voltage doubler rectifier.

5.2.4. Design of a multi-path pick-up

The multi-path pick-up structure is a combination of a conventional series-tuned pick-up and multiple LCL networks connected in parallel for power transfer regulation. The fundamental equation describing the power transfer of a conventional series-tuned pick-up under ideal tuning condition is \( P = Q I V_{oc} I_{sc} \). As described earlier, the current in the pick-up coil inductor \( L_2 \) is the summation of the current in the first inductor \( L_3 \) of each LCL network. A feature of the LCL network as described in Chapter 2 is that the input current \( I_{L3} = \frac{2\sqrt{2} V_{DC}}{\pi X} \) is constant providing \( V_{DC} \) is controlled to be constant by suitable action of the switches. Thus the apparent current quality factor of the basic series-tuned multi-path pick-up controller can be described by:

\[
Q_I = \frac{n}{X} \frac{2\sqrt{2} V_{DC}}{\pi I_{sc}}
\]

\[
Q_I = \frac{n\omega L_2}{X} \frac{2\sqrt{2} V_{DC}}{\pi V_{in.Avg}}
\]

Here \( n \) is the number of active LCL networks, \( X \) is the characteristic impedance of each LCL network, \( \omega \) is the primary operating frequency, and \( V_{in.Avg} \) is the average RMS value of the internal AC busbar (\( \approx V_{oc} \)) as discussed later in Section 5.3.1.

Compared to a conventional LC series-tuned buck converter, which in practice needs current limiting circuitry to control \( Q_I \), the \( Q_I \) of the series-tuned multi-path pick-up is directly controlled and limited by \( n \). This feature limits the inrush current in the pick-up coil at start up, which is a significant problem in a conventional series-tuned pick-up with buck converter [137]. The feature also prevents the pick-up coil current from exceeding the series tuning
capacitor rating during unexpected heavy loading conditions. Using (5.4), the power transferred by the multipath pick-up with the UPF single phase input power supply in [120] is then described by (5.5). The first \( \left( \frac{2\sqrt{2}}{\pi} \right) \) factor comes from the operation of the power supply that causes the track current to be modulated at double the mains frequency [120]. This modulation directly affects the average output DC current and will be explained in Section 5.3.1. With a constant \( V_{in.Avg} \), the power output can be simply varied by changing the average value of \( n \).

\[
P = \frac{2\sqrt{2}}{\pi} V_{in.Avg} I_{ac} Q_l
\]

\[
= n \frac{8}{\pi^2} \frac{V_{in.Avg} V_{DC}}{X}
\]  

(5.5)

A 300V DC, 2 kW multi-path pick-up prototype was designed and implemented for a battery charging system. It uses a similar primary power supply to that described in [120], with the specifications of Table 5-2. The supply is used to provide power over a pair of identical 420 mm diameter circular charging pads as described briefly in Chapter 3. The measured coupling factor and inductances of these pads are shown in Figure 5-5 with the pads centred but separated over a range of vertical offsets from 40 to 100mm. The design range of the pick-up coupled voltage \( V_{oc} \), hence the \( V_{in.Avg} \), is set between 210V RMS and 110V RMS. A simple step by step approach to designing the multi-path pick-up is outlined following:

- For a given maximum output power and output DC voltage, calculate the ratio of \( X/n \) with the minimum input voltage \( V_{in.Avg} \) using (5.5).
- Determine the number of active LCL networks “\( n \)” corresponding to the minimum \( V_{in.Avg} \) for output rated load, and use the ratio of \( X/n \) obtained earlier to calculate the characteristic impedance \( X \) of the LCL network.
- With the given \( \omega \) and \( X \), the value of \( L_3 \) and \( C_3 \) is then calculated for each network.
- Calculate the value of \( L_4 \) and \( C_4 \) based on \( X \) and the ratio of \( V_{DC}/V_{in.RMS} \) using the proposed graphical design approach detailed later in Section 5.3.
- Determine the required VA rating of each of the LCL network resonant components when operating under rated load at maximum \( V_{in.Avg} \) as each network now outputs more power directly related to the increase of \( V_{in.Avg} \).
In the application discussed here, a decision was made to use a 5 and a $\frac{1}{2}$ multi-path pick-up network. The half powered network has a characteristic impedance of 2X (compared with a characteristic impedance of X for a normal LCL network), and is used to enable improved output power resolution and control, ie: it allows a 10% step change in output power instead of a 20% step. The resolution helps to prevent the instantaneous output power of the pick-up exceeding the rating of the primary power supply when implementing average output power control. This will be explained in detail in Section 5.4.1.2.

![Figure 5-5: Parameters of two perfectly aligned 420 mm circular charging pads against vertical offset from 40 to 100mm. (a) Primary pad inductance with secondary being open-circuited and short-circuited. (b) the $V_{oc}$ and $I_{sc}$ of the pick-up pad with $I_1 = 27$A at a frequency of 50 kHz. (c) the mutual inductance and coupling coefficient](image-url)
5.3. A Graphical Design Approach for Designing Matching Network with 
Variable $V_{DC}/V_{in}$

A unity power factor (UPF) design topology using an LCL network with rectifier and 
DC capacitors at the output was developed in [116, 122, 123]. The objectives of this design 
approach were to ensure the output rectifier operates in continuous conduction and the 
resultant PF at the input of the LCL network is near unity in order to minimise reactive power 
in the circuit. This design approach has been proven to be effective for a fixed ratio between 
the input voltage ($V_{in}=V_{oc}$) and the output DC voltage ($V_{DC}$). However, in the application 
presented in this chapter, the induced voltage has an amplitude modulated envelope at double 
mains frequency and the average RMS value of $V_{oc}$ can vary between 110 to 210V as a result 
of the variation in the air gap between the circular pads. In consequence UPF cannot be 
achieved as the reactance introduced by the rectification process varies with $V_{DC}/V_{in,RMS}$. 
Consequently, the power factor to the input of the LCL network also varies, resulting in a 
variable reactive load which is reflected back to the primary track.

This load creates higher conduction losses in the pick-up coil and may be problematic 
for the primary track tuning. Since the UPF design approach described in [116, 122, 123] is 
for a fixed $V_{DC}/V_{in,RMS}$, it is desirable to develop an alternative design which selects $X_{L4}$ and 
$X_{C4}$ to minimise the variation in the input PF while maintaining the rectifier in continuous 
conduction. Here, a graphical design methodology is presented along with a brief summary 
of amplitude modulated track current phenomena and the UPF LCL pick-up design methodology.

5.3.1. An amplitude modulated track current

The single phase mains input power supply used in this design has an amplitude 
modulated track current at twice the utility frequency [120, 156]. In consequence, both the 
induced voltage in the pick-up coil ($V_{oc}$), and hence the voltage across the pick-up internal 
busbar ($V_{in}$) also have an amplitude modulated envelope as shown conceptually in Figure 5-6. 
Here the normalised conceptual waveform of $V_{in}$, the RMS envelope $V_{in,RMS}$ and the average 
RMS $V_{in, Avg}$ are plotted against a half mains period. When the track current is amplitude 
modulated at double mains frequency, $V_{in}$ is expressed by:

$$V_{in} = V_{in} \sin \omega t \left| \sin \omega_{m} t \right|$$  \hspace{1cm} (5.6)
Here $\omega_m$ responds to the mains frequency and $\hat{V}_{in}$ is the peak value of $V_{in}$. The relationship between the average RMS ($V_{in,Avg}$), the peak RMS envelope ($\hat{V}_{in,RMS}$) and the peak value ($\hat{V}_{in}$) are illustrated in Figure 5-6 and are defined by:

$$V_{in,Avg} = \frac{\hat{V}_{in,RMS}}{\sqrt{2}} = \frac{\hat{V}_{in}}{2} \quad (5.7)$$

As $V_{in}$ has a sine weighted envelope at double the mains frequency (5.6), the current $I_{L4}$ in the output inductor also has a sine weighted envelope. Therefore, the average value of the output current $I_{DC}$ of each network can be expressed by [120, 156]:

$$I_{DC} = \frac{4}{\pi^2} \frac{\hat{V}_{in}}{X} = \frac{4\sqrt{2}}{\pi^2} \frac{\hat{V}_{in,RMS}}{X} = \frac{8}{\pi^2} \frac{V_{in,Avg}}{X} \quad (5.8)$$

5.3.2. UPF design methodology operating at the critical point of the rectifier continuous conduction

When using the LCL network topology in applications with a rectifier and a DC capacitor at its output, the current $I_{L4}$ in the output inductor $L_4$ is necessarily distorted during the rectification process. This introduces an additional inductive reactance ($X_{rec}$) in the circuit in series with $L_4$ and $C_4$ as illustrated in Figure 5-7 where the output rectifier and DC load is
replaced by its equivalent reactance and resistive AC load [112, 116]. The UPF design approach, described in [116, 122, 123], designs the value of $X_{L4}$ and $X_{C4}$ to force continuous conduction of the rectifier and to compensate for the difference between the total reactance of $X_{L4}$ and $X_{rec}$, and the LCL impedance $X$. Thus, the LCL pick-up can be modelled with an ideally tuned network, with a resistive output, which results in a UPF input.

![LCL circuit diagram](image)

**Figure 5-7: A LCL circuit diagram with the rectifier and DC output load being represented by its equivalent reactance ($X_{rec}$) and AC resistive load ($R_{AC}$)**

This LCL pick-up UPF design methodology, described in [116, 123], is summarised with the steps shown below.

- For a given input average RMS voltage ($V_{in,Avg}$), output DC voltage, and desired power level, the LCL network impedance $X$ can be calculated using (5.9).

$$P_{in} = \frac{8}{\pi^2} \frac{V_{DC}V_{in,Avg}}{X} \tag{5.9}$$

- With the given ratio of output DC voltage to the input RMS voltage, the minimum inductance value of $L_4$ to ensure continuous conduction of the rectifier can now be calculated using (5.10). The inequality sign indicates that providing the ratio of $X_{L4}/X$ is greater than the right hand side, continuous conduction mode is guaranteed.

$$\frac{X_{L4}}{X} \geq \frac{V_{DC}}{\sqrt{2}V_{in,RMS}} \tag{5.10}$$

- With the designed value of $X_{L4}$ operating at the critical point of conduction, the equivalent reactance of the rectifier ($X_{rec}$) is expressed by [116]:

$$X_{rec} = 0.7337 \left( \frac{V_{DC}}{V_{in,RMS}} \right)^2 X_{L4} - X_{L4} \tag{5.11}$$

- Capacitor $C_4$ is used to compensate the difference between $X$ and $X_{L4}$ together with the $X_{rec}$ at the designed input voltage. Using (5.10) and (5.11), $C_4$ is then given by:
\[ \alpha C_4 = \frac{1}{1.4674 \omega L_4 - X} \]  

(5.12)

Using (5.10) and (5.11), the normalised impedances of both \( X_{L4} \) and \( X_{rec} \) as a function of the \( V_{DC}/V_{in,RMS} \) ratio for a rectifier operating at the critical point of conduction are shown in Figure 5-8.

![Normalised Impedance of \( X_{L4} \) and \( X_{rec} \) versus \( V_{DC}/V_{in,RMS} \) at the point of critical conduction](image)

**Figure 5-8: Normalised impedance of \( X_{L4} \) and rectifier designed at the point of critical conduction**

An LCL network is here designed with the UPF methodology shown above but with a variable \( V_{DC}/V_{in,RMS} \) to illustrate its drawbacks under such a condition. The design specification and the designed LCL parameters are given in Table 5-3. As mentioned previously, the minimum input voltage \( V_{in} \) for the LCL network is 110V average RMS, which is 155V peak RMS due to the amplitude modulation effect described in [120]. The maximum \( V_{in} \) is 210V average RMS, which is 297V peak RMS. With the UPF design methodology, the LCL network is designed to operate at the critical continuous conduction point and to have unity power factor at the input of the LCL network at the given input voltage. A decision was made to use 100V RMS as the design value for the input voltage, which ensures that the rectifier will continue to operate in continuous conduction with larger values of \( V_{in} \). Using (5.5), the \( X/n \) ratio required to output 2kW with a 100V input is calculated to be 6.64Ω. With a total number of 5.5 multi-path networks, the calculated \( X \) value is 36.5Ω but an \( X \) value of 36.19Ω is chosen with available components. The designed LCL network was simulated in...
SPICE to investigate its behaviour with a variable input voltage. Simulated results of the LCL input power factor are shown in Figure 5-9.

<table>
<thead>
<tr>
<th>Design Specification</th>
<th>Designed UPF LCL Parameter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{in,Avg}$ (minimum average RMS value)</td>
<td>100V</td>
</tr>
<tr>
<td>$V_{in,RMS}$ (minimum peak RMS value)</td>
<td>141V</td>
</tr>
<tr>
<td>$V_{in,Avg}$ (maximum average RMS value)</td>
<td>210V</td>
</tr>
<tr>
<td>$V_{in,RMS}$ (maximum peak RMS value)</td>
<td>297V</td>
</tr>
<tr>
<td>$P_{out}$ with minimum $V_{in}$</td>
<td>370W</td>
</tr>
<tr>
<td>Frequency</td>
<td>50kHz</td>
</tr>
<tr>
<td>$V_{DC}$</td>
<td>150V DC</td>
</tr>
</tbody>
</table>

Table 5-3: Design specification and the designed parameters for the UPF LCL network

![LCL input Power Factor vs. $V_{DC}/V_{in,RMS}$](image)

Figure 5-9: Input power factor of an LCL network designed with UPF methodology versus variation of $V_{DC}/V_{in,RMS}$

From Figure 5-9, it can be seen that the LCL input power factor is near unity at the designed input $V_{in}$ of 100V RMS ($V_{DC}/V_{in}$ ratio of 1.5). However, with any $V_{in}$ greater than the designed value the input power factor drops rapidly, given the ratio $V_{DC}/V_{in,RMS}$ decreases. If $V_{in}$ decreases below the designed value, the power factor variation is relatively small but the pick-up rectifier will operate in a discontinuous mode which results in less power. As indicated in Figure 5-9, the input PF varies from unity to 0.75 with a variation of $V_{DC}/V_{in,RMS}$ ratio from 1.5 to 0.5 respectively. A poor input PF creates unnecessary VA stress on the components and the reactive load will reflect back to the primary track. Normally any
constant reflected reactive power from the pick-up is compensated by adjusting the track tuning during installation, however when the pick-up power factor varies during operation, it is difficult to dynamically retune the primary without a self-tuning compensator of the like discussed in [154].

5.3.3. The Graphical Design Methodology with variable $V_{DC}/V_{in}$

As shown above, in applications where the input voltage is known to vary a different approach is required when designing the value of $X_C$ in order to minimise the resulting variation of input PF of the LCL network.

To help distinguish between the variable coupling application design which is proposed here and the conventional UPF design of Section 5.3.2, the reactance of the rectifier, $L_4$ and $C_4$ designed for operating with variable coupling application will be referred to as $X_{rec,V}$, $X_{L4,V}$ and $X_{C4,V}$ respectively. To analytically design the LCL network to achieve minimum input PF variation is difficult, thus in this section a simple graphical design approach is presented. This methodology firstly determines a sensible value of $X_{L4,V}$ and then examines a range of $X_{C4,V}$ in order to select the combination of $X_{L4,V}$ and $X_{C4,V}$ that achieves the desired minimum variation in the LCL network input PF.

5.3.3.1. Selection of $X_{L4,V}$

The initial approach for choosing the characteristic impedance $X$ of the LCL network for a given output power is the same as described in (5.9). For a given variation range of the input voltage $V_{in}$, $X_{L4,V}$ needs to be carefully chosen to force continuous conduction at the minimum input voltage level, which is the highest $V_{DC}/V_{in,RMS}$ ratio with a constant $V_{DC}$ output. However, as $V_{in}$ has a sine weighted amplitude modulation envelope, the LCL network will be forced to work in discontinuous conduction for a fraction of the mains period as continuous conduction is practically impossible when the envelope of the RMS value of $V_{in}$ is near zero (or $V_{DC}/V_{in,RMS}$ ratio nears a singular point). With the rectifier operating in discontinuous conduction, the output power level reduces and estimating it for control and design purposes is difficult. Thus, instead of using the minimum average RMS value of the input voltage to choose $X_{L4,V}$, a new voltage level relative to the peak RMS of the amplitude modulated waveform is chosen to ensure the majority of the power over the modulation cycle can be transferred with the rectifier in continuous conduction.
The instantaneous output power level of the LCL network is proportional to the $V_{\text{in}}$ envelope. This is because the LCL input current has a near constant envelope due to its constant output DC voltage [120]. As illustrated earlier in Figure 5-6, 80% of the total output energy is transferred between time $T/10$ to $2T/5$, which is to 60% of the mains period. The other 40% of the mains period (outside the dashed line) contributes the remaining 20% of the output power. At times $T/10$ and $2T/5$, the value of $V_{\text{in}}$ on the normalised RMS envelope in Figure 5-6 is 0.6 of the peak RMS value of $V_{\text{in}}$ ($\bar{V}_{\text{in,RMS}}$). Therefore, $0.6\bar{V}_{\text{in,RMS}}$ of the minimum average RMS $V_{\text{in,Avg}}$ is chosen as the threshold to design $X_{L4,V}$, given this choice ensures the LCL network operates under continuous conduction over the period where it transfers the majority of its power. Thus, in an amplitude modulated track current application the required $X_{L4,V}$ at the minimum average RMS value of $V_{\text{in}}$ is now given by:

$$X_{L4,V} = \frac{X \cdot V_{\text{DC}}}{\sqrt{2} \cdot 0.6 \cdot \bar{V}_{\text{in,RMS}}}$$  (5.13)

Here $X_{L4,V}$ is the impedance of $L_{4,V}$ designed for variable coupling applications.

With a $\bar{V}_{\text{in,RMS}}$ of 141V and $V_{\text{DC}}$ of 150V, the calculated $X_{L4,V}$ is 1.25X. A value of 1.3X for $X_{L4,V}$ is chosen from the available components.

5.3.3.2. Selection of $X_{C4,V}$

As outlined in Section 5.3.2, the value of $X_{C4}$ is chosen to compensate for the difference between the total reactance of $X_{L4}$ and $X_{\text{rec}}$, and the LCL impedance $X$ to achieve UPF at the input to the LCL network [116]. As (5.12) is only suitable for designing $X_{C4}$ at one fixed $V_{\text{DC}}/V_{\text{in,RMS}}$ ratio, it cannot be used here.

It is difficult to derive an analytical expression for $X_{\text{rec,V}}$ to facilitate the design of $X_{C4,V}$ for an LCL network operating with a variable $V_{\text{DC}}/V_{\text{in,RMS}}$ ratio, particularly given the value of $X_{\text{rec,V}}$ varies with changes in $V_{\text{DC}}/V_{\text{in,RMS}}$ [116]. As such the input PF of the LCL pick-up cannot be unity under variable $V_{\text{DC}}/V_{\text{in,RMS}}$ without using self-tuning circuitry like that described in [153, 154]. Therefore, a graphical design approach based on SPICE simulation results is suggested here for determining the value of $X_{C4,V}$. Rather than trying to achieve UPF at the input of the LCL pick-up, the value of $X_{C4,V}$ is selected to minimise the variation in the input PF while keeping it as near to unity as possible within the designed input voltage range.
This approach firstly defines a range of $X_{C4,V}$ values required to perform a series of SPICE simulation of LCL networks using the calculated $X_{L4,V}$ from (5.13) for the given $V_{DC}/V_{in,RMS}$ ratio. To find this range, an approximated rectifier reactance ($X_{rec,V}$) with the designed $X_{L4,V}$ operating at the given extreme points of $V_{DC}/V_{in,RMS}$ is required. Although (5.11) does not directly indicate the relationship between $X_{rec}$ and $X_{L4}$, it does suggest that with a constant $V_{DC}/V_{in,RMS}$ ratio $X_{rec}$ reduces with increasing $X_{L4}$. For the purposes of estimating $X_{rec,V}$, an inversely proportional relationship between $X_{L4}$ and $X_{rec}$ is assumed and $X_{rec,V}$ can then be approximated by:

$$\frac{X_{rec,V}}{X_{L4,V}} \propto \frac{X_{L4}}{X_{L4,V}}$$

So that with a $V_{DC}/V_{in,RMS}$ ratio within the given range, the $X_{rec,V}$ of the circuit with the designed $X_{L4,V}$ using (5.13) can be approximately scaled using (5.14). The values of $X_{L4}$ and $X_{rec}$ used in (5.14) are designed so that the LCL network operates at its critical conduction point with the same $V_{DC}/V_{in,RMS}$ ratio. This estimated value of $X_{rec,V}$ is then used to help determine the range of values for $X_{C4,V}$ that are evaluated in the simulation.

Calculated values of $X_{L4}$ and $X_{rec}$ must operate the LCL network at the critical point of conduction at the two extreme $V_{DC}/V_{in,RMS}$ ratios of 0.5 and 1.5 given in Table 5-4. Based on these values and the previously designed $X_{L4,V}$ of 1.3X, the estimated $X_{rec,V}$ using (5.14) are 0.045X and 0.4X. Combining the estimated values of both $X_{rec,V}$ and $X_{L4,V}$, the range of $X_{C4,V}$ to be used for the simulations is then calculated to be 0.345X to 0.7X. As the values of $X_{rec,V}$ are approximated, a wider range of $X_{C4,V}$ (0.3X to 0.8X) was actually chosen for the simulation.
A Practical Pick-up Design Topology for Variable Coupling Lumped Coil System

Table 5-4: Values of $X_{L4}$ and $X_{rec}$ of an LCL network operating at the critical point of conduction and the values of designed $X_{L4,V}$ and the estimated range of $X_{C4,V}$ for simulation analysis.

<table>
<thead>
<tr>
<th>$V_{DC}/V_{in,RMS}$</th>
<th>0.5</th>
<th>1.5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Critical point of rectifier continuous conduction</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$X_{L4}$</td>
<td>0.354X</td>
<td>1.06X</td>
</tr>
<tr>
<td>$X_{rec}$</td>
<td>0.165X</td>
<td>0.496X</td>
</tr>
<tr>
<td>Graphical</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$X_{L4,V}$</td>
<td>1.3X</td>
<td>1.3X</td>
</tr>
<tr>
<td>$X_{rec,V}$</td>
<td>0.045X</td>
<td>0.4X</td>
</tr>
<tr>
<td>$X_{C4,V}$</td>
<td>0.345X</td>
<td>0.7X</td>
</tr>
<tr>
<td>$X_{C4,V}$ range</td>
<td>0.3X</td>
<td>0.8X</td>
</tr>
</tbody>
</table>

With the chosen range of $X_{C4,V}$ given in Table 5-4, normalised SPICE simulation results for a $X_{L4,V}$ of 1.3X are shown in Figure 5-10. From the LCL network input PF results shown in Figure 5-10 (a), it can be seen that between a $V_{DC}/V_{in,RMS}$ ratio of 1.5 (equivalent to $V_{in}$ of 100V RMS) to 0.5 (equivalent to $V_{in}$ of 297V peak RMS) the $X_{C4,V}$ value of 0.35X demonstrates the least PF variation and is close to unity. The PF simulation results with the chosen value of $X_{C4,V}$=0.35X is compared in Table 5-5 along with two other values of 0.3X and 0.4X.

The simulation results of LCL network input PF angle shown in Figure 5-10 (b) indicate that the input PF angle of the LCL network changes sign resulting in either inductive or capacitive loading with varying $V_{DC}/V_{in,RMS}$ ratio. The point at which the sign change occurs depends on both the $V_{DC}/V_{in,RMS}$ ratio and the selection of $X_{C4,V}$. As shown in Figure 5-10 (b), the non-unity PF operation at the higher $V_{DC}/V_{in,RMS}$ ratio before the change of sign corresponds to additional capacitive power at the input to the LCL network, whereas at lower $V_{DC}/V_{in}$ ratios this corresponds to additional inductive reactive power at the input to the LCL network. This is because with $X_{L4,V}$ fixed, the estimated $X_{rec,V}$ at higher $V_{DC}/V_{in,RMS}$ ratio is larger than at lower $V_{DC}/V_{in,RMS}$ ratios as noted in Table 5-4.
Figure 5-10: (a) Input PF and (b) input PF angle of LCL network with $X_{L4_V} = 1.3X$ versus $V_{DC}/V_{in,RMS}$ with $X_{C4_V} = 0.3X$ to $0.7X$

<table>
<thead>
<tr>
<th>$X_{C4_V}$</th>
<th>$V_{DC}/V_{in,RMS} = 0.5$</th>
<th>$V_{DC}/V_{in,RMS} = 1.5$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.3X</td>
<td>0.998</td>
<td>0.975</td>
</tr>
<tr>
<td>0.35X</td>
<td>0.999</td>
<td>0.9819</td>
</tr>
<tr>
<td>0.4X</td>
<td>0.99</td>
<td>0.987</td>
</tr>
</tbody>
</table>

Table 5-5: LCL network input power factor with three $X_{C4_V}$ value 0.3X, 0.35X and 0.4X
5.3.3.3. Performance of the designed LCL network using graphical approach

With the analysis presented earlier, the designed parameters of the LCL network are given in Table 5-6. Instead of using a $X_{C4,V}$ value of 0.35X, 0.3X was chosen due to the ease of implementation with available components. A comparison of the input power factor between the graphical design presented here and the normal UPF design [116], is shown in Figure 5-11. Over the $V_{DC}/V_{in,RMS}$ range of interest, the power factor using the graphical design example remains high, even well outside the interested $V_{DC}/V_{in,RMS}$ range. This is an outstanding improvement in performance.

<table>
<thead>
<tr>
<th>Designed LCL Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>LCL impedance X:</td>
</tr>
<tr>
<td>$L_3$:</td>
</tr>
<tr>
<td>$C_3$:</td>
</tr>
<tr>
<td>$X_{L4,V}$:</td>
</tr>
<tr>
<td>$L_{4,V}$:</td>
</tr>
<tr>
<td>$X_{C4,V}$:</td>
</tr>
<tr>
<td>$C_{L,V}$:</td>
</tr>
</tbody>
</table>

Table 5-6: Designed LCL network parameters using the graphical design approach

![Graphical LCL input Power Factor vs. $V_{DC}/V_{in,RMS}$](image)

Figure 5-11: SPICE simulation of LCL input PF versus $V_{DC}/V_{in,RMS}$ for graphical design example and UPF design example

The LCL network with its parameters shown in Table 5-6 has been tested practically under conditions where the input voltage is varied from 40V RMS to 290V RMS. The measured input power factor (PF) is shown in Figure 5-12 (a). The measured and simulated
PF results show good agreement, eg: a maximum difference of 0.03. A comparison of the normalised simulation results and practical measurements of the output DC current are shown in Figure 5-12 (b). These measurements are normalised against the ideal maximum output current given by [116, 123]:

$$I_{DC_{\max}} = \frac{2\sqrt{2} \frac{V_{in_{RMS}}}{\pi}}{X}$$  (5.15)

As explained in [116, 123] practical output DC current magnitudes are normally lower than the ideal value by 5 to 15% due to the distortion from the output rectifier in $I_{L4}$. If the input voltage varies from 40V to 290V RMS, the measured output current and simulation results are agreed and well matched with less than 3% difference over the particular range of interest (100V RMS to 290V RMS).

![Figure 5-12: Measured and simulated results of graphical designed example LCL network. (a): LCL network input PF versus $V_{DC}/V_{in_{RMS}}$. (b): Normalised $I_{DC}$ versus $V_{in_{RMS}}$.](image-url)
5.4. Control and Practical performance of a 2kW multi-path pick-up

5.4.1. Power flow control of multi-path pick-up

As discussed in the earlier section, the power supply used in this EV charging system has a single phase mains input with minimal DC storage, and consequentially an amplitude modulated primary track current at double mains frequency. With this power supply characteristic any transient distortion occurring in the pick-up will easily reflect into the mains input current, since there is only minimal DC storage in the primary power supply. The input mains power factor of the primary power supply with a conventional LC parallel-tuned fast switching boost mode controller operating at constant duty cycle is 0.94-0.95. Compared with a conventional slow switching controller, the fast switching controller limits the transient distortion occurring in the pick-up resonant circuit [120, 144, 156]. Thus, a slow switched system should ideally synchronise its switching to the mains input voltage waveform to minimise the harmonic distortion in the power supply mains current. Two control methods could be used with the multi-path pick-up in this application. These are referred to as “step modulation control” and “average power control” and are discussed in the following sections.

5.4.1.1. Step modulation control

The step modulation control method proposed here is to individually control the on/off time of each LCL network to form a pick-up coil current \( I_2 \) envelope similar to a multi-level inverter voltage waveform. The on/off time of each network is controlled over each mains half period to output the desired average output power. A conceptual diagram of the \( V_{in,RMS} \) and \( I_2 \) envelope using 4 LCL networks switching over a quarter of the mains period is shown in Figure 5-13.
When the pick-up is delivering maximum power, all LCL networks are on for the whole mains period to form a square wave envelope for $I_2$. When the available input power is greater than the rated load, the LCL networks start switching in order to control its output power. In consequence, the pick-up coil current $I_2$ now has a multi-level step envelope. A conceptual envelope waveform of $I_2$ with four-levels or steps is shown in Figure 5-13 along with the envelope of $V_{in,RMS}$. At the beginning of the mains period, either one of the full or the half power LCL network is always on. The remaining networks are gradually switched on in sequence from $\theta_2$ to $\theta_4$ as the mains phase nears $\pi/2$. Each network is then switched off in reverse order until the mains phase voltages reaches “$\pi$” to ensure the waveform has a symmetrical envelope. Using (5.5), (5.6) and (5.7) the output power of the step modulation control can be expressed in (5.16). The calculation averages the total energy transferred from each channel over a quarter of the mains period.

$$P = \frac{\sum_{i=1}^{n} \int_{\theta_i}^{\theta_i+\pi/2} 2\sqrt{2} V_{DC} \hat{V}_{in,RMS} \sin \theta d\theta}{\pi X / 2}$$

(5.16)

The SPICE simulation results of $V_{in}$ and $I_2$ for the designed multi-path pick-up with 4 LCL networks using step modulation control are shown in Figure 5-14. The output rectifier used in this design is the symmetrical voltage doubler rectifier with the shunt load decoupling switch from Figure 5-3. The timing for turning on these 4 LCL networks is 0, 0.186, 0.398,
and 0.726 radians as demonstrated in Figure 5-13. Although each LCL network is switching at a rate of 200 switchings/second (switching twice in each half mains period), with each LCL network switching individually the equivalent switching speed of the multi-path pick-up can be as high as 1200 switchings/second for a 5.5 network multi-path pick-up. The average RMS of $V_{in}$ in this simulation is 160V RMS and the output power is 1913.4W, which is only 2.5% off the theoretical value calculated from (5.16) of 1963W.

![Figure 5-14: SPICE simulation results of a 4 network’s multi-path pick-up with step modulation control. Top trace is the internal busbar voltage $V_{in}$. Bottom trace is the pick-up coil current $I_2$.](image)

### 5.4.1.2. Average power control

The average power control method takes advantage of the zero crossing points of the envelope of the amplitude modulated track current which is synchronised to the mains voltage. Here the multi-path pick-up switches the half network in or out to regulate the average output power. As such it would be ideal to switch this network synchronously with the utility zero crossings to minimise switching loss and waveform distortion in the overall system. With a simple envelope detection circuit, the switches in the designed controller are operated slowly at the zero crossing of the modulated track current. By switching the half network in and out every full cycle of mains, the desirable average output power is achieved across two or more utility periods. This minimises the distortion on the IPT power supply.
mains input current and maintains a good input power factor. However, a disadvantage of this method is that the peak power drawn by the pick-up is slightly higher than the rated power. Thus, care must be taken to make sure this over rated power drawn by the pick-up is within the primary power supply capability.

A simple graphic, shown in Figure 5-15, is used to assist the determination of the average number of networks required for the rated output power with the given range of input voltage variation. At the minimum input voltage ($V_{in,Avg} = 110V$ RMS), all 5.5 networks are needed to be on to deliver the required output power. As the input voltage drops below the designed threshold, the pick-up controller is unable to deliver rated power. As $V_{in}$ increases, the number of networks may be reduced in half steps as shown in Figure 5-15 to counter the increase in the available input power. The dashed line is the controlled peak output power of the pick-up with variations in $V_{in,Avg}$, while the dotted line is the pick-up output power with one half network less than the dashed line. It is important to make sure the peak of the dashed line does not exceed the maximum power capability of the IPT power supply. With the half network step the peak output power is 1.16 per unit (pu), which corresponds to 2.3kW (here 3 and a half networks are on and are driven from an internal bus voltage of 198V average RMS). The instantaneous output power cannot be kept at the nominal 2 kW level. Instead the average power is controlled to be constant over a couple of the utility cycles, using a similar hysteresis control technique to that presented in [35] by controlling the number of active networks that are switched on. This number can be determined graphically from Figure 5-15 by considering the value in between the dashed and dotted lines. For example, with an input voltage of 160V average RMS the number of active networks will vary between 3.5 (at position A, output power = 0.94 pu) and 4 (at position B, output power = 1.07 pu) to give the averaged desired output power level.
Chapter 5  A Practical Pick-up Design Topology for Variable Coupling Lumped Coil System

5.4.2. Practical Results and Discussion

A prototype of the 5.5 network multi-path pick-up designed earlier was built and is shown in Figure 5-16. The decoupling switch used in this prototype uses a series AC switch and the symmetrical voltage doubler rectifier as presented in Figure 5-4. This prototype was constructed for experimental purposes and as such the physical size was not optimised. Because each network is only taking a small portion of the output power, it could be designed with surface mount semiconductors, and set up for machine placement manufacturing. The power control topology implemented in this prototype uses the average power control. It was chosen based on its operational simplicity and low loss, in that all switching instants occur at zero current and voltage ensuring minimum distortion in the overall system.
The multi-path pick-up has been tested on the proposed 2 kW IPT EV battery charger system with the designed $V_{\text{in, Avg}}$ variation from 100V RMS to 210V RMS. A Clarke-Hess power meter model 2335 was used to measure the power being transferred at the pick-up internal busbar to monitor the efficiency of the multi-path pick-up controller prototype. By measuring the power this way, the conduction loss in the circular charging pads is ignored and the efficiency measurements are only comprised of the losses in the multi-path pick-up controller. Oscilloscope captures of the EV battery charging system outputting 2 kW’s are shown in Figure 5-17. The pick-up efficiency for the given range of $V_{\text{in, Avg}}$ variation is shown in Figure 5-18, and the measured primary power supply input mains PF is shown in Figure 5-19.
Figure 5-17: Oscilloscope captures of a 5.5 network multi-path pick-up operating at 2 kW. Top and bottom traces of (a), (c) and (e) are AC busbar $V_{in}$ and $I_2$ respectively. Top and bottom traces of (b), (d) and (f) are the mains voltage and input mains current of the power supply respectively. (a) and (b) operating with $V_{in,Avg} = 113V$. (c) and (d) operating with $V_{in,Avg} = 136V$. (e) and (f) operating with $V_{in,Avg} = 207V$. 

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In Figure 5-17 (a) and (b) all 5.5 networks are on to deliver 2kW at the minimum \( V_{in,Avg} \) of 110V. From Figure 5-17 (c) and (e), it can be seen that the LCL network is controlled on and off in order to control the averaged output power to 2 kW. Because each of the LCL networks are switched on and off at the zero mains voltage, this switching action does not result in any observable transient in the mains current shown in Figure 5-17 (d) and (f). As a result, the mains input power factor is kept within 0.91 to 0.95 across all output
power ranges as presented in Figure 5-19. The maximum power efficiency and the part load efficiency of the multi-path pick-up, shown in Figure 5-18, is around 95 to 96%, while the low power efficiency is above 90% for all practical ranges of power output.

A breakdown of the losses in the multi-path pick-up controller outputting 2kW with an input voltage of 210V RMS is provided in Table 5-7. The loss in each component was calculated based on the measured waveforms and device characteristics provided from the datasheets. With the average power control technique, the multi-path pick-up has essentially zero switching loss in the series AC switches, but it necessary has a continuous conduction loss. The majority of the identified losses came from the AC series switch conduction loss and the output diode rectifier conduction loss. These losses are 2.1% of the output power 2kW. As mentioned in Section 5.2.2, despite the number of required passive components are more than a conventional IPT pick-up, the total VA of the active LCL networks is the same. Thus, the total conduction losses in the active passive components are also similar to a conventional IPT pick-up and it is only 0.45% of the rated output power. The rest of the 2.03% loss which are not specifically identified are a combination of the losses in the output DC filter capacitor, the mechanical joints in the circuitry and conduction losses in the printed circuit board.

<table>
<thead>
<tr>
<th>Item</th>
<th>Loss</th>
<th>%</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductor L3</td>
<td>1.6 W</td>
<td>0.08%</td>
</tr>
<tr>
<td>Inductor L4</td>
<td>6.0 W</td>
<td>0.30%</td>
</tr>
<tr>
<td>Capacitor C3</td>
<td>0.7 W</td>
<td>0.04%</td>
</tr>
<tr>
<td>Capacitor C4</td>
<td>0.6 W</td>
<td>0.03%</td>
</tr>
<tr>
<td>IGBT conduction loss</td>
<td>25.3 W</td>
<td>1.27%</td>
</tr>
<tr>
<td>Diode conduction loss</td>
<td>16.6 W</td>
<td>0.83%</td>
</tr>
<tr>
<td>Mechanical joint, DC capacitor and unknown</td>
<td>40.4 W</td>
<td>2.03%</td>
</tr>
<tr>
<td>Total Loss</td>
<td>91.2 W</td>
<td>4.58%</td>
</tr>
</tbody>
</table>

Table 5-7: Breakdown of calculated losses in the multi-path pick-up controller outputting 2kW with $V_{in, Avg} = 210V$

5.5. Conclusion

This chapter has proposed and presented the implementation of a slow switching pick-up controller called a multi-path pick-up controller for a domestic single phase input EV battery charging system. This multi-path pick-up uses a series-tuned LC pick-up with multiple LCL networks to regulate its output power. In order to use the LCL network with a
rectifier to produce a controlled DC output operating off of a variable input voltage, a design process using normalised SPICE simulations was used to modify the LCL network parameters. This is to minimise the additional reactive power present in the circuit resulting from expected variations in the pick-up open circuit voltage. Two methods for controlling the multi-path pick-up’s output power were discussed. These were entitled “step-modulation” and “average power” control, both of which were shown to be capable of dynamically regulating the output power with a variable input voltage while maintaining high mains input PF in the primary power supply. The average power control method was implemented because of its simplicity, low switching losses and minimum transient distortion in the overall system.

A 2kW 5.5 network multi-path pick-up prototype was presented. It is designed to regulate its output power with variations in the pick-up open circuit voltage of 110 to 210V average RMS. Under operation this multi-path pick-up achieved efficiencies of 96% at 2kW and above 90% for all output power ranges while maintaining a good mains input power factor in the primary power supply.
Chapter 6.

A Circulating Current Controller for Series-Parallel LCL pick-ups

6.1. Introduction

In an IPT system a secondary pick-up with a fast switching controller provides the ability to continuously vary and regulate the output power, as discussed in Chapter 5. This feature is not only desirable in lumped coil systems with variable coupling but is also advantageous for a multiple pick-up distributed track system where the primary power supply can be rated for the designed output power of all the pick-ups, instead of the total maximum power drawn by slow switching pick-ups while switched on. This is a key advantage of using fast switching controllers. However, the conventional LC parallel and series tuned fast switching controlled pick-ups are both hard-switched, which limits their efficiency and complicates the semi-conductor switch selection for power levels suitable for EV battery charging.

Recently a novel pick-up controller, named the AC processing pick-up, was investigated. It achieves zero voltage switching on both turn-on and turn-off. However, it was found to be impractical for applications with DC output voltages in the range of 200-400V when charging EVs. This is due to the voltage stress on the semi-conductor components during partial loading conditions when the peak resonant voltage is not limited by its input and regulated output voltages, and can increase without a sensible limit [157].

In this chapter, a “circulating current” fast switching controller for series-parallel LCL tuned pick-ups [158, 159] is described using the unity power factor (UPF) LCL pick-up methodology presented in [116, 123]. This controller adopts characteristics such as soft-
switching of the switches and diodes during either turn on or turn off, while the components’ voltage and current stresses are limited by the input and output voltages.

A steady state AC analysis of the proposed controller is presented. To model the characteristics of this topology, both a linear approximation and a polynomial approximation for the phase difference between the AC output voltage and current is used. This analysis is supported with simulation results and experimental measurements. A design example of a 1.5kW 300V DC EV battery charging system prototype is presented and includes measurements of both its efficiency and reflected impedance under operation, along with the expected operational waveforms.

6.2. Fundamentals of Circulating Current Duty Cycle Control

A feature of the series-parallel tuned LCL network is that a constant input voltage ($V_{oc}$) will produce a constant resonant current in the output inductor fixed by the LCL network characteristic impedance $X$. Thus, for a slow switching LCL pick-up with a load decoupling switch, the same resonant current circulates through the output inductor and diode bridge rectifier whether the load decoupling switch is on or off. Therefore, by switching the rectifier synchronously with the resonant current to control the portion of the resonant current that is rectified, the pick-up output power can be continuously controlled [158, 159].

The proposed circulating current control circuit with a unity power factor (UPF) LCL pick-up, shown in Figure 6-1, comprises four diodes in a diode bridge configuration and two switches ($S_1$ and $S_2$) in parallel with the bottom two diodes ($D_3$ and $D_4$) [158, 159]. Closing switches $S_1$ and $S_2$ causes the resonant current ($I_{L3}$) in $L_3$ to recirculate in the resonant network through $S_1$ and $D_4$ or $S_2$ and $D_3$ for positive and negative periods of $I_{L3}$ respectively. In practice, the bottom two shunted diodes of the diode bridge may comprise the body diodes of MOSFET’s or IGBT’s forming the switches $S_1$ and $S_2$. $V_{g1}$ and $V_{g2}$ are the PWM gate signals which drive $S_1$ and $S_2$, and are synchronized with $I_{L3}$. This circuit may be operated in two modes referred to as Mode I and II [158, 159].
6.2.1. Mode I operation

In Mode I, D₁ or D₂ is allowed to conduct at the beginning of either the positive or negative period of I₃. S₁ or S₂ is then turned on after conduction to clamp part of the rectified I₃ to regulate the output power. A conceptual single cycle operating waveform for Mode I is shown in Figure 6-2 when D₁ starts conducting at the beginning of the positive period of I₃ at t₀ [158, 159]. Figure 6-2 shows a sequence of the circuit’s operation. The corresponding circuit diagram detailing each of these steps is given in Figure 6-3.

In Mode I, the rising edge of the gate signals and the duty cycle of V₁ and V₂ are controlled by the diode conduction interval (θ₁) referenced to the respective negative-to-positive and positive-to-negative zero crossings of I₃ (shown in Figure 6-2). The sequence of Mode I operation is described by the following steps:

- At t₀, shown in Figure 6-3 (a), current I₃ has just turned positive. With S₁ being held in the off state, D₁ starts to conduct. Current I₃ is then transferred to the load (R_DC) via D₁ and D₄ for a predetermined portion θ₁ of the positive period of I₃. The instantaneous output voltage of the LCL network (V_LCL) during the period t₀ to t₁ is equal to +V_DC (where V_DC is held constant throughout the operation under the assumption of a large DC capacitor C_DC or battery).
- At t₁ (when θ₁ is reached), shown in Figure 6-3 (b), S₁ is turned on and I₃ circulates through S₁ and D₄ in order to stop power being transferred to the load for the remaining portion of the positive period of I₃. This remaining time is called the switch conduction interval (θ₂).
- At t₂ = T/2, shown in Figure 6-3 (c), I₃ turns negative so that D₂ turns on with a duration of θ₁ and the circuit is completed through D₃. While D₃ is conducting, the shunt switch S₁ can now be turned off with zero current. The instantaneous output
voltage of the LCL network \( V_{\text{LCL}} \) is equal to \(-V_{\text{DC}}\) between \( t_2 \) and \( t_3 \). The power transferred to the load is identical in the second half cycle providing the diode conduction interval \( \theta_1 \) is maintained constant.

- At \( t_3 \), shown in Figure 6-3 (d), the gate signal \( V_{g2} \) turns on \( S_2 \) to keep \( I_{L3} \) circulating through \( S_2 \) and \( D_3 \). The gate signal \( V_{g1} \) can be applied to turn the shunt switch \( S_1 \) off any time between \( T/2 \) and \( T \) as \( I_{L3} \) circulates through \( D_3 \).

With switches \( S_1 \) and \( S_2 \) controlled such that they are synchronised with the zero crossing of \( I_{L3} \), the output current \( (I_{D1} + I_{D2}) \) is essentially a rectified chopped quasi sine wave. By controlling the diode conduction interval \( \theta_1 \) or equivalently the switch conduction interval \( \theta_2 \), the average output current can be directly and smoothly controlled. The diode and switch conduction intervals are related by:

\[
\theta_2 = \pi - \theta_1
\]  

(6.1)
Figure 6-2: Typical single cycle operating waveform of the circulating current controller operating in Mode I
Chapter 6

A Circulating Current Controller for Series-Parallel LCL pick-up

6.2.2. Mode II operation

Mode II operation is very similar to Mode I, but with a different switching sequence. The single cycle operating waveform of Mode II is shown in Figure 6-4 [158, 159]. In Mode II, S₁ or S₂ is made to conduct at the beginning of the positive or the negative period of $I_{L3}$ after which the switch is turned off to allow part of $I_{L3}$ to be rectified and transferred to the load. The falling edge of the gate signals $V_{g1}$ and $V_{g2}$ and thus the diode conduction interval $\theta_1$ are controlled with a phase delay $\theta_2$ referenced to $I_{L3}$ as shown in Figure 6-4. Before current $I_{L3}$ turns positive at $t_0$, diode D₃ is already turned on with $I_{L3}$ flowing in the negative direction. Thus, turning on S₁ during the negative period of $I_{L3}$ will result in zero current/zero voltage switching. The sequence of Mode II operation is described by the following steps:

- At $t_0$, shown in Figure 6-5 (a), $I_{L3}$ turns positive. Switch S₁ is turned on and $I_{L3}$ is forced to circulate through S₁ and D₄. No power is transferred to the load $R_{DC}$ for this portion of the positive period of $I_{L3}$ (equivalent to the period of $t_1$-$t_2$ in Mode I operation)
- At $t_1$ (when the switch conduction interval $\theta_2$ is reached), shown in Figure 6-5 (b), S₁ is turned off and $I_{L3}$ circulates through D₁ and D₄ to transfer power to the load. During
the diode conduction interval $\theta_1 (t_1 - t_2)$, the instantaneous LCL output voltage ($V_{LCL}$) is $+V_{DC}$. At any time between $t_0$ to $T/2$, $V_{g2}$ can be used to turn on $S_2$ with zero current and zero voltage switching-like $S_1$.

- At $t_2 = T/2$, $I_{L3}$ turns negative so that $D_1$ turns off softly and $I_{L3}$ recirculates through $S_2$ and $D_3$.
- At $t_3$, the gate signal $V_{g2}$ turns off $S_2$ to allow $I_{L3}$ to transfer power to the load through $D_2$ for the predetermined diode conduction interval $\theta_1$. The value of the LCL instantaneous output voltage during the period $t_3 - T$ is $-V_{DC}$.

With $\theta_1$ and $\theta_2$ maintained equal, the power transferred to the load over each half cycle is identical. By controlling the switch conduction interval $\theta_2$ of both switches $S_1$ and $S_2$ synchronously with the zero crossings of $I_{L3}$, on the AC side of the circuit both the phase and the fundamental component of $V_{LCL}$ (indirectly seen by the fundamental component of $I_{L3}$) change concurrently. As a result of this variation in magnitude and phase, the output power is controlled. This will be explained in detail in the next section.

On the DC side of the circuit the average rectified current is directly and incrementally controlled by controlling $\theta_2$ for both $S_1$ and $S_2$. Therefore, similar to other fast switching pick-up controllers the proposed topology for an LCL pick-up is capable of varying its output power continuously to give a variable coupled power from the pick-up coil. Similar to the suggested operation for a slow switching parallel LC tuned pick-up in [144], the proposed control method could also be used in a slow switching LCL pick-up to limit any transient overshoot occurring in the system. Such control may be achieved by limiting the rate of change of the diode conduction interval from zero to full conduction when turning on the circuit or vice versa when turning it off - but this approach and any subsequent investigation is outside the scope of this thesis.

A characteristic of the proposed controller is that the maximum current $I_{L3}$ that circulates through $L_3$, the switches and diode rectifier, depends on the coupled voltage $V_{oc}$ and the designed secondary current $Q_l$ and its RMS value is given by [116, 123]:

$$I_{L3} \approx \frac{V_{oc}}{X} = Q_l I_{sc} \quad (6.2)$$

Here $I_{sc}$ is the short circuit current of the pick-up coil $L_2$ and $Q_l$ is the pick-up current $Q$ which is given by [16, 35, 108, 123]:
Thus the current rating of each of these components needs to be designed under the maximum coupled voltage condition. However, the voltage stress on the semiconductor switches and diodes is limited to the output DC voltage despite the coupling condition.

\[
Q_i = \frac{\omega L_2}{X} = \frac{L_2}{L_2 - \frac{1}{\omega^2 C_{L2}}} = \frac{C_2}{C_{L2}} + 1 \quad (6.3)
\]

Figure 6-4: Typical single cycle operating waveform of circulating current controller operating in Mode II
6.3. Steady State AC Analysis with UPF LCL Pick-Up

To gain insight on the operation of the circulating current controller, a steady state AC analysis is presented in this section. As demonstrated in Section 6.2, the phase of $V_{LCL}$ relative to $I_{L3}$ changes during operation. This will necessarily introduce additional reactive loading in the circuit which will reflect back to the primary power supply. Most fixed frequency power supplies have a limitation on the track inductance variation, therefore it is important to understand the reflected impedance characteristic of this new circulating current controller.

6.3.1. Circuit operation model using fundamental components

In order to simplify the real and reactive power flow behaviour analysis of the circulating current controller, the output diode rectifier and the shunt switches are initially replaced by an equivalent controllable output voltage source. As described in the previous section, with either $D_1$ or $D_2$ conducting, the output voltage of the LCL network is $\pm V_{DC}$ and with either $S_1$ or $S_2$ turned on, the output voltage is zero. Thus, with the assumption that the rectifier of the LCL network is always operating in continuous conduction while the shunt switches are off, the output voltage can be modelled as a variable width square wave voltage,
with its on time being defined as $\theta_1$, as illustrated conceptually in Figure 6-6. This is similar to a conventional single phase inverter bridge voltage. The magnitude of the fundamental component of this voltage waveform is well known and is given by [115, 160]:

$$V_{LCL,1} = \frac{2\sqrt{2}}{\pi} V_{DC} \sin \left( \frac{\theta_1}{2} \right)$$  \hspace{1cm} (6.4)

Figure 6-6: A symmetrical variable width single-phase inverter bridge output voltage waveform.

The LCL pick-up circuit with circulating current control can now be modelled with two voltage sources as shown in Figure 6-7.

Figure 6-7: A circulating current control LCL pick-up being modelled with a controllable output voltage source

As mentioned in Chapter 2 and in [116, 122, 123], the impedance combination of each branch of a series-parallel LCL network is designed for the characteristic impedance $X$ at the designed operating frequency. However as described in Chapter 5, an inductive reactance ($X_{rec}$) is introduced by the $I_{L3}$ waveform which is distorted during the rectification process, thereby introducing a phase delay between the fundamental component of $I_{L3}$ ($I_{L3,1}$) and the fundamental component of $V_{LCL}$ ($V_{LCL,1}$) [116, 123]. $X_{rec}$ can be compensated by ensuring that the total series reactance of $L_3$ and $C_3$ ($jX_{L3} - jX_{C3}$) is equal to $j(X - X_{rec})$. This ensures that the input voltage $V_{oc}$ continually sees an equivalent pure LCL tuned circuit with an output resistive load which results in a unity power factor (UPF) input. The equations for designing the value of $L_3$ and $C_3$ were presented in Chapter 5 and are shown here again in (6.5) and (6.6) for convenience [116, 123].
\[ \frac{\omega L_3}{X} = \frac{V_{dc}}{\sqrt{2}V_{oc}} \] \hspace{1cm} (6.5)

\[ \omega C_3 = \frac{1}{1.4674\omega L_3 - X} \] \hspace{1cm} (6.6)

Here (6.5) is used to design the LCL pick-up to operate at the critical point of continuous conduction and (6.6) is used to compensate the combined additional reactance of \( X_{L3} \) and \( X_{rec} \) to ensure UPF at the input of the LCL network.

The real and reactive power flow in an LCL network can be modelled adequately by only considering the fundamental components of the output voltage \( V_{LCL} \) and the circulating current \( I_{L3} \), while ignoring the harmonic components in \( I_{L3} \). Therefore, the equivalent circuit for Figure 6-7, replacing the square wave output voltage \( V_{LCL} \) by its equivalent fundamental component \( V_{LCL, 1} \) is used for the analysis and illustrated in Figure 6-8. Each branch of the tuned LCL pick-up is represented by its characteristic impedance \( X \). An additional reactance component \((-j0.4674X_{L3}) \) in series with the output voltage source \( V_{LCL, 1} \) is used to represent the total impedance mismatch of \( X_{L3} \) and \( X_{C3} \) as indicated in Figure 6-8.

As the LCL pick-up is designed to achieve UPF at the input of the network, it is easier to model the output loading impedance \((R_{load} + jX_{load})\) generated from the circulating current controller at the output of the (modelled) ideally tuned LCL network and in front of the mismatched reactance component \((-j0.4674X_{L3})\). Under these circumstances, the modelled \( R_{load} \) and \( X_{load} \) is the output impedance seen by the ideal LCL network and is used to describe the input impedance \((Z_2)\) seen by the pick-up open-circuit voltage \( V_{oc} \) which will be discussed in Section 6.3.3.2.

**Figure 6-8:** Equivalent circuit diagram of an unity power factor LCL pick-up by using its fundamental components
One characteristic of a tuned LCL network is that the magnitude of the fundamental component of the output current is controlled by its input voltage and vice versa [117, 119]. This can be demonstrated by analysing the fundamental component of the output current \( I_{L3,1} \) using the superposition principle with a Norton transformation of the input voltage \( V_{oc} \) and the fundamental output voltage \( V_{LCL,1} \) as shown in Figure 6-9. Two current components, \( I_{L3a} \) and \( I_{L3b} \), are defined here for applying the principle of superposition to the analysis of \( I_{L3} \) and \( I_{L3,1} \). \( I_{L3a} \) is sourced from \( V_{oc} \) and \( I_{L3b} \) (and its fundamental component \( I_{L3b,1} \)) is sourced from \( V_{LCL} \).

![Equivalent circuit of an LCL pick-up using superposition principle.](a) with respect to \( V_{oc} \), (b) with respect to \( V_{LCL,1} \)

As shown in Figure 6-9 (b), the impedance of the LCL network seen by the fundamental component of \( V_{LCL} \) (\( V_{LCL,1} \)) is equivalent to an open-circuit when the LCL network is tuned at the operating frequency, so that the current component \( I_{L3b,1} \) is essentially zero. Thus \( I_{L3,1} \) consists of \( I_{L3a} \) only. As demonstrated by the Norton equivalent circuit shown in Figure 6-9 (a), the magnitude and the phase (referenced to \( V_{oc} \)) of the current component \( I_{L3a} \) and essentially the fundamental component \( I_{L3,1} \), is given by:

\[
I_{L3a} = I_{L3,1} = \frac{V_{oc}}{j \omega L_2 + \frac{1}{j \omega C_{L2}}} = -j \frac{V_{oc}}{X} = -j Q_1 I_{sc} \tag{6.7}
\]

### 6.3.2. A linear approximation of the relative phase variation of \( V_{LCL,1} \)

As illustrated previously in Figure 6-2 and Figure 6-4, both the relative phase and the magnitude of \( V_{LCL,1} \) vary simultaneously with changes in the diode conduction interval \( \theta_1 \). Therefore, for modelling the real and reactive power flow in the pick-up it is necessary to have an expression for the phase variation of \( V_{LCL,1} \) relative to \( I_{L3,1} \) as a function of \( \theta_1 \) and this can be approximated by first considering the extremes of operation where \( \theta_1 \) is either \( \pi \) or \( 0 \).
With the circuit operating at $\theta_1 = \pi$, the circuit operation is identical to a conventional diode bridge rectifier as mentioned in Section 6.3.1 and the load reactance $X_{\text{load}}$ is zero when operating in either Mode I or Mode II. Therefore the phase difference ($\alpha$) between $V_{\text{LCL-1}}$ and $I_{L3-1}$ with $\theta_1 = \pi$ in both Mode I and Mode II results in an inductive load which is compensated by the mismatched reactance component (-j0.4674$X_{L3}$). This reactive load is then given by:

$$0.4674X_{L3}I_{L3-1}^2 = V_{\text{LCL-1}}I_{L3-1}\sin\alpha_0$$

(6.8)

Using (6.4), (6.5), (6.7) and (6.8), this phase difference $\alpha_0$ is:

$$\alpha_0 = \sin^{-1}\left(\frac{0.4674\pi}{4}\right) = 21.537^\circ = 0.3759\text{rad}$$

(6.9)

When $\theta_1$ is close to 0 the circuit is essentially short circuited, despite the fact that the magnitude of $V_{\text{LCL-1}}$ is near zero in both Modes. In Mode I $V_{\text{LCL-1}}$ is leading $I_{L3-1}$ by $\pi/2$ while in Mode II $V_{\text{LCL-1}}$ is lagging $I_{L3-1}$ by $\pi/2$.

Therefore, for $\theta_1$ varying between $\pi$ and 0, the $V_{\text{LCL-1}}$ phase relative to $I_{L3-1}$ is varying between $\alpha_0$ (0.3759rad) to $\pi/2$ for Mode I and between $\alpha_0$ (0.3759rad) to $-\pi/2$ for Mode II operation. This is illustrated in the phasor diagrams in Figure 6-10 (a) and (b) for Mode I and Mode II respectively.
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Figure 6-10: Conceptual phasor diagram of $V_{oc}$, $V_{LCL_1}$ and $I_{L3_1}$ (a) Mode I, (b) Mode II. Note: the length of the phasor here does not accurately represent the magnitude of the corresponding component.

As explained in the previous section, the fundamental component ($I_{L3_1}$) of $I_{L3}$ is constant and is the dominant component of $I_{L3}$ regardless of the magnitude of its harmonic components generated by $V_{LCL}$. Therefore, across the entire operation range the phase difference between the zero crossing of $I_{L3_1}$ and $I_{L3}$, which controls the voltage transition of $V_{LCL}$, is small. This phase difference is not linear due to the harmonic components in $I_{L3}$.

However, as indicated in (6.9) and in the phasor diagrams shown in Figure 6-10, when the controller operates at either full diode conduction, or zero diode conduction, $\alpha$ is fixed regardless of the circuit parameters of the UPF LCL pick-up. A UPF LCL pick-up, designed with a $V_{DC}/V_{oc}$ ratio of 1 and a $Q_l$ of 1, is simulated in SPICE for operating in both Mode I and Mode II. The simulation results of the phase variation $\alpha$ for both modes are shown in Figure 6-11. These results show that the variation of $\alpha$ with respect to $\theta_1$ is not linear because of slight distortions due to the phase difference between the zero crossing of $I_{L3}$ and $I_{L3_1}$. However, as this distortion is normally small and the two extreme points of $\alpha$ are fixed, $\alpha$ can be regarded as approximately linear versus $\theta_1$ and is modelled using a linear approximation for simplicity purposes. A full detailed discussion on the exact variation of $\alpha$ across the entire
operational range (θ₁ between 0 to π) is presented in Section 6.4. In the detailed analysis, it is proven that α is consistent for a given θ₁ regardless of \( V_{oc}, V_{DC} \) and \( Q_i \).

The analysis presented in this section uses this linear approximation to model the circulating current control characteristics. The linear approximation for α in both modes of operation is illustrated in Figure 6-12 and is expressed by:

\[
\alpha = \frac{\alpha_0 \mp \frac{\pi}{2}}{\pi} \theta_1 \pm \frac{\pi}{2}
\]

(6.10)

For Mode I, the sequence of the signs are minus, plus while for Mode II, this sequence is plus, minus.
6.3.3. Characteristic of circulating current duty cycle control

6.3.3.1. Characteristic of the output power

If it is assumed that there is zero power loss in the output rectifier and the associated shunt switches, the output real power of the circulating current controller can be simply described by the fundamental components of the output voltage $V_{LCL,1}$ (from (6.4)), the output current $I_{L3,1}$ (from (6.7)) and the power factor (from (6.10)). This results in:

$$P_{out} = V_{LCL,1} I_{L3,1} \cos \alpha$$

$$= \frac{2\sqrt{2} V_{DC} V_{oc}}{\pi X} \sin \left( \frac{\theta_1}{2} \right) \cos \alpha$$

$$= \frac{2\sqrt{2} V_{DC} V_{oc} Q}{\omega L_2} \sin \left( \frac{\theta_1}{2} \right) \cos \alpha$$

(6.11)

Most IPT pick-ups operate with a controlled (near constant) output DC voltage for conventional industrial applications, so that the output power is directly proportional to the average output DC current. Therefore, it is convenient to have an expression of the normalised output DC current in order to analyse its output power characteristic versus the variation of the diode conduction intervals and different $V_{DC}/V_{oc}$ ratio.

The ideal maximum DC output current ($I_{DC,\text{max}}$) of an LCL pick-up is also defined as [115, 116, 123]:

$$I_{DC,\text{max}} = \frac{2\sqrt{2} V_{oc}}{\pi X}$$

(6.12)

With (6.12) the conventional definition of DC power ($P=V_{DC}I_{DC}$), the normalised DC output current is expressed as:

$$I_{DC,n} = \sin \left( \frac{\theta_1}{2} \right) \cos \alpha$$

(6.13)

This expression indicates that the relationship between the output current and the diode conduction interval $\theta_1$ is not affected by $V_{DC}/V_{oc}$. This is expected and results from the output current source characteristic of the LCL pick-up. SPICE simulation results of the normalised output current for a range of $V_{DC}/V_{oc}$ ratio and an analytical result for a $V_{DC}/V_{oc}$ ratio of 1 using (6.10) and (6.13) for Mode I and Mode II are shown in Figure 6-13.
of the diode conduction interval used in the simulation results is “degrees” for the purpose of a finer grid scale in the graph. The simulation results for the normalised output current versus diode conduction interval remain constant regardless of the $V_{DC}/V_{oc}$ ratio. This concurs with the observation from (6.13). The analytical results modelled using the linear approximation of $\alpha$ for both modes are similar to the simulation results, but there are observable differences introduced by the linear approximation of the phase variation $\alpha$.

![Figure 6-13: Normalised pick-up output DC current ($I_{DC}/I_{DC\_max}$) as a function of diode conduction interval $\theta_1$ for a fixed $Q_1$ of 1 and various $V_{DC}/V_{oc}$ ratio and the analytical results using linear approximation of $\alpha$ with a $V_{DC}/V_{oc}$ and $Q_1$ of 1 (a) Mode I, (b) Mode II](image)

In both modes of operation, the normalised output current with a maximum diode conduction interval is around 0.95 instead of 1. This is because the output DC current has harmonics in it introduced by the rectifier. This causes the rectified DC current to be slightly distorted from an ideal rectified sine wave [116, 123]. This phenomenon could also be explained by looking at the phase difference between $V_{LCL\_1}$ and $I_{L3\_1}$. As discussed earlier, by default there is a phase difference $\alpha_0$ between $V_{LCL\_1}$ and $I_{L3\_1}$ with the circulating current controller operating with full diode conduction. This non-unity power factor at the AC output of the LCL pick-up prevents it from outputting its ideal maximum power with a diode conduction interval of 180°. However, as shown conceptually in Figure 6-12 (b), $\alpha$ is close to zero in Mode II while the diode conduction interval remains high, so that the product of the sine and cosine function in (6.13) will have a slight increase with $\theta_1$ being slightly less than 180°. This effect is shown in the simulation, and in the analytical results of Figure 6-13 (b). These results suggest that for Mode II, the normalised output current has a maximum and is close to one with $\theta_1$ around 160°. This means that the Mode II pick-up controller needs to
take account of this phenomena during operation. The simplest approach is to limit its
to 160° of the diode conduction interval to prevent a “lock up” condition
because of the change in the sign of the slope between 160° and 180°.

The modelling errors introduced in the use of linear approximation in the output
current calculation are within 10% of the maximum rated output current in both modes of
operation as shown in Figure 6-13. Consequently this simple modelling analysis is sufficient
for pick-up design purposes, where a controller can varying the output voltage to the level of
accuracy required.

6.3.3.2. Characteristic of the reflected impedance

It is important to gain insight into both the real and reactive power loading conditions
of the pick-up on the primary power supply from an overall system design perspective. Thus,
both the real and reactive power sourced from the primary track needs to be determined. If
only the fundamental components in the analysis are considered, the real and reactive loads
reflected back onto the primary track from the pick-up are given by:

\[ P = \text{Re}(Z_r) I_1^2 \]  
\[ \text{VAR} = \text{Im}(Z_r) I_1^2 \]

Here \( Z_r \) is the pick-up reflected impedance and is given by [106, 109]:

\[ Z_r = \frac{\omega^2 M^2}{Z_2} \]

where \( Z_2 \) is the impedance seen by the pick-up open circuit voltage source \( V_{oc} \) of the
series-parallel tuned LCL pick-up as shown in Figure 6-8.

When tuned at the operating frequency \( \omega \), this LCL pick-up can be modelled with a
perfectly tuned network using the equivalent circuit shown in Figure 6-8. The input
impedance of this equivalent LCL network is given by [155]:

\[ Z_2 = \frac{X^2}{Z_{load}} = \frac{X^2}{R_{load} + jX_{load}} \]
where $R_{\text{load}}$ and $X_{\text{load}}$ are the output loading impedance of the circulating current controller after the model of a perfectly tuned LCL network (as indicated in Figure 6-8).

Using (6.7) and (6.11) the output loading resistance is then given by:

$$R_{\text{load}} = \frac{2\sqrt{2}}{\pi} \frac{V_{\text{dc}}}{V_{oc}} \frac{\omega L_2}{Q_i} \sin \left( \frac{\theta_1}{2} \right) \cos \alpha$$ (6.18)

The output reactive power can be derived in the similar fashion as the output real power expression shown in Section 6.3.3.1. If only the fundamental components of the output voltage and current are considered, the output reactive power of the circulating current controller is given by:

$$VAR_{\text{out}} = V_{\text{LCL}} I_{L3} \sin \alpha - 0.4674 X_{L3} I_{L3}^2$$

$$= 2\sqrt{2} \frac{V_{\text{dc}} V_{oc}}{X} \sin \left( \frac{\theta_1}{2} \right) \sin \alpha - 0.4674 \frac{V_{\text{dc}} V_{oc}}{\sqrt{2} X}$$ (6.19)

and the output reactance is then given by:

$$X_{\text{load}} = 2\sqrt{2} \frac{V_{\text{dc}}}{\pi} \frac{\omega L_2}{Q_i} \sin \left( \frac{\theta_1}{2} \right) \sin \alpha - 0.4674 \frac{V_{\text{dc}}}{\sqrt{2} \omega L_2}$$ (6.20)

The output reactive load of the circulating current controller is composed of a variable term, which is similar to the output power expression, and a constant value which represents the mismatched impedance of $X_{L3}$ and $X_{C3}$ as indicated in Figure 6-8. Using (6.16), (6.17), (6.18) and (6.20) the normalised reflected resistive and reactive impedances of the circulating current controlled pick-up can be expressed by (6.21) and (6.22) respectively.

$$\text{Re}(Z_{\text{ref}}) = \frac{\text{Re}(Z_r)}{Z_0} = \frac{2\sqrt{2}}{\pi} \frac{V_{\text{dc}}}{V_{oc}} Q_i \sin \left( \frac{\theta_1}{2} \right) \cos \alpha$$ (6.21)

$$\text{Im}(Z_{\text{ref}}) = \frac{\text{Im}(Z_r)}{Z_0} = \frac{2\sqrt{2}}{\pi} \frac{V_{\text{dc}}}{V_{oc}} Q_i \sin \left( \frac{\theta_1}{2} \right) \sin \alpha - 0.4674 \frac{V_{\text{dc}}}{\sqrt{2} V_{oc}} Q_i$$ (6.22)
As shown, both the resistive and reactive components are normalised using:

\[ Z_0 = \frac{\omega M^2}{L_x} \]  

(6.23)

As expected, these expressions for both the normalised reflected resistive and reactive impedance of the circulating current controller are very similar to a conventional UPF LCL pick-up with load decoupling switch. The reflected impedance of the UPF LCL pick-up while delivering power was discussed in Chapter 2 [36, 116, 123]. Here it is derived for a working UPF LCL pick-up with DC output regulator switching with duty cycle D:

\[
Z_{LCLr,n} = \frac{Z_{LCLr}}{Z_0} = \frac{2\sqrt{2} V_{DC}}{\pi V_{ac}} Q_i (1 - D) - j \frac{0.4674 V_{DC}}{\sqrt{2} V_{ac}} Q_i D
\]

(6.24)

Here D is the duty cycle of the load decoupling switch and is either 0 or 1.

The expressions for the reflected impedance of both controllers show they are proportional to the product of Q_i and the ratio of V_{DC}/V_{ac}. The difference (as shown), is that the magnitude of the reflected load of the circulating current controller can now be controlled continuously from zero to full load, which is not possible in the slow switching hysteresis controller of the traditional LCL pick-up.

The reflected reactance of (6.22) is composed of two terms: one is a variable of \( \theta_1 \) and \( \alpha \), and the other one is a constant value. In Mode I, as \( \theta_1 \) varies from 180° to 0, \( \alpha \) varies from \( \alpha_0 \) (21.537°) to 90°. As such, while \( \theta_1 \) remains relatively high the reflected reactance will be inductive with increasing \( \alpha \) because the phase of \( V_{LCL,1} \) leads \( I_{L3,1} \) (as illustrated in the phasor diagram of Figure 6-10 (a)). As \( \theta_1 \) is reduced the magnitude of \( V_{LCL,1} \) reduces. When \( \theta_1 \) is below 90°, the magnitude of the reflected reactance will start to decrease despite the continually increasing \( \alpha \) and it will become capacitive when \( \theta_1 \) is below ~40° which will be shown in the simulation results in the next section. In Mode II, \( \alpha \) varies from \( \alpha_0 \) to -90°, so the reflected reactance will be capacitive for the entire operational range of \( \theta_1 \).
6.3.3.3. Simulation results of the reflected impedance

The SPICE simulation results of the normalised reflected resistive load for a fixed $Q_I$ of one and a range of $V_{DC}/V_{oc}$ ratio are shown in Figure 6-14 (a) and Figure 6-15 (a) for Mode I and Mode II respectively. The reflected resistance with respect to the diode conduction interval at the various operating $V_{DC}/V_{oc}$ ratios matches the normalised DC output current shown in Figure 6-13, as both describe the real output power of the controller. The simulation results of the controller with a $V_{DC}/V_{oc}$ of 1 are compared with the analytical results using (6.21) with the same operating conditions in Figure 6-14 (b) and Figure 6-15 (b) for Mode I and Mode II respectively. Similar to the earlier comparisons of the normalised DC output current, the analytical results shown here are a good approximation to the simulation results but with errors which are introduced by the modelling error in the phase variation $\alpha$.

![Figure 6-14: Mode I - Normalised reflected resistance $Re(Z_{r,n})$ versus diode conduction interval $\theta_1$ with a $Q_I$ of 1 and a range of $V_{DC}/V_{oc}$ (a): SPICE simulation results. (b) SPICE simulation results and analytical results using linear approximation of $\alpha$ with $V_{DC}/V_{oc}=1$](image-url)
Chapter 6  A Circulating Current Controller for Series-Parallel LCL pick-up

Figure 6-15: Mode II - Normalised reflected resistance $\text{Re}(Z_r, n)$ versus diode conduction interval $\theta_1$ with a $Q_1$ of 1 and a range of $V_{\text{DC}}/V_{\text{oc}}$ (a): SPICE simulation results. (b) SPICE simulation results and analytical results using linear approximation of $\alpha$ with $V_{\text{DC}}/V_{\text{oc}}=1$

The SPICE simulation results for the normalised reflected reactive load for a fixed $Q_1$ of one and a range of $V_{\text{DC}}/V_{\text{oc}}$ for Mode I and Mode II are shown in Figure 6-16 (a) and Figure 6-17 (a) respectively. As predicted by (6.22), the reflected reactance variation profile of Mode I swings between an inductive and a capacitive load with the diode conduction interval varying over its operational range. When $\theta_1$ is maintained between $40^\circ$ to $180^\circ$, the reflected load seen by the primary track is inductive and has a maximum value at around $90^\circ$. This inductive load slightly increases the track inductance. When $\theta_1$ varies between 0 to $40^\circ$, the reflected reactance to the primary is capacitive which slightly reduce the track inductance. For Mode II, the reflected reactance is purely capacitive and has a maximum value at a diode conduction interval of around $90^\circ$.

The reflected reactance simulation results with a $V_{\text{DC}}/V_{\text{oc}}$ of 1 are compared with the analytical results using (6.22) with the same operating conditions in Figure 6-16 (b) and Figure 6-17 (b) for Mode I and Mode II respectively. Similar to the reflected resistance, the analytical results for both Mode I and Mode II show similar trends to the simulated results but again with some differences. Despite the errors in the linear approximated model of $\alpha$, for most cases the maximum and the minimum value of the analytical results of both the reflected resistive and reactive impedance are sufficiently accurate for design purposes. However, if an operational analysis of the circulating current controller is required a more precise expression of the phase variation $\alpha$ is needed. This could be useful in an overall system optimisation for
an application with a distributed track and multiple pick-ups where different pick-ups could operate in different modes to minimise the total reflected impedance on the primary track. A detailed analysis of the variation of $\alpha$ is presented in Section 6.4 which enables a more accurate approximation of the reflected impedances.

Figure 6-16: Mode I - Normalised reflected reactance Im($Z_{r,n}$) versus diode conduction interval $\theta_1$ with a $Q_1$ of 1 and a range of $V_{DC}/V_{oc}$ (a): SPICE simulation results. (b) SPICE simulation results and analytical results using a linear approximation for $\alpha$ with $V_{DC}/V_{oc}=1$

Figure 6-17: Mode II - Normalised reflected reactance Im($Z_{r,n}$) versus diode conduction interval $\theta_1$ with a $Q_1$ of 1 and a range of $V_{DC}/V_{oc}$ (a): SPICE simulation results. (b) SPICE simulation results and analytical results using a linear approximation for $\alpha$ with $V_{DC}/V_{oc}=1$
As mentioned earlier, the variation in the reflected reactance of the controller on the primary track must be taken into account at the system level, given most power supplies operating at fixed frequency have limitations on the track inductance variation which can be tolerated. However, with the reflected impedance being a mixture of resistive and reactive components across its designed power range, it is also important that the magnitude of the total reflected impedance does not exceed the primary power supply volt-amp rating within its operation range.

Simulation results of the magnitude of the total reflected impedance for various \( V_{DC}/V_{ac} \) ratios’ are shown in Figure 6-18. From this it is clear that for Mode I, the maximum reflected impedance occurs when the controller is operating with a full diode conduction interval, corresponding to maximum output power. For Mode II, the maximum reflected impedance occurs at a diode conduction interval of 135°. The magnitude of the total reflected impedance at this diode conduction interval is 10% higher than the maximum reflected resistance from the controller when operating at maximum power, which is operating at a diode conduction interval of around 160° (as outlined in Section 6.3.3.1). This shows that the VA rating of the primary power supply needs to be 20% above its designed power rating when the pick-up controller is operating in Mode II.

![Figure 6-18: Magnitude of the reflected impedance on the primary track (a) Mode I, (b) Mode II](image)
6.4. A Detailed Analysis of the Phase Relation $\alpha$ Between $V_{LCL,1}$ and $I_{L3,1}$

In the previous section, an assumption was made that the phase variation $\alpha$ is not affected by the harmonic components in $I_{L3}$ but is only controlled by variations in the diode conduction interval. This results in some errors in the linear approximation method. In this section, an analytical expression for the current $I_{L3}$ is presented. However, it is also shown that it is difficult to derive an analytical expression for the phase variation $\alpha$. Consequently an alternative polynomial approximation for $\alpha$, which includes the impact of the harmonic content in $I_{L3}$, is presented. This polynomial approximation enables the characteristics of the circulating current controller to be more accurately modelled for system analysis.

6.4.1. An expression for $I_{L3}$ including the harmonic components

In Section 6.3.1 it was demonstrated, using the principle of superposition and considering only the fundamental components that the current $I_{L3,1}$ only consists of $I_{L3a}$, which corresponds to the input voltage $V_{oc}$. However, to obtain an exact expression for $I_{L3}$, the harmonic components in $I_{L3b}$ generated from $V_{LCL}$ need to be included. A series of conceptual waveforms of the circulating current controller operating in Mode I are shown in Figure 6-19. Here the phase difference between $V_{LCL,1}$ and $V_{oc}$ is defined as $\varphi$. $\theta_a$ is defined to be the zero crossing of $I_{L3}$ and is also the voltage transition of $V_{LCL}$ from 0 to $V_{DC}$. 
Using $V_{oc}$ (from (6.25)), as the phase reference, the symmetrical variable width single-phase square wave voltage $V_{LCL}$ can be described using a Fourier series [115, 160].

$$V_{oc}(\theta) = \hat{V}_{oc} \cos(\theta)$$  \hspace{1cm} (6.25)

$$V_{LCL}(\theta) = \sum_{h=1}^{\infty} \left( V_{LCL}^h \right) \cos \left( h(\theta - \varphi) \right)$$  \hspace{1cm} (6.26)

Here $\left( V_{LCL}^h \right)_h$ is the magnitude of the $h^{th}$ harmonic of the square wave and is given by

$$\left( V_{LCL}^h \right)_h = \begin{cases} 0 & h = \text{even} \\ \frac{4}{\pi h} V_{DC} \sin \left( h \frac{\theta_1}{2} \right) & h = \text{odd} \end{cases}$$  \hspace{1cm} (6.27)

Using the principle of superposition, the equivalent circuit of Figure 6-7 with respect to the $h^{th}$ harmonic of $V_{LCL}$ is shown in Figure 6-20.
Figure 6-20: An equivalent circuit of an LCL pick-up using super position principle with respect to $V_{LCL}$.

The impedance of the LCL network $Z_{out}$ at the $h^{th}$ harmonic seen by $V_{LCL}$ is then given by:

$$Z_{out\ h} = jX_{L_2\ h} - jX_{C_1\ h} + \frac{1}{jX_{L_2\ h} - jX_{C_2\ h} - jX_{C_3\ h}} + \frac{1}{jX_{L_2\ h} - jX_{C_2\ h} - jX_{C_3\ h}}$$

(6.28)

where $Z_{out\ h}$ is the impedance corresponding to the $h^{th}$ harmonic component.

Using (6.5) and (6.6) with the definition of $Q_l$ from (6.3), the impedance of each component in the equivalent circuit is given below. The impedances are represented by the designed LCL characteristic impedance $X$ at the operating frequency and the harmonic order $h$.

$$X_{L_2\ h} = hXQ_l$$

(6.29)

$$X_{C_2\ h} = (Q_l - 1)\frac{X}{h}$$

(6.30)

$$X_{C_3\ h} = X/h$$

(6.31)

$$X_{L_3\ h} = \frac{V_{dc}}{\sqrt{2V_{oc}}} hX$$

(6.32)

$$X_{C_1\ h} = \frac{0.4674 V_{dc}}{\sqrt{2V_{oc}}} hX = 0.3305 \frac{V_{dc}}{V_{oc}} \frac{X}{h}$$

(6.33)

Therefore, $I_{L3b}$ is then expressed by:

$$I_{L3b}(\theta) = \sum_{h=1}^{\infty} \left( \frac{V_{LCL}}{h Z_{out\ h}} \right) \cos(h(\theta - \phi))$$

(6.34)
With the definition of $I_{L3a}$ and $I_{L3,1}$ given in (6.7), the exact expression for $I_{L3}$ can now be written as:

$$I_{L3}(\theta) = -j \frac{V_{oc} \cos(\theta)}{X} \sum_{h=1}^{\infty} \left( \frac{h}{Z_{out,h}} \right) \cos \left( h(\theta - \phi) \right)$$

(6.35)

However, the exact total output impedance $Z_{out,h}$ at the corresponding harmonic frequency is different with the LCL network designed for different values of $V_{DC}/V_{oc}$ ratio and $Q_t$. Therefore, given $\theta_1$ analytically solving the above expression for the phase difference $\phi$ and $\alpha$ is very difficult. Alternatively, a polynomial approximation of the phase variation $\alpha$ for an UPF LCL pick-up calculated using the SPICE simulator may be used to describe the phase variation $\alpha$ and is presented in the next section.

### 6.4.2. Polynomial approximations of the phase variation $\alpha$

For approximating the phase variation $\alpha$ using polynomial equations, ideally a separate approximation equation is required for every particular combination of the designed values of $V_{DC}/V_{oc}$ and $Q_t$, given the total output impedance $Z_{out,h}$ is different. This is however practically impossible given the very large number of possible $Q_t$ and $V_{DC}/V_{oc}$ ratio combinations. However in practice a 3rd order polynomial approximation of the phase variation $\alpha$ is accurate enough for modelling the characteristic of the circulating current controller with UPF LCL pick-up for all practical combinations of $V_{DC}/V_{oc}$ and $Q_t$. For example, the UPF LCL pick-up design topology is not suitable for $V_{DC}/V_{oc}$ less than 1 [116].

By looking at the expression of $Z_{out,h}$ given in (6.28) and the equivalent circuit in Figure 6-20, the series-parallel LCL network is composed of a series resonant network ($L_3$ and $C_3$) and a 2nd or 3rd order parallel resonant network ($L_2$, $C_{L2}$ and $C_2$), which is identical to a band pass filter. The impedance ($Z'_{out}$) of a 2nd or 3rd order LC band pass filter with high quality factor $Q$ (which results in low bandwidth given $BW \propto \omega/Q$) is ideally $\infty$ at the tuned operating frequency. Once outside the tuned frequency range, the impedance reduces significantly. Therefore, when $h>1$ the total impedance $Z_{out,h}$ is dominated by $X_{L3,h}$ and $X_{C3,h}$ as given in (6.32) and (6.33) respectively. Under this circumstance, $Z_{out,h}$ can be approximated as:

$$Z_{out,h} \approx j \frac{V_{DC} h X}{\sqrt{2V_{oc}}} - j0.3305 \frac{V_{DC} X}{h V_{oc}}$$

(6.36)
where $h > 1$.

As indicated in (6.36), $Z_{\text{out},h}$ is now directly proportional to the ratio of $V_{\text{DC}}/V_{\text{oc}}$ and $X$. The normalised MATLAB simulation results of $Z_{\text{out},h}$ with respect to $X$ for the $3^{\text{rd}}$, $5^{\text{th}}$, $7^{\text{th}}$ and the $9^{\text{th}}$ order of harmonic components, shown in Figure 6-21, agree with the observation from (6.36). The normalised $Z_{\text{out},h}$ is directly proportional to the ratio of $V_{\text{DC}}/V_{\text{oc}}$ regardless of $Q_1$ given $Q_1$ is included in $X$.

As the harmonic components of $V_{\text{LCL}}$ are directly proportional to $V_{\text{DC}}$, $I_{\text{L3b}}$ and hence the harmonic components of $I_{\text{L3}}$ are proportional to $V_{\text{oc}}/X$. Given the definition of $I_{\text{L3,1}}$ in (6.7) is also proportional to $V_{\text{oc}}/X$, the magnitude of the $I_{\text{L3}}$ harmonic components relative to its fundamental component $I_{\text{L3,1}}$ is near constant and independent of the ratio between $V_{\text{DC}}/V_{\text{oc}}$ and $Q_1$. Therefore, for a given $\theta_1$ the phase of the zero crossing of $I_{\text{L3}}(\theta_a)$ is also approximately consistent regardless of the value of $V_{\text{DC}}/V_{\text{oc}}$ ratio or $Q_1$ of the LCL pick-up. In consequence, for a given $\theta_1$ the phase variation $\alpha$ remains independent of $V_{\text{DC}}/V_{\text{oc}}$ and $Q_1$ for all practical designs.
This observation concurs with the SPICE simulation results of the normalised reflected impedance shown in Figure 6-18 where the simulation results profile is directly proportional to the ratio of $V_{DC}/V_{oc}$. The value of $\alpha$ calculated using the SPICE simulator for Mode I and Mode II were given in Figure 6-11 in Section 6.3.2 and is shown here again in Figure 6-22. Two $3^{rd}$ order polynomial equations can be used to model these results as given in (6.37) and (6.38) for Mode I and Mode II respectively and they are plotted in Figure 6-22 to compare against simulation results. These two equations are also compared against practical measurements in Section 6.5.

Two 3rd order polynomial equations can be used to model these results as given in (6.37) and (6.38) for Mode I and Mode II respectively and they are plotted in Figure 6-22 to compare against simulation results. These two equations are also compared against practical measurements in Section 6.5.

$$Mode \ I:\ 
\alpha = 3.505 \times 10^{-5} \times \theta_1^3 - 9.34 \times 10^{-3} \times \theta_1^2 + 1.676 \times 10^{-1} \times \theta_1 + 90 \tag{6.37}$$

$$Mode \ II:\ 
\alpha = 3.535 \times 10^{-5} \times \theta_1^3 - 9.465 \times 10^{-3} \times \theta_1^2 + 1.18 \times \theta_1 - 90 \tag{6.38}$$

6.5. Design Example, Experimental Results and Discussions

6.5.1. Specification and implementation of design example

A practical prototype was built for an EV battery charging application using the proposed circulating current control topology with an air gap of 270mm. The rated power output of the prototype is 1.5kW at 300V DC. This prototype is tested with a 3kW fixed
frequency IPT power supply which also uses an LCL resonant network [120]. The magnetic structure used in this application is a pair of 760 x 400mm Double D pads which were developed and presented in Chapter 3. The parameters of the primary power supply and the Double D charging pads aligned horizontally with an air gap of 270mm are listed in Table 6-1. With these given parameters, the UPF LCL pick-up used with the circulating current controller is designed according to [116, 123] and its parameters are given in Table 6-1.

<table>
<thead>
<tr>
<th>3kW Power Supply</th>
<th>Magnetic Details of 760x400mm Double D pads</th>
<th>Pick-up Controller</th>
</tr>
</thead>
<tbody>
<tr>
<td>V_d 300V DC</td>
<td>L_1 148.79μH</td>
<td>C_2 0.422μF</td>
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<tr>
<td>Frequency 20kHz</td>
<td>L_2 149.53μH</td>
<td>C_3 0.219μF</td>
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<td>Rs,L_1 56mΩ</td>
<td>X 18.86Ω</td>
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<td></td>
<td>Rs,L_2 57mΩ</td>
<td>V_{oc} 103.4V RMS</td>
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<td></td>
<td>Q_{L_1} (L_1 native Q) 330</td>
<td>L_3 313.56μH</td>
</tr>
<tr>
<td>Position (x,y,z) 0, 0, 270mm</td>
<td></td>
<td>I_{sc} 5.5A RMS</td>
</tr>
<tr>
<td>Q_{L_2} (L_2 native Q) 326</td>
<td></td>
<td>Q_{L_3} (L_3 native Q) 555</td>
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<tr>
<td></td>
<td>RS,L_3 71mΩ</td>
<td>V_{DC} 300V</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Po_{ud} 1.5kW</td>
</tr>
</tbody>
</table>

Table 6-1: Parameters of 3kW IPT power supply, a pair of 760 x 400mm Double D charging pad with an air gap of 270mm and the designed UPF LCL pick-up (Parameters are taken with Agilent LCR meter Model E4980A)

The implemented controller for the LCL pick-up circulating current control is shown conceptually in Figure 6-23 and a photograph of the practical circuit is shown in Figure 6-24. Diodes D_1 to D_4 used in the implemented circuit were Fairchild FFH60UP40S. The switches S_1 and S_2 are Fairchild IGBT FGH60N60UF. The microcontroller used here is a Cypress PSoC CY8C29466-24PXI. A current transformer (CT) and associated circuitry, which is not shown, is used to sense the zero crossing of I_{L3} for synchronising the gate drive signal V_{g1} and V_{g2}. As explained in Section 6.4, the zero crossing of I_{L3} varies during operation with changes of diode conduction interval due to the magnitude variation of harmonic components in I_{L3}. This variation in zero crossing of I_{L3} during operation affects the microcontroller’s
ability to synchronise its gate driving waveform. Therefore, in the implemented prototype a phase lock loop (PLL) is used as a buffer between the microcontroller and the CT feedback to smooth out the change in the zero crossing of $I_{L3}$ with changes in the diode conduction interval.

Figure 6-23: Conceptual diagram of the implemented circulating current controller

Figure 6-24: A photo of the circulating current controller prototype
6.5.2. Experimental results and Discussion

The experimental setup for measuring the performance and the reflected impedance of the circulating current controller prototype is shown conceptually in Figure 6-25. The primary power supply is powered by an Agilent power supply 6813B. This setup is similar to the experiments presented in Chapter 5. The difference is that the Clark-Hess Model 2335 power meter is placed on the primary side of the charging pad to measure the reflected real and reactive power from the circulating current controller on the primary side as demonstrated in Figure 6-25. The inductor $L_t$ is the designed output inductor of the resonant LCL network in the power supply and capacitor $C_t$ is used to fully series compensate the inductance of the charging pad. With this experimental setup the voltage measured by the power meter is ideally only the reflected voltage ($V_r$) from the pick-up. Thus ideally the phase difference between $V_r$ and $I_1$ measured by the power meter only represents the reflected load from the pick-up to ensure a good accuracy is achieved in the measurements.

![Figure 6-25: Conceptual diagram of experimental setup for power and VAR measurements](image)

Oscilloscope captures of the prototype operating at one third, two thirds and full rated power are shown in Figure 6-26 for both Mode I and Mode II. For consistency of comparison between both modes, here the full rated power in Mode II refers to the case where the circuit is operating with maximum diode conduction interval instead of operating at the maximum output power with $\theta_1 = 160^\circ$ as mentioned in Section 6.3.3.1. The top trace in each capture is the current $I_{L3}$, the second trace is the LCL AC output voltage $V_{LCL}$, while the third trace is the total output current of $D_1$ and $D_2$. As expected, when a full diode conduction interval is applied to both modes, the waveforms are identical to each other. These captures demonstrate that the AC output voltage $V_{LCL}$ and the output DC current ($I_{D1} + I_{D2}$) are successfully controlled as the system regulates the output power to the load $R_{DC}$.

As demonstrated in Section 6.2.1, in Mode I both diodes $D_1$ and $D_2$ are softly turned on but hard switched at turn off once $\theta_1$ is less than $180^\circ$. This hard turn off of the diodes results in diode reverse recovery currents in the circuit. These are observable in the diode
output current traces in Figure 6-26 (a) and (b). In comparison, the Mode II diode output current traces in Figure 6-26 (d) and (e) do not have any observable diode reverse recovery currents as the diodes D1 and D2 are turned off softly as explained in Section 6.2.2. As a result, the waveforms of Mode II have less noise present in the system compared with Mode I under operation as shown in Figure 6-26.

The output power, the reflected reactance and the efficiency measurements of the pick-up for both Mode I and Mode II are shown in Figure 6-27, Figure 6-28 and Figure 6-29 respectively. The output power and reflected reactance profile are compared against simulation and analytical results using the polynomial approximation explained in Section 6.4.2. The practical measurements show good agreement with the simulation and analytical results.

In this prototype system, the maximum reflected inductive impedance from the pick-up operating in Mode I back on to the track is 0.2Ω which is 1.6µH at 20kHz. This occurs with a diode conduction interval of 90° as indicated in Figure 6-28 (a). This reflected inductive load of 1.6µH is 3.8% of the rated power supply inductance. A practical IPT power supply is normally designed with a track inductance tolerance of +/- 5-10%, so that the reflected inductive load from this circulating current control prototype would not be an issue in this design. However, an inductive load at the output of a power supply adopting an LCL resonant network topology results in a capacitive load on the supply inverter bridge [117, 119, 120, 156]. Therefore, good practice suggests that this reflected inductive load should be considered while designing the power supply to ensure that no resultant capacitive loading on the inverter bridge occurs.

On the other hand, the maximum reflected capacitive load from the controller operating in Mode II occurs at θ1 = 90° as indicated in Figure 6-28 (b). This capacitive impedance is -0.7Ω which is 11.37µF at 20kHz. This reflected capacitive load corresponds to reducing the track inductance by 5.57µH which is 13% of the designed track inductance. Although this variation exceeds the general 10% tolerance, a power supply with LCL resonant network has better tolerance to track inductance variations which are less than its designed value [117, 119]. Although the extra reactive load introduced by the circulating current controller slightly detunes the primary power supply, in practice this reactive load can easily be handled by normal system parameter tolerances.
Figure 6-26: Oscilloscope captures of circulating current controller operating in Mode I (a) – (c) and in Mode II (d) – (f). Traces from top to bottom are $I_{L3}$, $V_{LCL}$ and total diode output current ($I_{D1} + I_{D2}$) for (a) (d) 1/3 of rated power, (b) (e) 2/3 of rated power and (c) (f) full power.
Chapter 6  A Circulating Current Controller for Series-Parallel LCL pick-up

Figure 6-27: Output DC current $I_{DC}$ of simulation, analytical and practical results versus diode conduction interval for (a) Mode I (b) Mode II

Figure 6-28: Reflected reactance $\text{Im}(Z_r)$ of simulation, analytical and practical results versus diode conduction interval for (a) Mode I, (b) Mode II

Figure 6-29: Efficiency measurements versus output power for (a) Mode I, (b) Mode II
Power efficiency measurements of the prototype system were taken at the DC input to the primary power supply $V_d$, the AC output of the primary power supply and the DC output of the circulating current controller $V_{DC}$, as indicated in Figure 6-25. The efficiency of the overall system, the pick-up controller and the charging pads is shown in Figure 6-29. The pick-up efficiency which includes the pick-up pad and the controller circuitry is calculated using the measurements at the AC output of the power supply minus the calculated conduction loss in the primary pad with parameters given in Table 6-1. The charging pad conduction loss is calculated using the native quality factor ($Q_L$) of the charging pad inductor with its self inductance, its winding and core loss. This is given by:

$$P_{loss} = \frac{\omega L L_1^2}{Q_L}$$  \hspace{1cm} (6.39)

The overall system efficiency for both modes is 80% at the designed maximum power and 70% when operating at half of the designed output power. In Figure 6-29 the gradient of the overall system efficiency measurements for both modes is still positive at the designed 1.5kW. This is because both the power supply and the charging pad used in this prototype system are not operating close to their original rated power level which is for 3kW applications. Therefore, the overall system efficiency should increase and is expected to be between 85 to 90% when operating at the designed 3kW rated power.

The pick-up efficiency in both modes is similar. Both have an efficiency of 95% at the maximum power level and remain above 85% efficient over most of the power range. A breakdown of the pick-up losses while operating in Mode II at a diode conduction interval of 100° is provided in Table 6-2. The loss in each component is calculated based on the measured waveforms and its datasheet. The majority of the identified losses are the pick-up pad, the IGBT’s and the diode rectifier. The remaining 35.4W losses are unaccounted for but could be a combination of resistive mechanical joints in the circuitry, losses in the output DC filter capacitor and conduction and eddy current losses in the printed circuit board.
# A Circulating Current Controller for Series-Parallel LCL pick-up

## Chapter 6

### Item Loss %

<table>
<thead>
<tr>
<th>Item</th>
<th>Loss</th>
<th>%</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pick-up pad</td>
<td>14.1 W</td>
<td>1.52%</td>
</tr>
<tr>
<td>Inductor L3</td>
<td>2.9 W</td>
<td>0.31%</td>
</tr>
<tr>
<td>Capacitor C2</td>
<td>0.3 W</td>
<td>0.03%</td>
</tr>
<tr>
<td>Capacitor C3</td>
<td>0.6 W</td>
<td>0.07%</td>
</tr>
<tr>
<td>IGBT switching and conduction loss</td>
<td>13.4 W</td>
<td>1.44%</td>
</tr>
<tr>
<td>Diode conduction loss</td>
<td>16.6 W</td>
<td>1.79%</td>
</tr>
<tr>
<td>Mechanical joint, DC capacitor and unknown</td>
<td>35.4 W</td>
<td>3.82%</td>
</tr>
<tr>
<td><strong>Total Loss</strong></td>
<td>83.2 W</td>
<td>8.97%</td>
</tr>
</tbody>
</table>

*Table 6-2: Breakdown of calculated losses in the pick-up operating in Mode II at $\theta_1 = 100$ degrees*

### 6.6. Conclusion

A new circulating current controller for a series-parallel tuned LCL pick-up is proposed in this chapter. The proposed topology regulates the power flow by adjusting the equivalent AC output voltage of the LCL network using two diodes and two switches switching synchronously to the IPT frequency. This enables continuous power control from no load to full load for the LCL pick-up structure.

The proposed controller has two control modes which are referred to here as Mode I and Mode II. A detailed steady state AC analysis and the characteristics of both modes has been presented. Due to the nature of the controller operation, the circulating current controller characteristically also reflects a small amount of reactive impedance back onto the primary track, that slightly detunes the primary power supply during operation. In Mode I, the controller reflects either an inductive or capacitive impedance depending on the diode conduction interval. In Mode II, the controller reflects only a capacitive load back onto the track.

A 1.5 kW 300V DC prototype system for an EV battery charging was successfully implemented as a design example. The measured profile of the output power and the reflected impedance supports the presented AC analysis of the controller. In this practical design example, the reflected reactance variation from the controller can easily be accommodated within normal acceptable tolerances of the power supply. The efficiency of the pick-up charging pad and its controller for both modes is 95% at maximum output power and is still above 85% when operating at one third of the rated power.
Chapter 7.

A Design Example for a Circulating Current Controller in Variable coupling system

7.1. Introduction

The novel fast switching circulating current controller for series-parallel LCL pick-ups, presented in Chapter 6, is redrawn in Figure 7-1 (a) along with its conceptual operating waveforms in Figure 7-1 (b) [158, 159]. This controller was shown to have good efficiency and the ability to vary and regulate its output power smoothly from zero to full power. As explained in Chapter 5, this feature is very desirable for IPT EV battery charging applications where the magnetic coupling between the charging pads, and in consequence the coupled open circuit voltage ($V_{oc}$), can vary while the pick-up controller is still required to provide constant and regulated output power to the battery.

In this chapter, the steady state analysis approach presented in Chapter 6 is used again to examine the power flow of the circulating current controller and its reflected impedance back on to the primary track while operating with variable mutual coupling. In Chapter 6, a design example for a 1.5kW 300V DC EV battery charging system prototype was presented. This pick-up controller prototype was designed for a system with fixed coupling and an air gap of 270mm. At the end of this chapter, the same 1.5kW pick-up controller prototype, with added modifications to enable operation with a variable $V_{oc}$, is used to investigate its performance when operating with air gaps that could vary between 220mm and 270mm using the same 760x400mm Double D charging pads.
7.2. Steady State Analysis for operating in Variable Coupling Application

As discussed in [116, 122, 123] and in Chapter 5, a key characteristic of the series-parallel LCL pick-up with rectifier DC output is that the current $I_{L3}$ is slightly distorted during the rectification process. This necessarily introduces additional inductive reactance, (referred to as $X_{rec}$ in Chapter 5), in series with $L_3$ and $C_3$. The value of $X_{rec}$ varies with different $V_{DC}/V_{oc}$ ratios [116, 123].

The design procedure presented in Chapter 5 is again used to design the combination of $L_3$ and $C_3$ of this LCL network to enable operation in this variable coupling application. The purpose of this approach is to minimise the variation in the total combined reactance of $X_{L3}$, $X_{C3}$ and $X_{rec}$ and to keep this total reactance near the value of the LCL characteristic impedance $X$ within the given variation of $V_{DC}/V_{oc}$ ratio. Consequently, when both switches $S_1$ and $S_2$ are off to enable full power delivery, near unity power factor (UPF) is then achieved at the input of the LCL network despite the varying input voltage ($V_{oc}$). For simplicity the analysis presented in this section assumes that the input PF of the LCL pick-up with both switches $S_1$ and $S_2$ off remains unity.
7.2.1. Fundamental analysis

It was shown in Chapter 6 that the real and reactive power flow in the circulating current controller can be modelled adequately by considering only the fundamental components of the output voltage $V_{LCL}$ ($V_{LCL_1}$) and the circulating current $I_{L3}$ ($I_{L3_1}$), and the phase difference $\alpha$ between them. Therefore following the analysis of Chapter 6, the output rectifier, the controlling switches and the output DC capacitors (of Figure 7-1 (a)) can all be modelled with a controllable output voltage source. By considering only the fundamental components, the equivalent circuit of the circulating current controller is shown in Figure 7-2. Each branch of the tuned LCL pick-up is represented by its characteristic impedance $X$. An additional mismatch reactance component ($-jX_{mm}$) in series with the output voltage source $V_{LCL_1}$ is used to represent the total impedance mismatch of $X_L3$ and $X_C3$ as indicated in Figure 7-2. This mismatched reactance ($X_{mm}$) is given by:

$$jX_{mm} = jX - j(X_{L3} - X_{C3})$$  \(7.1\)

Unlike the UPF LCL pick-up presented in [116, 123], where the mismatched reactance ($0.4674X_{L3}$) is a fixed portion of the circuit parameter $X_{L3}$ for the given $V_{DC}/V_{oc}$ ratio, $X_{mm}$ of the LCL pick-up designed to operate with variable input voltages varies with different $V_{DC}/V_{oc}$ specifications as discussed in Chapter 5. In consequence, the phase variation $\alpha$ between $V_{LCL_1}$ and $I_{L3_1}$ for a given diode conduction interval $\theta_1$ also varies with different combinations of $L_3$ and $C_3$. As such, the 3rd order polynomial approximation for $\alpha$ presented in Chapter 6 cannot be applied here. Therefore, in this chapter $\alpha$ is modelled using the alternative linear approximation (presented in Chapter 6) with modifications to estimate the power flow in the circuit and consequently its reflected real and reactive load back on to the primary track. This linear approximation method was proven to be effective for estimating the reflected loading profile of the circulating current controller for design purposes and will be explained in Section 7.2.2.
Chapter 7  A Design Example for a Circulating Current Controller in Variable coupling system

Figure 7-2: A fundamental components equivalent circuit diagram of a LCL pick-up designed for use in system with variable mutual coupling

Apart from the difference between the mismatched reactance of 0.4674X_{L3} and X_{mm}, the equivalent circuit of Figure 7-2 is identical to the one analysed in Chapter 6. Thus, by expressing the mismatched impedance in terms of X_{mm}, the analysis derived in Chapter 6 Section 6.3.3 can be directly used to analyse the circuit operation under variable mutual coupling conditions. The key analytical expressions, which are given below, are the output DC current (I_{DC}) and the normalised reflected impedance (Z_{r,n}). With these expressions and an estimation of α with respect to θ_1, (discussed in the next section), the real and reactive power flow of the pick-up can be described.

Referring to Chapter 6 Section 6.3.3.1, the output DC current (I_{DC}) is given by:

$$I_{DC} = \frac{2\sqrt{2} Q_I V_{oc}}{\pi \omega L_2} \sin\left(\frac{\theta_1}{2}\right) \cos \alpha$$

(7.2)

Here Q_I is the designed current Q of the LCL network [116].

Referring to Chapter 6 Section 6.3.3.2, the output real and reactive load (R_{load} + jX_{load}) are described by:

$$R_{load} = \frac{2\sqrt{2} V_{DC} \omega L_2}{\pi V_{oc} Q_I} \sin\left(\frac{\theta_1}{2}\right) \cos \alpha$$

(7.3)

$$X_{load} = \frac{V_{LCL_1}}{I_{L3_1}} \sin \alpha - X_{mm}$$

$$= \frac{2\sqrt{2} V_{DC} \omega L_2}{\pi V_{oc} Q_I} \sin\left(\frac{\theta_1}{2}\right) \sin \alpha - X_{mm}$$

(7.4)

The pick-up reflected impedance (Z_r) in terms of its output loading impedance (R_{load} + jX_{load}) is given by:

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Using (7.3), (7.4) and (7.5) the normalised reflected impedance of the circulating current controlled pick-up operating with variable $V_{oc}$, can be expressed by:

$$Z_r = \left( \frac{\omega M}{X} \right)^2 \left( R_{load} + jX_{load} \right) \quad (7.5)$$

Using (7.3), (7.4) and (7.5) the normalised reflected impedance of the circulating current controlled pick-up operating with variable $V_{oc}$, can be expressed by:

$$\text{Re}(Z_{r,n}) = \frac{\text{Re}(Z_r)}{Z_0} = \frac{2\sqrt{2}}{\pi} \frac{V_{DC}}{V_{oc}} Q_i \sin \left( \frac{\theta_1}{2} \right) \cos \alpha \quad (7.6)$$

$$\text{Im}(Z_{r,n}) = \frac{\text{Im}(Z_r)}{Z_0} = \frac{2\sqrt{2}}{\pi} \frac{V_{DC}}{V_{oc}} Q_i \sin \left( \frac{\theta_1}{2} \right) \sin \alpha - Q_i^2 \frac{X_{mn}}{\omega L_2} \quad (7.7)$$

where $Z_0$ is defined by:

$$Z_0 = \frac{\omega M^2}{L_2} \quad (7.8)$$

Using (7.2), (7.6) and (7.7), the key characteristics of the circulating current controller can be estimated.

### 7.2.2. Linear estimation of phase variation $\alpha$

In Chapter 6, two methods were used to estimate the value of $\alpha$ as a function of $\theta_1$: these are the linear approximation and the 3rd order polynomial approximation. Both approximations were proven to be effective for system design purposes but the 3rd order polynomial expression provided a more accurate modelling of the circuit’s behaviour. However, this polynomial expression cannot be adopted here for use in variable coupling applications, because it was developed specifically for a UPF LCL pick-up topology. In a variable coupling application, where the LCL pick-up is normally not designed using the UPF LCL topology described in [116, 122, 123], there are a very large number of possible $L_3$ and $C_3$ combinations. In consequence it is impractical to develop a polynomial expression and therefore the linear approximation of $\alpha$ is adopted in this chapter.

Despite operating under variable coupling conditions, the phase difference $\alpha$ with $\theta_1=0$ (which is essentially a short circuit) is $\pi/2$ for Mode I and $-\pi/2$ for Mode II regardless of the circuit parameters and the operating $V_{DC}/V_{oc}$ ratio. The linear approximation graphs of $\alpha$ for both modes, demonstrated previously in Chapter 6 Section 6.3.2, is illustrated here again in Figure 7-3 and is expressed by:
Chapter 7  A Design Example for a Circulating Current Controller in Variable coupling system

\[ \alpha = \frac{\alpha_0 \mp \pi}{\pi} \theta_1 \pm \frac{\pi}{2} \]  

(7.9)

For Mode I, the sequence of the signs are minus, plus while for Mode II, the sequence is plus, minus.

![Figure 7-3: Linear approximation of phase difference \( \alpha \) (a) Mode I, (b) Mode II](image)

When the circuit operates with \( \theta_1=\pi \), the circulating current controller behaves identically to a diode bridge rectifier as explained in Chapter 6 Section 6.3.1. An assumption was made earlier in this section that with \( \theta_1=\pi \) the input PF of the LCL network is assumed to be unity within the given range of \( V_{DC}/V_{oc} \) ratios. Therefore, the controller output load reactance \( X_{load} \) with \( \theta_1=\pi \) can be assumed to be zero for both Mode I and Mode II. Equating (7.4) to 0, the phase difference \( \alpha_0 \) at \( \theta_1=\pi \) is then given by:

\[ \alpha_0 = \sin^{-1} \left( \frac{\pi}{2\sqrt{2}} \frac{V_{oc} Q_l X_{mm}}{V_{DC} \omega L_2} \right) \]  

(7.10)

Under the chosen LCL network design (\( X_{mm} \) and \( Q_l \) is constant) and assuming it operates within the given range of \( V_{DC}/V_{oc} \) ratio, (7.10) shows that the value of \( \alpha_0 \) varies with \( V_{DC}/V_{oc} \). This means that there is a different linear approximation for \( \alpha \) as a function of \( \theta_1 \) for each \( V_{DC}/V_{oc} \) value. Thus, in practice the \( \alpha \) approximation is calculated at the given extreme points of \( V_{DC}/V_{oc} \) to estimate variations in the circuit reflected impedance for system design purposes. The analytical results obtained using the linear approximation for \( \alpha \) are compared against both simulation and practical results for a design example discussed in the next section.
7.3. Design example, Experimental Results and Discussions

The 1.5kW 300V DC EV battery charging prototype presented in Chapter 6 Section 6.5 was designed to operate at a fixed air gap of 270mm with fixed coupling between the horizontally aligned 760x400mm Double D charging pads. Here in this section, the performance of this prototype system is examined at two chosen air gaps of 220 and 270mm with the LCL pick-up modified to operate with the selected variations in air gap.

7.3.1. Specification and design of prototypes for operating with variable coupling

The parameters of the primary power supply and the magnetic details of the Double D charging pads when perfectly aligned but with the chosen air gaps of 220mm and 270mm are listed in Table 7-1 below.

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<thead>
<tr>
<th>3kW Power Supply (PS)</th>
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</thead>
<tbody>
<tr>
<td>V_{in}</td>
</tr>
<tr>
<td>Frequency</td>
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Magnectics detail of 760x400mm Double D pads

<table>
<thead>
<tr>
<th>Air gap (mm)</th>
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<th>270</th>
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<tbody>
<tr>
<td>L_1 (μH)</td>
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<td>148.79</td>
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<tr>
<td>Q_{L1} (L_1 native Q)</td>
<td>311</td>
<td>330</td>
</tr>
<tr>
<td>L_2 (μH)</td>
<td>151.36</td>
<td>149.53</td>
</tr>
<tr>
<td>Q_{L2} (L_2 native Q)</td>
<td>352</td>
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</tr>
<tr>
<td>M (μH)</td>
<td>30.27</td>
<td>20.57</td>
</tr>
<tr>
<td>k</td>
<td>0.2</td>
<td>0.138</td>
</tr>
<tr>
<td>V_{oc} (V)</td>
<td>152</td>
<td>103</td>
</tr>
<tr>
<td>I_{oc} (A)</td>
<td>8</td>
<td>5.5</td>
</tr>
</tbody>
</table>

Table 7-1: Parameters of a 3kW IPT power supply, a pair of 760x400mm Double D charging pads with air gaps of 220mm and 270mm.

With the given variation of magnetic coupling and the resulted V_{oc}, the LCL pick-up design is modified using the graphical design approach presented in Chapter 5 Section 5.3.3 to achieve a minimised variation and near unity input PF of the LCL pick-up. The designed parameters of the LCL pick-up for operating with the given variable V_{oc} is listed in Table 7-2. Because the LCL pick-up prototype in Chapter 6 was designed for operating with a V_{DC}/V_{oc} value of 3 at 1.5kW for the same pick-up magnetic structure, therefore C_2 and L_3 does not
need to be changed to operate within the given $V_{DC}/V_{oc}$ range of 2-3. The only alternation required is the value of $C_3$ to minimise its input PF.

<table>
<thead>
<tr>
<th>C_2</th>
<th>0.422μF</th>
<th>L_3</th>
<th>313.56μH</th>
</tr>
</thead>
<tbody>
<tr>
<td>X</td>
<td>18.86Ω</td>
<td>X_{C3}</td>
<td>1.6X</td>
</tr>
<tr>
<td>X_{L3}</td>
<td>2.1X</td>
<td>C_3</td>
<td>0.266μF</td>
</tr>
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</table>

Table 7-2: Parameters of the designed 1.5kW 300V DC LCL pick-up for operating with a $V_{DC}/V_{oc}$ ratio between 2 to 3

The designed LCL pick-up with full diode conduction interval $\theta_1$ was simulated in SPICE and its simulated input PF with a range of $V_{DC}/V_{oc}$ between 1.5 to 3.5 is shown in Figure 7-4. As demonstrated by the SPICE simulation, its input PF is above 0.99 for the range of $V_{DC}/V_{oc}$ ratio (between 2 to 3) of interest. With this minimal variation of PF, the input PF can reasonably be assumed to be unity with both switches $S_1$ and $S_2$ off.

![LCL Input Power Factor vs. $V_{DC}/V_{oc}$](image)

Figure 7-4: SPICE simulation results of LCL input PF versus $V_{DC}/V_{oc}$ for adopting the graphical design approach in Chapter 5

7.3.2. Experimental results and Discussion

The experimental setup for measuring the performance and the reflected impedance of the 1.5kW prototype is identical to that described in Chapter 6 Section 6.5.2. The prototype was tested individually at both air gaps of 220mm and 270mm according to the specification given in Table 7-1.

Oscilloscope captures of the prototype outputting 1.5kW at the two given air gaps are shown in Figure 7-5 and Figure 7-6 for Mode I and Mode II respectively. These captures
demonstrate that when the pick-up operates with an air gap of 270mm (which corresponds to a \(V_{oc}\) of 100V) the pick-up controller operates with a full diode conduction interval \(\theta_1\) to output the required 1.5kW. On the other hand, for both modes of the controller operating with an air gap of 220mm (corresponding to a \(V_{oc}\) of 150V) the output power is successfully controlled and regulated at the rated 1.5kW by operating with a smaller diode conduction interval (around 105° for both modes).

Figure 7-5: (a) and (c) are oscilloscope captures of circulating current controller in Mode I with 220mm air gap and \(\theta_1=107^\circ\) and (b) and (d) are with 270mm air gap, \(\theta_1=180^\circ\). Traces from top to bottom in (a) and (b) are \(I_{L3}, V_{LCL}\) and total diode output current \((I_{D1}+I_{D2})\). Traces from top to bottom in (c) and (d) are \(V_{DC}\) and \(I_{DC}\) respectively.
Chapter 7
A Design Example for a Circulating Current Controller in Variable coupling system

Figure 7-6: (a) and (c) are oscilloscope captures of circulating current controller in Mode II with 220mm air gap and $\theta_1=105^\circ$ and (b) and (d) are with 270mm air gap and $\theta_1=176^\circ$. Traces from top to bottom in (a) and (b) are $I_{L3}$, $V_{LCL}$ and total diode output current ($I_{D1}+I_{D2}$). Traces from top to bottom in (c) and (d) are $V_{DC}$ and $I_{DC}$ respectively.

The output power, the reflected reactance and the efficiency measurements of the pick-up operating in Mode I at the two given air gaps are shown in Figure 7-7, Figure 7-8, and Figure 7-9 respectively. For Mode II, the output power, the reflected reactance and the efficiency measurements at the two given air gaps are shown in Figure 7-10, Figure 7-11, and Figure 7-12 respectively. The output power and the reflected reactance profiles are compared against SPICE simulation and analytical results using the linear approximation explained in Section 7.2.2. Measurements of reflected reactance under Mode I operation at 220mm air gap have magnitude variations with similar trends but which are slightly out of phase (as illustrated in Figure 7-8 (a)). The rest of the practical measurements of both modes show good agreement with the simulation results at both air gaps. Analytical results show similar trends to both the simulated and measured results but with errors which are introduced by the modelling error introduced in the linear approximation of the phase variation $\alpha$. The errors between the measured and analytical results of the output DC current are within 10% of the maximum available output current (7A for 220mm air gap and 4.9A for 270mm air gap).
error in the analytically modelled reflected impedance is about 15% of the measured maximum reflected reactance in both modes of operation. Thus, although the analytical results show some modelling error, but the linear approximated analysis is sufficient for system design purposes.

The measured results of the prototype system, operating with the two given air gaps, indicates that the maximum reflected inductive load from the pick-up occurs in Mode I while operating with $\theta_1=95^\circ$ and at an air gap of 220mm ($V_{oc}$ of 100V) as shown in Figure 7-8 (b). This reflected inductive load is $0.42\Omega$ which is $3.34\mu H$ at $20kHz$. This is 7% of the designed inductance of the primary power supply and is within the general $\pm5$-% variation tolerance. However, as the power supply used in this prototype system is voltage fed, and operates with fixed frequency using an LCL resonant network, any reflected inductive load on the primary track will result in a capacitive loading on the power supply inverter bridge which is undesirable [117, 130]. In consequence, it is best to adjust the primary tuning network during the system design process to avoid any capacitive loading effect on the inverter bridge.

The maximum reflected capacitive reactance from the controller occurs in Mode II, while operating with $\theta_1=90^\circ$ and an air gap of 270mm ($V_{oc}$ of 150V), is -1$\Omega$, which is $7.96\mu F$ at $20kHz$ as shown in Figure 7-11 (a). This load corresponds to reducing the primary track inductance by $7.96\mu H$ (19% of the designed track inductance). This exceeds the nominal $\pm10$% variation tolerance but a power supply using an LCL tuning network has better tolerance to a lower track inductance [117, 120, 127, 155, 156]. In practice this additional VA stress on the primary power supply does need to be taken into consideration during the power supply design process, but this is outside the scope of this thesis.

The overall system efficiency measurements of the prototype systems are similar for both modes of operation at both air gaps as demonstrated in Figure 7-9 and Figure 7-12. The overall system operates with near 80% efficiency at the rated 1.5kW and is above 70% when operating at half rated power. This overall system performance is very similar to the prototype presented in Chapter 6 Section 6.5.2 and is expected to be between 85% to 90% when operating close to the 3kW designed rating of the power supply and the charging pads. The pick-up efficiency is between 92 to 95% when operating at the designed 1.5kW and remains above 85% when operating at half the designed power.
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Figure 7-7: Output DC current $I_{DC}$ of simulation, analytical and practical results versus diode conduction interval for Mode I at (a) 220mm and (b) 270mm air gap

Figure 7-8: Reflected reactance $\text{Im}(Z_r)$ of simulation, analytical and practical results versus diode conduction interval for Mode I at (a) 220mm and (b) 270mm air gap

Figure 7-9: Efficiency measurements for Mode I operation versus output power at (a) 220mm and (b) 270mm air gap
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Figure 7-10: Output DC current $I_{DC}$ of simulation, analytical and practical results versus diode conduction interval for Mode II at (a) 220mm and (b) 270mm air gap

Figure 7-11: Reflected reactance $\text{Im}(Z_r)$ of simulation, analytical and practical results versus diode conduction interval for Mode II at (a) 220mm and (b) 270mm air gap

Figure 7-12: Efficiency measurements for Mode II operation versus output power at (a) 220mm and (b) 270mm air gap

100% 90% 80% 70% 60% 50% 250 500 750 1000 1250 1500 Efficiency Output Power (W)

(a) Overall system

(b) Charging Pad and Pick-up Controller

(c) Pick-up Controller
7.4. Conclusion

In this chapter a circulating current controller with a series-parallel LCL pick-up, as presented in chapter 6, has been examined for use in an EV battery charging system with variable coupling between the charging pads. For the LCL pick-up to be used with variable coupling, the pick-up design had to be modified using the graphical design approach presented in Chapter 5 to achieve near UPF at the input of the LCL network. Under the assumption of UPF, the analysis derived in Chapter 6 can be directly adopted here with knowledge of the circuit parameter variations from the graphical design approach.

The 1.5kW 300V DC output prototype, presented in Chapter 6, was used for the performance evaluation. The air gap between the charging pads was varied between 220mm and 270mm while the $V_{oc}$ of the pick-up varies between 150V to 100V, which corresponds to a change in the $V_{DC}/V_{oc}$ ratio from 2 to 3.

The controller demonstrated its ability to regulate its output power at 1.5kW despite the increase in $V_{oc}$, and to operate with good efficiency, close to 95%. The measured power profile shows good agreement with both the SPICE simulations and the analytical results. In this practical example, when operating under Mode I the reflected reactance from the pick-up controller is inductive and is within the 10% tolerance band of the designed inductance of the primary power supply. When operating under Mode II the reflected reactance is capacitive and under worst case conditions is 19% of the nominal designed power supply inductance. A power supply with an LCL resonant topology has better tolerance to track inductance variations which cause the inductance to be lower than its designed value, so this variation is acceptable but the additional reactive loading may need to be included in the design of the primary power supply.
Chapter 8.

Conclusions

8.1. General Conclusions
8.2. Suggestions for Future Work
8.3. Concluding Remarks

8.1. General Conclusions

The goal of the research work carried out in this thesis was to develop a stationary IPT EV battery charging system for domestic household and opportunistic charging applications where high efficiency and good tolerance to misalignment are highly desired. A number of new magnetic and control solutions have been proposed for an IPT charging system that together improve its capability to transfer power to a stationary EV with good efficiency and good tolerance of movement. These are the Double D charging pad, multi-path pick-up controller and circulating current controller.

Research into means of charging EV batteries using inductive power transfer has been ongoing for the past two decades. The types of developed lumped coil systems can be broadly categorised into two groups: charging paddle systems and hands-free loosely coupled systems. The most typical charging paddle system is the MagneCharge system which formed the basis of the North American SAE J1773 standards outlined in Chapter 1 and 2 [5]. This requires a user to insert an inductive plug into the vehicle, similar to a traditional direct contact conductive coupler system, but limits the chance of using opportunistic charging, particularly in places where EV’s spend little time, or must move often, e.g: at bus stops and taxi stands.

In Chapter 1, the development of a number of loosely coupled lumped coil systems for EV battery charging were discussed. Such systems require mechanical assistance to reduce the air gap between the charging pads before charging could commence. This increases system cost and complexity, and reduces its durability. Chapter 2 undertakes a general review of IPT technologies that could be used in EV battery charging applications. Following this, this thesis focuses on possible improvements to the system, beginning with the development of a new charging pad magnetic structure, called a Double D pad, in order to enhance
coupling performance. Practical considerations for designing lumped coil systems with wide coupling variation were also investigated. This was then followed by the development of two new pick-up controllers, called a multi-path pick-up and a circulating current controller. Both are capable of regulating power flow on the secondary side while reducing the burden on the primary power supply when the system is operating under variable coupling applications.

8.1.1. Polarised Single-Sided Flux Lumped Coil System

In Chapter 3, the concept of a flux pipe is introduced, together with two flux transmitter/receiver pole areas in lumped coil IPT systems to enhance magnetic coupling for large air gap applications. Based on this flux pipe concept, two novel IPT lumped coil systems, named the flat bar type charging pad and the Double D charging pad were developed. Both systems demonstrated improved coupling performance over large air gaps compared with conventional circular charging pads. However, the flat bar type charging pad was found to be impractical because it generates magnetic flux on both sides of the pad. This makes it sensitive to nearby metallic objects so that when aluminium shielding is introduced, the losses are unacceptable for high power transfer.

The Double D pad is structured to only generate flux on one side so that aluminium shielding only removes leakage fields and has little loss. This is similar to a conventional circular pad, which has been the preferred structure for EV charging systems until now but the field lines naturally couple over twice the height of the circular pad making large air gap couplers possible without unnecessarily oversize the magnetic structure. The performance of a 760x400mm Double D pad prototype was compared with a 700mm diameter circular pad to demonstrate the improvement and the results are summarised in Table 3-9. The Double D pad prototype has superior coupling performance compared to a circular pad, with an improvement of 50%, in both vertical and lateral displacement while it is physically smaller and it uses 30% less ferrite. This improved coupling and compact lumped coil design is a key to the future success of loosely coupled EV battery charging systems.
Conclusions

<table>
<thead>
<tr>
<th></th>
<th>Circular Pad</th>
<th>Double D Pad</th>
</tr>
</thead>
<tbody>
<tr>
<td>Physical size (mm)</td>
<td>700 mm diameter</td>
<td>760 x 400 mm</td>
</tr>
<tr>
<td>$k$ range between air gap of 150 mm to 250 mm</td>
<td>0.25 – 0.11</td>
<td>0.335 – 0.16</td>
</tr>
<tr>
<td>$k$ range between perfectly aligned to 120mm offset in both horizontal direction with air gap of 200 mm</td>
<td>0.145 – 0.07</td>
<td>0.231 – 0.143</td>
</tr>
<tr>
<td>Ferrite volume ($m^3$)</td>
<td>$1.5 \times 10^{-3}$</td>
<td>$1 \times 10^{-3}$</td>
</tr>
<tr>
<td>Native quality factor at 20 kHz ($Q_L$)</td>
<td>220</td>
<td>291</td>
</tr>
<tr>
<td>Length of Litz wire (m)</td>
<td>33</td>
<td>35.1</td>
</tr>
</tbody>
</table>

Table 8-1: Comparison of 700 mm Circular and 760 mm Double D pad

8.1.2. Design Considerations for Variable Coupling Lumped Coil Systems

In Chapter 4, a number of practical design considerations for variable coupling lumped coil systems were discussed. As a result of the variable coupling and the proximity of Ferrite in each structure (which is required to enhance the coupling for power transfer), the self-inductance of both the primary and the secondary charging pads necessarily vary with changes in the air gap. This causes mistuning in both the primary and secondary resonant networks. The resultant impact of both a mistuned pick-up and primary track on the power supply was investigated and a design methodology to mitigate the additional reactive load on the inverter bridge was presented. This design methodology begins by determining the appropriate tuning of the pick-up pad and then calculates the tuning of the primary pad by including the variable reflected load from the pick-up taking into account likely pad misalignment.

A 1.2kW design example using the new design strategy with a pair of 700mm diameter circular pad was presented. The performance of the design example was compared analytically against two common tuning approaches which tune each of the pads at either the maximum or the minimum coupling position within a specified power transfer zone. The analytical results demonstrated the effectiveness of the proposed design methodology which was verified using SPICE simulations.
8.1.3. Pick-Up Controller Designs for Variable Coupling Applications

Pick-up controllers can be categorised based on slow or fast switching operation as explained in Chapter 2, each having their own strengths and weakness based on selected design aspects. In this thesis, two new pick-up power flow controllers (one from each category) were developed for applications with variable magnetic coupling called the multi-path pick-up controller (slow switching) and the circulating current controller (fast switching). Both controllers allow efficient power flow control to be implemented.

8.1.3.1. Multi-Path Pick-Up Controller

The limitations of a slow switching controller operating under variable coupling was discussed in Chapter 5. Perhaps the biggest problem is its inability to dynamically regulate the output power. However, the slow switching controller has a simple and robust controller design and low switching losses, both of which are attractive when considering the power levels required for EV battery charging applications. The multi-path pick-up consists of a series-tuned LC pick-up together with multiple LCL networks and rectifier sections that are all connected in parallel between the fully series-tuned pick-up and the DC output filter and load. The number of active LCL networks is controlled in order to maintain a desirable output power level. For LCL network with rectifier to be used with variable input voltage, a new design approach (based on [116, 123]) was developed to minimise the extra reactive power in the circuit due to variations in the ratio between input voltage and output DC voltage, which is assumed to be held normally constant by the action of the controller.

A 2kW 5.5 network multi-path pick-up prototype was designed and implemented. It demonstrated the ability to regulate the output power while the coupled pick-up open circuit voltage varied between 110 and 210V RMS. This multi-path pick-up achieved efficiencies of 96% at 2kW and above 90% for all output power ranges while maintaining a mains input power factor above 0.93 in the primary power supply and nearly all output power levels.

8.1.3.2. Circulating Current Controller for Series-Parallel LCL pick-up

In Chapter 6, a new circulating current controller for a series-parallel tuned LCL pick-up was presented. This controller consists of two diodes and two switches switching synchronously at the IPT frequency to perform both power regulation and rectification. This enables continuous power control from no load to full load for the series-parallel LCL pick-up structure as is essential for a pick-up operating under variable coupling conditions.
The circulating current controller has two control modes referred to as Mode I and Mode II in this thesis. A detailed steady state analysis based on the fundamental components of the controller was presented in Chapter 6. The controller varies the magnitude of the LCL output AC voltage and its phase relative to the LCL network output resonant current. This necessarily introduces a small amount of reactive load in the system. In Mode I, the controller reflects an inductive load back onto the track for a diode conduction interval between 40 to 180°, and a capacitive load for a diode conduction interval between 0 to 40°. In Mode II, the controller reflects only a capacitive load back onto the track and its maximum value occurs at a diode conduction interval of 90°.

A 1.5kW 300V DC prototype system designed for operating with fixed coupling was also presented. The measured profile of the output power and the reflected impedance supports the presented steady state analysis of the controller. The prototype demonstrated a good efficiency of 95% when operating at maximum load and above 85% when operating at one third of the rated power.

In Chapter 7, the circulating current controller from Chapter 6 was modified to enable its operation under variable coupling applications. This modification requires the LCL network to be redesigned using the approach presented in Chapter 5 to achieve near UPF at its input. The analysis derived in Chapter 6 can then be directly applied in the variable coupling application for system design purposes. A 1.5kW prototype was developed and tested under variable coupling conditions. The coupled voltage varies from 100V to 150V RMS while the output voltage is regulated at 300V DC. The controller demonstrated its ability to regulate its output power with changes in the coupled voltage and with good efficiency, close to 95% at rated load.

### 8.2. Suggestions for Future Work

This thesis has demonstrated the feasibility of designing IPT systems for stationary EV battery charging with good efficiency while also having a wide tolerance to physical movement and changes in the coupling in general compared with conventional loosely coupled systems. The research has investigated various techniques including magnetic design, tuning design considerations, and pick-up controller topologies. Suggestions for future extensions to the work presented in this thesis are outlined in this section.
8.2.1. Tolerance Extending Charging Pad design

A new quadrature pick-up topology, shown in Figure 8-1, was recently proposed for a distributed track in an Automatic Guided Vehicle (AGV) application [30, 66, 161]. This topology consists of one coil capturing the magnetic flux in the vertical direction and one coil capturing the flux in the horizontal direction [30, 66, 161]. This greatly enhances the lateral movement tolerance of the pick-up.

![Figure 8-1: Quadrature pick-up design for distributed track AGV applications](image)

This concept could be used in the Double D pick-up pad to improve its lateral tolerance relative to the primary in the x direction of the pad (indicated in Figure 8-2 (a)). This limitation of lateral tolerance in the x direction was explained in Chapter 3 and is due to the nature of the generated flux profile of the primary pad. By adding a quadrature coil, illustrated in Figure 8-2 (b), in the pick-up pad the tolerance to lateral movement can be improved. Note that this quadrature coil is where the usual pick-up coil in a circular pad would be expected to be. Under normal circumstances while the pick-up pads are perfectly aligned laterally, the Double D coil picks up the magnetic flux in the horizontal direction. When the pick-up pad has a misalignment in the x direction, the quadrature coil moves to be more on top of one of the pole areas to pick-up the vertical magnetic flux [147, 148]. The concept of combining the Double D pad with a quadrature coil is illustrated in Figure 8-2 (c). Preliminary results demonstrate a promising improved performance in lateral tolerance which is very attractive for the application of Roadway Powering of EVs (RPEVs) [72] as discussed in the next section.
8.2.2. Dynamic Powering of Electric Vehicles on a Rowadway

“Opportunity charging” of EVs helps to reduce the required onboard storage of EVs by maintaining the battery state of charge above a critical level. In order to minimise the need for onboard storage and the depth of battery discharge, the concept of RPEVs using multiple lumped coil systems was outlined in [72]. This concept proposes installing the Double D pads along a highway as illustrated in Figure 8-3. Compared with conventional distributed track systems for AGV applications, this concept distributes lumped coils along the highway, which coils can be selectively energised underneath an EV to couple power as required in order to achieve a similar coupling performance, for both vertical and lateral movement, to a stationary lumped coil system.

In this proposed system, a single primary power supply is used on the side of a roadway to drive multiple pads but as stated only the pad that is directly underneath the vehicle needs to be energized to minimise standing loss in the system. This selectivity ensures that a conventional vehicle can travel along the roadway without acting as an undesirable load since there is no heating of the underside caused by magnetic fields being generated above the road surface.
Figure 8-3: Conceptual diagram of distributed lumped coil system for Roadway powering IPT system

The goal of an RPEV IPT system is to supply continuous power to a vehicle along the highway with a coupled uncompensated power ($S_U$) as wide and smooth as possible. To realise this, a thorough analysis as to the size, shape and positions of the charging pad along the highway is needed. These factors can be grouped into two main criteria which contribute to the fluctuation of $S_U$ profile along the highway. These are the spacing between pads and the skew angle of the pad as shown in Figure 8-4 (a) and (b) respectively.

Figure 8-4: Conceptual diagram to illustrate the pad layout criteria in a Dynamic Roadway IPT System
(a) spacing between pads and (b) pad misalignment in x direction

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The spacing between each primary pad cannot be too close due to the possibility of flux cancellation and cross coupling between active and inactive primary pads. However, the spacing cannot be too large to avoid a discontinuous or highly fluctuating SU power profile as illustrated conceptually in Figure 8-5 (a) for two different pad spacing A and B.

![Figure 8-5: Conceptual SU profile along the highway in the y direction of the two main criteria spacing between pads](image)

While the vehicle moves down the highway in the centre of the lane, the pick-up pad is aligned laterally with the primary pad. Under this situation, only the Double D coil is transferring power because the quadrature coil has insignificant coupled SU. While the vehicle is driving near to the boundary of the vehicle lane, the quadrature coil is responsible for the majority of the output power. Hence, there will be a continuous fluctuation of coupled power between the Double D coil and the quadrature coil with lateral misalignment. This can be improved by deliberately adding skew to the pick-up pad, relative to the primary pads in the roadway, as shown in Figure 8-4 (b) to ensure that both coils are active at all times.

While skew of a polarised magnetic structure may be an important variable, it has never been investigated in any practical application, particularly since with stationary charging, the skew is negligible [38]. Thus, if as discussed earlier the primary pad and the pick-up pad are each skewed (a small angle) but in opposing directions as illustrated in Figure 8-4 (b) to reduce the fluctuations in the SU profile of the Double D coil and the quadrature coil with lateral misalignment, this improvement would likely come with a reduction in the maximum coupling. This is because by skewing the pads in the opposite direction to each other, the quadrature coil may have a wider SU profile for capturing more flux in the x direction but the skew may also affect the coupling performance of the Double D coil. However, compromising the peak coupling performance to reduce the magnitude of fluctuations in the power profile is seen as an advantage in RPEV applications and is an interesting avenue of future research.
8.2.3. Comparison of pick-up controllers for EV battery charging application

The developed multi-path pick-up and circulating current controller both have their strengths and weakness. A comparison between these two is required in order to ensure the selected controller is best suited for a given design specification. One important design criteria is the operating frequency.

With advances in silicon carbide (SiC) power transistor technology [162], a high power IPT system operating at a higher frequency (between 80 to 150kHz) is feasible in the near future. The advantage of operating at higher frequencies well above 10-40kHz as presently used is that less ferrite and Litz wire are required and this directly relates to the weight of the charging pad. However, a frequency increase necessarily complicates the design of the circulating current controller which switches synchronously with the primary operating frequency. Furthermore there are limited available capacitors suitable for operations at higher frequencies. On the other hand, the multi-path pick-up controller could fully utilise the frequency increase to improve its power density by reducing the physical size of its passive components. The control algorithm can also remain the same as it “slow-switches”.

8.3. Concluding Remarks

The developed techniques presented in this thesis demonstrate the feasibility of using a fixed frequency, single pick-up IPT system to transfer power efficiently across a large air gap, with variable coupling, for EV charging applications. These techniques include magnetic designs of the charging pad, tuning network design and two pick-up control topologies and they can be directly used for applications where a single power supply is required to deliver power to multiple vehicles. The research presented in this thesis is still ongoing and recommendations have been made for further extending the developed techniques for RPEV applications.
Appendix A. Derivation

This section presents the derivation of the reflected impedance of a mistuned parallel-tuned pick-up shown in Chapter 4 (4.15).

The conceptual diagram of a mistuned parallel-tuned pick-up and its Norton equivalent circuit diagram are shown in Figure A-1.

When \( L_2 \) and \( C_2 \) are tuned at the operating frequency \( \omega \) the equivalent impedance seen by the current source \( I_{sc} \) is given by:

\[
Z_2' = \frac{1}{\frac{1}{R_{AC}} + j\omega \Delta C_2} = \frac{1}{R_{AC}} - j\omega \Delta C_2
\]

Therefore, the pick-up resonant voltage \( V_2 \) is given by:

\[
V_2 = I_{sc} Z_2'
\]

Using \( V_2, V_{oc} \) and \( L_2 \) the pick-up coil current \( I_2 \) is described by:
The input impedance $Z_2$ is defined by:

$$Z_2 = \frac{V_{oc}}{I_2}$$

and the reflected impedance $Z_r$ is defined by:

$$Z_r = \frac{\omega^2 M^2}{Z_2}$$

Using the definition of $Z_2$ and $Z_r$, the mistuned parallel-tuned pick-up reflected impedance ($Z_{rp}$) is then given by:

$$Z_{rp} = \frac{\alpha M^2}{\omega M I_1} \left( \frac{1}{R_{ac}^2} - j \alpha C_2 \right)$$

$$= \left( \frac{L_2}{R_{ac}^2} + L_2^2 \left( \alpha C_2 \right)^2 \right) - j \omega M^2 \left( \frac{L_2}{R_{ac}^2} + L_2^2 \left( \alpha C_2 \right)^2 \right) - j \omega M^2 L_2$$
References


References


[99] 大聖 泰弘. エネルギー使用合理化技術戦略的開発エネルギー有効利用基盤技術先導研究開発非接触給電装置の研究開発. 東京都, 早稲田大学, 2005, p. 16.


