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AC Processing Controllers for IPT Systems

By

Hunter Hanzhuo Wu

A thesis submitted in partial fulfilment of the requirements for the degree of Doctor of Philosophy in Electrical and Computer Engineering, The University of Auckland, 2009

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Sincere thanks to Mum and Dad for all their support
Abstract

Inductive Power Transfer (IPT) technology allows electrical energy to be transferred between two loosely coupled inductors over relatively large air gaps. An IPT system can be divided into two sections – a primary supply and one or more secondary pickups and controllers. Currently, IPT applications have been used in a variety of industrial and commercial applications.

This thesis proposes a novel AC processing controller which directly regulates power in AC form, hence producing a controllable high-frequency AC source. The pickup has significant advantages in terms of increasing system efficiency and reducing pickup size compared to traditional pickups that also produce a controlled AC output using complex AC-DC-AC conversion circuits.

The parallel AC processing pickup employs switches operating under Zero Voltage Switching (ZVS) conditions to clamp parts of the resonant voltage across a parallel tuned LC resonant tank to achieve a controllable AC current source. The derivation of key power electronic specifications such as component ratings, output harmonic content and power factor are shown along with the pickups normalized characteristics. Practical implementation aspects such as a new synchronization scheme using clamp time, series and parallel connection of switching devices to achieve higher ratings and a new PWM control technique to reduce conduction losses are investigated. By adding a rectifier, a controlled DC output can be produced. Practical examples including a lighting system and an EV charging system have shown the pickup to achieve very high operating efficiencies above 96%.

The series AC processing pickup uses an AC switch operating in series with a resonant network to produce a controllable AC voltage source. When a rectifier is cascaded onto this pickup, it can also produce a precisely controlled DC voltage. The circuit is analytically analyzed and the maximum efficiency for a 1.2kW prototype is measured to be 93%.
A direct AC-AC IPT system which takes 50Hz mains input and has 50Hz mains output without requiring a DC link is also proposed. This technique is based on the AC processing concept with matrix converters giving it the advantage of high efficiency. Other AC-AC systems at the track frequency are also proposed to enable controllable intermediate IPT links suitable for powering separate IPT tracks. Both of these later systems, appear to have significant future potential but need much further study to understand and overcome practical limitations which could form the basis of separate Ph. D. studies in their own right.
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- Hunter Hanzhuo Wu

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Nomenclature

Acronyms

AGV  Automated Guided Vehicle.
EV   Electric Vehicle.
EMI  End Effect Voltage Offset.
ESR  Internal Combustion Engine.
IGBT Insulated Gate Bipolar Transistor.
IPT  Inductive Power Transfer.
LCL  Inductor-Capacitor-Inductor connection.
MOSFET Metal Oxide Silicon Field Effect Transistor.
PI   Proportional and Integral control.
PLL  Phase Lock Loop.
PWM  Pulse Width Modulation.
Q    Quality Factor.
rms Root Mean Square.
ZCS  Zero Current Switching.
ZVS  Zero Voltage Switching.

Symbols

$\alpha$ Ratio of actual tuning capacitance and nominal tuning capacitance.
$C_1$ The track (primary) tuning capacitor.
$C_2$ The pickup (secondary) tuning capacitor.
$C_{20}$ Nominal pickup tuning capacitance.

$C_{dc}$ DC capacitor in pick-up regulator circuit.

$f$ IPT system frequency.

$\phi$ Controlled phase delay angle.

$I_1$ Primary (track) current.

$I_2$ Current in the pick-up coil.

$I_{L_{dc}}$ DC current flowing through DC inductor.

$I_{sc}$ Short circuit current of the pick-up coil.

$k$ Coupling coefficient.

$L_1$ The track inductance.

$L_2$ The inductance of the pick-up coil.

$L_{20}$ Nominal inductance of the pickup coil.

$L_{dc}$ DC inductor in pick-up regulator circuit.

$M$ Mutual inductance.

$N_1$ Number of turns in the track inductor.

$N_2$ Number of turns in the pickup coil.

$Q_2$ Quality factor of the pickup tuned circuit.

$Q_{20}$ Nominal quality factor.

$\theta$ Phase between $V_{oc}$ and fundamental component of $V_c$.

$R_{ac}$ Equivalent AC resistance of rectifier and load.

$R_L$ Load resistance on the pickup regulator.

$R_r$ Reflected resistance of the pickup onto the track.

$S_U$ Uncompensated power output of the pickup coil (defined as $V_{oc} \cdot I_{sc}$).
\( T \)  
IPT system period.

\( V_{dc} \)  
DC output voltage of the pick-up regulator.

\( V_c \)  
Voltage across the tuning capacitor of the pickup.

\( V_{oc} \)  
Open circuit voltage of the pickup coil.

\( \omega \)  
Operating frequency of the power supply.

\( \omega_f \)  
Damped resonant frequency.

\( X_r \)  
Reflected reactance of the pickup onto the track.

\( Z_r \)  
Reflected impedance from the pickup to the track.
1 Introduction

1.1 Background

Inductive Power Transfer (IPT) technology enables the delivery of contact-less electric power from a primary source to one or more galvanically isolated secondary loads using time varying magnetic fields. The fundamental laws which govern IPT technology were founded by Ampere and Faraday over 100 years ago. Ampere found that a magnetic field is generated when current flows through a conductor while Faraday discovered that a time varying magnetic field in a coil can induce an electric voltage potential across it [1, 2]. By understanding the combination of these two laws, the essence of IPT systems is easily understood. In IPT systems, a time varying magnetic field is generated when high frequency current flows through a conductor and the changing magnetic field will in turn induce a voltage across a secondary coil. This induced voltage source is the basis of electrical power transferred to the load. Thus, electrical power can be transferred over a large air gap without any physical contact. This is similar to the operating principle of a conventional transformer. However, the very low coupling nature of IPT systems has made it difficult to realize power transfer in an efficient and cost effective manner [3, 4].

An IPT system consists of two electrically isolated parts as shown in Figure 1.1. The primary side comprises a resonant power converter which generates a high frequency current in a track loop. The secondary side consists of an inductor coil tuned to resonance at the track frequency and a power regulator. The time varying magnetic field produced by the primary track current links the primary and secondary to deliver electrical energy over an air gap. After the power is received by a pickup inductor, the pickup regulator converts it to a desirable form to drive a load.
1.2 IPT Applications

The development of power electronics technology in the last few decades has allowed IPT systems to be utilized in residential, commercial and industrial applications. Some of these applications include contact-less recharging of electric vehicles and wireless powering of biomedical implants. Other applications include people movers, Automatic Guided Vehicles (AGV), road studs and materials handling systems. The power rating for these systems varies from a few hundred milli-watts in biomedical applications, to several kilo-watts in Electric Vehicle (EV) applications. The distance of power transmission for these applications ranges from a few millimeters to a few hundred millimeters. Some of these applications are discussed below.

For materials handling and AGV applications, any movement of the secondary usually follows a rigidly defined path given by the location of the primary track. Electrical power is continuously transferred to the secondary load as the pickup moves along the track. A Japanese company Daifuku has successfully commercialized the technology for clean room applications and some of these systems are shown in Figure 1.2. Currently, more than 80% of clean rooms worldwide use this technology because of its advantages in terms of dust free,
Chapter 1. Introduction.

maintenance free, safe, and robust operation.

![Figure 1.2. Clean Room IPT Systems including (a) Material Handling System and (b) AGV.](image)

A road stud is the most common form of IPT lighting application used for traffic guidance. The road stud is used to delineate the traffic lanes very clearly in poor weather conditions such as fog, rain and snow. A primary track is buried under the ground and a pickup is positioned on the surface of the highway to receive electrical energy to power onboard Light Emitting Diode’s (LED’s). The road stud has been used in many countries and some of these applications are shown in Figure 1.3.

![Figure 1.3. IPT Road Stud Systems in (a) Terrace Tunnel, Wellington, New Zealand and (b) in Intersection Traffic Guidance, USA.](image)

In EV applications, movement of the vehicle is usually very flexible, so on board energy
storage using batteries and super capacitors is required. To recharge the energy storage, a plug based charging system can be used to recharge the batteries. However, normal members of the public with little or no training may have concerns about potential safety hazards when connecting high power charger plugs with power ratings that are much higher than the standard mains plug. Conversely, IPT provides an ideal solution as the inherent electrical isolation provides important safety features. In addition, it is now possible, and convenient to recharge the batteries automatically as the vehicle enters the charger bay without the user ever forgetting to recharge the vehicle. A typical recharging configuration using IPT technology is shown in Figure 1.4. The primary charger pad is located at either a house garage or a recharge station. The secondary charger pad is attached under the vehicle. The recharging process begins when the secondary charger pad moves on top of the primary pad. The recharging time usually takes between a few hours to half a day depending on the power level of the IPT system.

![Recharging Platform for Off-road Electric Vehicles (EV)](image)

**Figure 1.4. A Recharging Platform for Off-road Electric Vehicles (EV).**

In biomedical applications, electrical power at about 5-20W is required to keep an implanted heart-assist device operating continuously. A heart assist device is a mechanical pump which helps a failing heart to circulate blood through the body. The traditional method of delivering power to implants is to use percutaneous wires from the outside of the skin to deliver power to the implanted device inside the patient’s body. A major problem is that the percutaneous wire penetrating the skin is the source of many infections when it is exposed to
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The outside environment over extended periods of time. The connected wires also constrain mobility of patients because the heart assist device has to remain in a very clean environment during the treatment process to reduce the likelihood of infection. Here, IPT technology can offer an excellent solution to eliminate the possibility of the skin infection caused by percutaneous wires [5]. Another name for the same technology, used in biomedical applications is Transcutaneous Energy Transfer (TET). As shown in Figure 1.5, a TET system has a small primary coil outside the human body driven by an external driver to generate a high frequency magnetic field, and another small secondary coil under the skin to pickup power from the field and regulate it to a form suitable for driving the implanted heart assist device. Although there are strong patient health advantages related to this new technology, whether it can actually be used in patients depends on many factors such as the reliability, power efficiency, and implant size [5].

![Figure 1.5. An IPT Biomedical System.](image)

1.3 IPT System Overview

1.3.1 Magnetic Coupling Structures

Magnetic coupling is a measure of how well the magnetic field linkage is between the primary and secondary of the system. It is governed by the geometrical structure of the
primary track and secondary pickup inductor [6]. In IPT systems, a better coupling can be summarized as having two advantages. Firstly, for a given amount of power transferred to the load, an enhancement in coupling will reduce the required primary track current. This will substantially decrease both the $I^2R$-losses in the primary track and the component losses in the primary converter. Secondly, better coupling increases the amount of power the IPT system can transfer. This property enables an improvement in the overall efficiency or the cost of the IPT system [5].

Magnetic cores can be used to enhance the linkage of the magnetic field or coupling between the primary and secondary. Magnetic cores have advantages in terms of increasing coupling between the system and reducing the generated Electromagnetic Interference (EMI) to other systems [7]. The most common type of material used for magnetic cores in IPT applications has been ferrites. These materials have very low eddy current and very low hysteresis losses which make them ideal for high frequency applications such as IPT. In addition, the conductor windings for the coil usually use Litz wire to mitigate skin and proximity effect at high frequencies [8].

There are two coupling configurations commonly used in IPT applications. These can be classified into distributed systems that have multiple secondary pickups on one distributed primary track, or lumped systems that have one secondary pickup coupled to one lumped primary coil. Each of these configurations is shown in Figure 1.6. Multiple pickup systems are usually used in material handling, people moving, AGV, and road stud applications [9-12]. In these systems, many secondary pickups are powered by one single primary converter and track. For systems that require on board energy storage such as biomedical and Electric Vehicle applications [13-19], each secondary pickup is powered by one primary lumped coil.
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For an IPT system with distributed track, the ferrite structures for the secondary pickup have many shapes depending on different applications. Common types of ferrite structures are shown in Figure 1.7. E and H cores are used in industrial applications where the secondary pickup is running along a primary track with some sections of the core slotted into the cable [11]. The E and H cores have been recently improved to an S core which has a much tighter coupling [20, 21]. Lastly, flat-E cores are used in applications where the track is buried and small lateral movements of the pickup along the track are required, however this tolerance to movement comes at the cost of lower coupling [3, 22].
In a lumped primary coil IPT system such as a biomedical application, the coils are normally in the shape of a pancake [23-31]. This coil may use additional ferrite to enhance coupling although this results in a heavier and larger pickup. Unlike the distributed track system, where the secondary pickup moves along the track with a fixed separation, the lumped coils can move both in the vertical and horizontal directions. Misalignments between the coils can cause the coupling of the whole system to change over several hundred percent in some applications [29, 32]. The main objective when designing such systems is to keep the overall coupling as high as possible for the range of alignments in both the vertical and horizontal directions.

In the past, many authors have discussed the design and optimization of TET coils for biomedical applications. A full design process for the transcutaneous transformer using simple mathematical equations is outlined in [33]. Soma [29] and Donaldson [34], introduce both simple and efficient mathematical approaches to optimizing the coil dimensions. Grover [35], outlines a complete set of normalized calculations which cover almost all coil structures. In addition, mathematical analysis of coil optimization in terms of dimensions and number of turns have been proposed [36-45]. These cover a wide range of coil structures using many different analytical analysis techniques. In summary, over the years, many breakthroughs in analysis and optimization techniques have helped develop physical insights on the design of transcutaneous transformers.
1.3.2 Primary Power Converter

The primary power converter described in IPT systems refers to the power supply shown in Figure 1.1. This converter governs the operating frequency and primary track current. For a given coupling condition, a higher operating frequency or primary current allows more power to be delivered to the load. There are many different primary converters used for medium to high power industrial applications. These include current fed converters [46-52], voltage fed converters [53] and LCL inverters [54-56] as shown in Figure 1.8, Figure 1.9 and Figure 1.10. The current fed converter, shown in Figure 1.8, has advantages in terms of concise circuitry and low cost [25, 48], and is also commonly used in biomedical applications [57]. In addition, high efficiency operation is maintained because only the real current is sourced through the switches and the high reactive track current is confined to the LC tank. Voltage fed converters can tolerate systems with long tracks as the voltage across the track inductor can be increased beyond the input voltage of the power supply [4]. However, high resonant current flows through both the switch and the track, which lowers the efficiency [4]. Recently, a new type of LCL inverter using a composite tuning topology has been proposed. This converter has the benefits of a current source property when driving the track and presents unity Displacement Power Factor (DPF) to the input stage [55].

![Figure 1.8. Current Fed Converters showing (a) Full Bridge and (b) Push-Pull Converter.](image-url)
For biomedical applications, any system running with a higher primary current will lead to a lower system efficiency. On the other hand, an increase in operating frequency will not lead to a reduction in efficiency if the converter topology is chosen carefully. Hence, research over the years has concentrated on investigating new circuit topologies which can operate at higher frequencies without significantly compromising efficiency. In the first description of a primary converter in the late 1960’s [58], a very lossy hard switched converter was proposed to generate high frequency currents at 30 kHz and its switching losses increased linearly with increase in operating frequency. From the 1980-1990’s, inefficient hard switching techniques were soon substituted with soft switching operation enabling these converters to easily operate around 100 kHz while maintaining high efficiency operation [15, 30, 59-62]. In the past two decades, biomedical power supplies have utilized the class-E converter proposed by the Sokal’s [63], due to its very high efficiency operation at frequencies much higher than other previously published converters. The class-E converter operates from 1 – 10 MHz and
Chapter 1. Introduction.

achieves conversion efficiencies beyond 90% [14, 23, 27, 28, 64-77]. In addition, the low harmonic output, low sensitivity and low component stress has truly made this converter the most popular for many TET systems today. There are many publications on the design and analysis of the class-E converter written over the decades [69, 76, 78-80]. TET systems using the class-E converter are described in [14, 23, 27, 28, 64, 66-68, 72, 81].

![Class E Converter Diagram](image)

**Figure 1.11. A Class E Converter.**

### 1.3.3 Secondary Pickup

The secondary pickup converts AC power received from the pickup coil to DC, enabling most electronic devices to be driven directly. The secondary inductor coil is tuned with either a parallel or a series tuning capacitor to increase the output power by a factor known as the quality factor $Q$. For the parallel tuned pickup, the output voltage is boosted by $Q$. Similarly, for the series tuned pickup, the output current is boosted by $Q$ [4]. In a practical system, the highest $Q$ reachable is limited to about 10 as the pickup operational bandwidth is inversely proportional to $Q$ and component tolerances make higher $Q$ circuits impractical. Furthermore, the VA of the pickup is proportional to $Q$ squared so that practical size and thermal ratings also naturally limit $Q$ in commercial systems.
After the tuning stage, a rectifier and some extra filtering components are used to produce a DC output from the high frequency AC source. The power losses in the diodes of the rectifier are significant in the pickup stage because the ratio of the diode forward voltage over the output voltage is high. Recently, the rapid development of MOSFET devices has generated a new area of research in synchronous rectifiers. Some highly efficient pickups designed with synchronous rectifiers are outlined in [61, 82, 83].

1.3.4 Power Flow Regulation

The purpose of power flow regulation is to provide the load with the power it needs on a moment by moment basis. The coupling condition and load resistance will usually change depending on changes in alignment between the coils and the power required by the load. As a result, the output voltage delivered to the load will vary without a controller. A higher output voltage may damage the device and a lower output voltage may not be enough for proper operation. Hence, power flow regulation must be used for delivering the amount of power the load instantaneously requires rather than its average power demand. Power flow controllers can be implemented on either the primary or the secondary side. The first description of such a power regulation circuit was outlined in the late 1980’s using a linear regulator [84]. This inefficient technique was very soon replaced with a system which tuned and detuned the resonant tank by switching a capacitor in and out of the system [13]. Boeing’s patent also describes a possible technique for regulating power using a switch mode converter to control
Chapter 1. Introduction.

phase delay [85]. Soon after, a novel decoupling control technique was proposed by Boys et al whereby a shorting switch was used to directly decouple the pickup from the primary converter [12, 86]. This technique is very simple and easy to implement for most high power IPT systems. Recently, Si and Hu [25], proposed a dynamic tuning system, whereby a variable capacitor is produced by switching a fixed capacitor with an AC switch. Power regulation using primary current control to adjust power transfer is also used in TET systems [28, 64, 68, 72, 87].

For medium to high power systems with multiple pickups, secondary side control is the only option as primary side control cannot deliver the exact amount of power each pickup needs. Boys’ decoupling technique is a common method to regulate power. In Electric Vehicle applications, however the decoupling control technique may have high losses under varying coupling conditions forcing the components to be overrated when significant increases in the coupling condition arise. A more complete discussion on this section will be provided in chapter 2 of the thesis.

1.4 Motivations of the Thesis

The objective of the thesis is to improve the overall performance of an IPT system by implementing power regulation techniques using secondary pickup controllers. This includes improvements in power transfer efficiency and capability of the pickup to handle large variations in coupling. This research has a broad range of applications from high power EV’s to low power biomedical systems.

Presently, most tuned IPT pickups can deliver a controlled DC output. If a controlled AC output is required to drive a light (as an example), an extra DC-AC inverter is required to produce a controllable AC source. This conversion topology requires additional components that increase cost and lower efficiency. Other circuits less commonly used to regulate power deliberately tune or detune the pickup using a variable inductor. Although this technique can
vary the AC power directly, the use of a variable inductor in the non-linear region of the B-H curve may limit the efficiency of the overall system. In addition, the variable inductor is expensive to manufacture because it has to manage the high resonant current without completely saturating.

Large changes in coupling of the order of a few hundred percent are quite common in both EV and biomedical applications. Existing secondary controllers have limited ability to cope with extra power transferred from the primary, as such a very inefficient design technique of overrating most components for normal operation is used.

1.5 Contributions of the Thesis

As a consequence of the above considerations, this thesis proposes a new AC processing pickup which has several advantages. Firstly, the controller can control extra power transferred over very large coupling variations without overrating any devices in the system. This could prove extremely useful in both EV and biomedical applications. Secondly, the proposed regulator can directly generate a controlled AC voltage or current source with a very simple circuit configuration. This simple circuit structure with soft switching operation dramatically increases the operating efficiency of the system compared to traditional AC-DC-AC conversion topologies. Lastly, the switching action in parallel with a fixed tuning capacitor generates an equivalent variable capacitor which can increase the operating $Q$ of the pickup to very high levels compared to previously described techniques.

1.6 Scope of the Thesis

This research project will thoroughly investigate the control and operation of an AC processing pickup. The research topic is focused on the theoretical development of the pickup.

Chapter 2 is a full literature review on all power regulation topologies used to date. A summary of the advantages and disadvantages relating to both primary and secondary side
control is given. In addition, a discussion of controllers used in different applications is outlined.

Chapter 3 presents the operation of a parallel AC processing pickup in steady state. A full steady state analysis of the system is presented. The derivation of key power electronic specifications such as component ratings, output harmonic content and power factor is shown. Both a simple and an exact circuit analysis are proposed to investigate the operation of the circuit.

Chapter 4 investigates practical implementation strategies for the parallel AC processing pickup. A new synchronization scheme using clamp time detection is proposed. In addition, normalized graphs are produced to aid designers. Furthermore, both series and parallel connection to enable switching devices to achieve higher ratings in the AC processing pickup are outlined. Moreover, a new PWM control technique is introduced to reduce conduction losses in the AC switch. Lastly, the design of a 1.2kW lighting system is described to verify the proposed AC processing pickup.

Chapter 5 discusses the operation of an AC processing pickup with a rectifier and DC filtering components to produce a controlled DC output. The steady state operation along with full design procedures is derived in this section. A practical EV battery charging system is used to verify the proposed design technique.

Chapter 6 investigates a new series AC processing pickup that provides either a controlled AC or DC voltage source to the load. This new pickup topology allows the output voltage to be controlled without the need of an extra buck converter stage. In addition, it provides a solution to the control of the transient inrush current which is present in conventional series tuned IPT pickups.

Chapter 7 investigates the use of the AC processing pickup in an IPT system that directly outputs 50Hz AC without significant DC energy storage. A matrix converter is used in the secondary pickup to convert the high frequency AC waveform to 50Hz AC while achieving
power flow control. Theoretical analysis is presented and simulation results are also presented.

Chapter 8 summaries the conclusions obtained in this thesis. Future research directions are also provided.
2 Overview of IPT Power Controllers

An essential part of an IPT system is the power flow controller. This chapter provides a review on the current state of the art for controllers. The purpose of using a power controller is to control the amount of power delivered to a load using a reference, rather than an arbitrary amount, dependent on the system parameters of the IPT system. Some of the common parameters in an IPT system which vary the instantaneous power transfer in the IPT system can be grouped into parameters that change relatively quickly, typically from several microseconds to seconds, and parameters that change over longer periods of time, typically over weeks or even years.

The fast changing parameters usually vary when the IPT system is in operation and are listed below

1. Load Resistance – the power required by the load can change significantly depending on the application. One common load that has a significant variation on the equivalent load resistance seen at the output of a pickup regulator is the electric motor on a moving trolley for material handling systems [88]. For example, the electric motor may require the maximum power when the trolley is accelerating near its maximum predetermined speed, but under standby conditions, the motor requires no power as the trolley stays stationary in a fixed location. For most practical systems, the instantaneous variations in power demand cannot be predicted during the design phase, and closed-loop feedback systems are required.

2. Coupling Condition – the power that can be transferred to the load changes significantly as the coupling condition in the system changes [6, 17, 89]. Variable coupling conditions normally arise in lumped IPT systems where ferrite is present both in the primary and secondary coils. In such systems, coupling varies with the relative movement of the secondary coil with respect to the primary coil during operation. Many lumped IPT applications (from high power electric vehicle battery charging systems to low power biomedical implant
systems) exhibit this behavior when the alignment of the primary and secondary coils or charge pads is unpredictable.

The slow changing parameters usually vary very slowly. The IPT system can regard this kind of variation as a fixed error that the system has to work with during operation. Some of these variations are listed below:

1. Component Aging – components used in any application age over time. Typical examples of these are capacitors that change in value as they degrade [90]. Other examples of this include current or voltage sensors, which may lead to changes in primary track current and system operating frequency.

2. Component Tolerance – any circuit when mass manufactured always has a finite tolerance due to the incentive to reduce cost. Hence, the IPT system always has to operate with some tolerance.

These two behaviours identified above must be considered differently during design. As an example for fast changing parameters, the controller must use some sort of feedback loop to monitor the output voltage or current and control to a fixed reference. Slow changing parameters can be considered a fixed error in the IPT system. These errors must be tolerated and handled by the controller during normal operation. As a result of these fast and slow changing parameters, the IPT system requires a power flow controller to keep the output voltage or current constant depending on the requirements of each specific application.

This chapter begins by introducing tuning topologies because the type of controller that can be used is dependent on tuning. Following this, a review of all power flow controllers will be presented along with an outline of their existing limitations.

2.1 Tuning Topologies

For an IPT pickup inductor coupled onto a primary track, the equivalent electrical model is a voltage source in series with the self inductance of the pickup ($\omega L_2$). The Thevenin and Norton equivalent circuits for the pickup can now be drawn as shown in Figure 2.1. Note that
while these circuits are electrically equivalent, the circuit (a) corresponds to real components (voltage and inductance) while circuit (b) has purely virtual components that do not physically exist.

Figure 2.1. Equivalent Circuits of IPT Pickup in (a) Thevenin Form and (b) Norton Form.

Power can be transferred by connecting a load directly to the pickup inductor. But, due to the large self inductance, the maximum output power will be limited. The maximum output power occurs when the load resistance \( R_2 \) is equal to the reactance of the self inductance, and can be expressed as:

\[
P = \frac{V_{oc} I_{sc}}{2}
\]

(2-1)

where \( V_{oc} \) and \( I_{sc} \) are the open circuit voltage and short circuit current, respectively. In order to improve the power output capacity, some compensation, or tuning in the pickup is necessary.

Series and parallel tuning are shown in Figure 2.2. As the name suggests, compensation capacitors are either connected in series or parallel with the pickup inductor or coil. For these two compensated pickups, basic power transfer properties are listed in Table 2.1. Note that these results assume perfectly tuned conditions. The \( Q_2 \) for the series tuned circuit is defined as \( \frac{\omega L_2}{R_2} \) whereas the \( Q_2 \) for the parallel tuned circuit is defined as \( \frac{R_2}{\omega L_2} \). It can be seen that a fully series tuned pickup yields a pure voltage source while a fully tuned parallel pickup gives a pure current source. Therefore, theoretically they have unlimited power output capability. However, practically if the load resistance is too small for the series tuned pickup or too large for the parallel one, the quality factor \( Q_2 \) of the circuit would be very large. A large \( Q_2 \) will make the tuning difficult and also make the system too sensitive to parameter
variations. Therefore, the maximum power that can be transferred is limited by a specified maximum \( Q_2 \), which is normally limited to less than 10 in practical applications as discussed in chapter 1. In addition, as \( Q_2 \) increases, the VA rating for the tuning components also increases.

\[
R_{ac} = \frac{\pi^2 R_{dc}}{8}
\]

(2-2)

Similarly, for a series resonant tank with a diode bridge rectifier and a DC capacitor output

Table 2.1. Output Properties of the Different Pickup Compensations.

<table>
<thead>
<tr>
<th>Outputs at Maximum Power</th>
<th>No Compensation</th>
<th>Series Compensation</th>
<th>Parallel Compensation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage</td>
<td>( V_{oc}/\sqrt{2} )</td>
<td>( V_{oc} )</td>
<td>( Q_2 V_{oc} )</td>
</tr>
<tr>
<td>Current</td>
<td>( I_{oc}/\sqrt{2} )</td>
<td>( Q_3 I_{oc} )</td>
<td>( I_{oc} )</td>
</tr>
<tr>
<td>Power</td>
<td>( V_{oc}I_{oc}/2 )</td>
<td>( Q_3 V_{oc}I_{oc} )</td>
<td>( Q_2 V_{oc}I_{oc} )</td>
</tr>
</tbody>
</table>

The most common type of load in industrial applications requires a controlled DC output. To power a DC load, a rectifier is used to transform the AC voltage and current into a DC voltage and current, the harmonics of which can then be reduced by employing an output filter.

To simplify the resonant circuit analysis with a rectifier, an equivalent resistance model can be developed for the rectifier to represent the pickup loading system [7]. The parallel resonant tank uses a diode bridge rectifier with a DC inductor output filter. If it is assumed that the DC inductor current is continuous and only the fundamental component is modeled, the equivalent resistance shown in [7] is given by:

\[
R_{ac} = \frac{\pi^2 R_{dc}}{8}
\]

(2-2)
Chapter 2. Overview of IPT Power Controllers.

filter, the equivalent resistance is given by:

\[ R_{eq} = \frac{8R_0}{\pi^2} \] (2-3)

2.2 Power Flow Controllers

Figure 2.3 shows the power flow controllers used in IPT systems. Each of these are discussed in the sections below.

![Controller Block Diagram](image)

**Figure 2.3. Overview of Controller Block Diagram.**

2.2.1 Voltage Regulator Controllers

An old and conventional method to regulate DC power is to use a voltage regulator on the secondary side to maintain a constant voltage delivered to the load [25, 84]. This technique does not actually control the power transferred in the IPT system, but rather dissipates the extra power not required by the load in the voltage regulator. The basic structure of a series tuned and parallel tuned pickup with a voltage regulator and a rectifier is shown in Figure 2.4. For the series tuned pickup, a series voltage regulator is used to form a voltage divider so that the output voltage to the load can be controlled. Likewise, for the parallel tuned pickup, a shunt voltage regulator is used to form a current divider so that the output current to the load can be controlled.

A major downside of these techniques is that the voltage regulator has to dissipate a large amount of power when the load changes over a wide range. As such, this type of circuit is usually limited to low power applications. One other rare application that uses this controller...
is when the open circuit voltage and output load are precisely known and do not vary, the voltage regulator can be designed with a minimum amount of power dissipation. In most practical IPT applications, these conditions are never met.

![Figure 2.4. Voltage Regulator Pickup Topologies for (a) Series Tuned and (b) Parallel Tuned Configuration.](image)

### 2.2.2 Coupling Controllers

Coupling controllers refer to controllers that change the coupling coefficient between the primary and secondary pickup inductors. This controller generally requires some sort of mechanical movement. One obvious approach is to move the secondary inductor away from the primary to reduce the coupling, however moving the system mechanically is very tedious and may not be acceptable in many applications. Other coupling controllers use a metal shield in between the track and the pickup to screen the magnetic field linking the primary and secondary. However, this type of approach is also difficult or often impractical to implement in some applications. As such, coupling controllers have rarely if ever been adopted in IPT systems as many other elegant solutions have been found over the years.

### 2.2.3 Decoupling Controllers

Decoupling controllers are incorporated in the secondary pickup and use some sort of Switch Mode Power Supply (SMPS) to regulate the output voltage or current. This technique is far more efficient than the linear regulator approach, and thereby has higher operating efficiencies. Depending on the type of tuning topology, either a buck or boost circuit topology are commonly used. Other DC-DC converter topologies can also be used as decoupling
controllers, but are uncommon ones.

2.2.3.1 Buck Controller

This type of decoupling controller is used with a series tuned IPT pickup. Here, the series tuned pickup behaves like a voltage source and the power taken from the pickup depends on how much current is flowing to the load. The buck circuit topology is cascaded to the series IPT pickup, shown in Figure 2.5, and controls the current delivered to the load by switching at a duty cycle $d$. The behavior of this controller is exactly like a buck converter. The series IPT pickup is like a constant DC voltage source. The buck converter usually steps down a higher input voltage to a lower output voltage while increasing the input to output current ratio. In this way, the output voltage of the buck converter can be regulated to any reference lower than the input voltage. In practice the buck converter controls the duty cycle to regulate the output voltage to the desired value, and the current delivered by the pickup directly depends on the load after the buck converter. The output of the buck converter will behave similarly to an ideal voltage source.

2.2.3.2 Boost Controller

The boost topology decoupling controller is usually used with a parallel tuned IPT pickup, but it can also be used with series tuned pickups. The series pickup with boost converter topology is less common and the discussion here will be in context with the boost controller for a parallel tuned IPT pickup. The basic structure of the boost controller is shown in Figure

![Figure 2.5. Series IPT Pickup Cascaded with Buck Converter.](image-url)
2.6. There are two different types of operation for the boost controller - fast switching and slow switching. Each of the switching techniques has its own advantages and is discussed below individually.

![Parallel IPT Pickup Cascaded with Boost Converter](image)

**Figure 2.6. Parallel IPT Pickup Cascaded with Boost Converter.**

### 2.2.3.2.1 Fast Switching Overview

Fast switching refers to the switching frequency of the boost converter switch to be relatively high and is not necessarily related to the IPT frequency. The condition is meet, when the switching frequency is high enough to allow the boost converter to boost the voltage. Under fast switching conditions, the controller fully acts as a boost converter, increasing the input to output voltage ratio and decreasing the input to output current ratio. To regulate the output voltage, the power delivered by the pickup must change as the load changes. The output of this pickup can also be considered as a voltage source, with one limitation that the output current of this voltage source is limited by the short circuit current of the pickup multiplied by the duty cycle. Typical waveforms of a fast switching controller are shown in Figure 2.7.
Chapter 2. Overview of IPT Power Controllers.

2.2.3.2.2 Slow Switching Overview

Slow switching refers to the switching frequency of the boost converter switch to be relatively low. In a typical example, a switching frequency of 1-100Hz is considered slow. Under slow switching, the controller acts very differently as compared to a boost converter in the fast switching operation. Since the output of the pickup is a current source the switch closes and opens, depending on whether the load needs the current. Under the switch open condition, the output current of the pickup will charge capacitor $C_3$ and the output voltage will rise. Under the switch closed condition, the pickup will be decoupled from the IPT system as shorting a current source would dissipate no energy, while the output voltage drops as $C_3$ discharges through the load resistor. Usually some kind of hysteresis controller is required to regulate the output voltage within certain bounds. To regulate the output voltage as the load changes, the time the circuit stays in the open mode or the closed mode changes to keep the output voltage constant within hysteresis limits. Typical waveforms for a slow switching controller are shown in Figure 2.8.
2.2.3.2.3 Comparison of Slow and Fast Switching Operation

Unlike the fast switching operation, the slow switching operation completely collapses the resonant capacitor voltage when the switch is shorted. The resonant capacitor voltage builds up when the switch is opened. The resonant energy will be completely transferred to the DC inductor and dissipated in the internal resistance in the switch and rectifier diodes each time it is shorted. However the losses are not too bad as the frequency is low. For fast switching, the resonant capacitor voltage is not collapsed, however the switching frequency is much higher which can result in greater switching losses in the boost converter stage if it is not designed with care.

When the circuit is not requiring full power, fast switching operation would always keep the resonant circuit VAR’s at a minimum while keeping the output voltage constant. However, the slow switching topology would always alternate the resonant circuit between full power and decoupled conditions. From this, the steady state losses in the resonant tank would be lower for a fast switching controller as lower resonant currents would lead to a squared time less dissipation in $I^2R$ losses in the ESR of the pickup inductor and tuning capacitor, while the slow switching operation would only be linearly proportional.
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Switching noise can also be reflected back onto the mains frequency through the primary power supply from the secondary decoupling controller. The fast switching controller would be perceived as a fixed load averaged by the fast switching frequency. However, switching transients will be generated by the slow switching controller as it turns on and off. In addition, due to the hysteresis nature of the controller, the switching frequency changes as the load varies. This makes the EMI filter on the primary supply difficult to design as a wide range of low frequency components are reflected back onto the mains.

Despite what is discussed above, in practice it is hard to specifically compare which switching strategy is better. It is rather application dependent. So the designer must decide on the type of controller topology desired based on application.

Decoupling controllers have been found to be excellent in IPT systems where multiple secondary pickups are coupled onto one long track. Over the years, weaknesses in the decoupling controller have been slowly identified in applications. One weakness of the decoupling controller is that it cannot tune the system back into resonance if any of the tuning components drift from their ideal value. One method of switching small capacitors into the resonant circuit to tune the system is presented in [90]. Another method using a saturable inductor to tune an LCL pickup has been presented in [91].

2.2.4 Tuning/Detuning Controllers

Variable tuning controllers adjust the power flow in an IPT system by tuning the resonant circuit if more power is required by the load and detuning the resonant circuit if less power is required. Using the detuning aspect of resonant circuits, power transfer can be achieved in an efficient manner. These controllers have advantages over decoupling controllers as they have the ability to continuously retune the system if maximum power is required.

The tuning of the resonant circuit is varied by either adjusting the pickup inductance or its tuning capacitance. In the most practical application, designing a pickup inductor that can achieve a wide variation in inductance is difficult. Hence, the tuning capacitor is usually adjusted. If it can be assumed that the equivalent capacitance of the resonant circuit is
controlled, the resonant circuit can tune and detune itself using this variable equivalent capacitor. Although the tuning idea can be equally applied to a series tuned circuit, here a simple parallel tuned LC circuit is discussed because it is the most commonly used. The rectifier, DC filtering components, and the DC load can be considered as an AC equivalent resistor as discussed in the section 2.1. In addition, the primary track current and the operating frequency of the IPT system are assumed to be constant. By using the well known voltage divider relationship, the output voltage can be written as:

\[
|V_c| = \frac{V_{oc}}{\sqrt{(1 - \omega^2 L_2 C_c)^2 + (\omega L_2 / R_c)^2}}
\]

(2-4)

\[
\angle V_c = \arctan \left( \frac{\omega L_2 / R}{1 - \omega^2 L_2 C_c} \right)
\]

(2-5)

Let the equivalent tuning capacitance of the circuit be:

\[
C_2 = \xi C_0
\]

(2-6)

where

\[
C_{20} = \frac{1}{\omega L_2}
\]

(2-7)

\[
\xi \in \mathbb{R}
\]

To demonstrate the control of the output voltage using different \(\xi\) values, (2-6) can be substituted into (2-4) and expressed as:

\[
|V_c| = \frac{V_{oc}}{\sqrt{(1 - \xi^2)^2 + \left(\frac{1}{Q_2}\right)^2}}
\]

(2-8)

where the quality factor is defined as:

\[
Q_{20} = \omega R_2 C_0 = \frac{R_2}{\omega L_2}
\]

(2-9)
Chapter 2. Overview of IPT Power Controllers.

The variation of the output voltage against different $\xi$ values for different circuit quality factors is shown in Figure 2.9. The output voltage is decreased as the tuning capacitance value is changed from the fully tuned value of $\xi=1$. Hence, by changing the tuning capacitance of the circuit, the power transfer of the IPT system is controlled. It should also be noted from Figure 2.9 that the rate of change of output voltage increases as the $Q_2$ of the circuit gets higher. This increases the sensitivity of the circuit and more precise control systems have to be used to control the output voltage accurately.

![Figure 2.9. Output Voltage vs. Equivalent Tuning Capacitance.](image)

One main advantage of using a variable tuning controllers is that it can significantly boost the operating $Q_2$ of the circuit if the circuit is not already in tune. This is particularly useful in applications where the coupling is poor and the $Q_2$ of the IPT system is limited to 10 due to finite tolerance in the tuning capacitor value. By allowing the tuning capacitor to be controlled very precisely, the resonant circuit can be tuned much more accurately. This enables the operating $Q_2$ of the pickup to be higher than for passively tuned pickups. Consequently, a higher $Q_2$ directly leads to more power transferred in the IPT system. Similarly, for an IPT system that delivers the same amount of power, a higher $Q_2$ could potentially reduce the requirement for a good coupling configuration, a high track current and higher operating frequencies. Another advantage of varying the tuning of the system is that the resonant current in the secondary system reduces dramatically when it enters the detuned state when no power is required by the load. This reduces the standing losses in the inductor coil and improves the
efficiency of the system. However, a higher $Q_2$ pickup increases the VA of the resonant tank which lead to higher ratings for the components. In addition, this may reduce the efficiency of the pickup as more current is circulating in the resonant tank.

The reflected power factor of the pickup on the primary track is poorer than a traditional IPT decoupling controller for a parallel tuned pickup. The variable tuning pickup reflects variable VAR’s back on the primary track, which the primary power supply has to source. This may result in more stress in the primary supply as it has to source both the real and imaginary power required by the pickup.

There are several methods to realise a variable tuning capacitor to regulate power delivered to a load. In practice, these are:

2.2.4.1 Variable Tuning Technique using a Fixed Reactance Bank

Power regulation can be implemented by switching fixed reactive components in and out of a resonant circuit to tune and detuned the circuit operation. This technique is widely adopted in power systems to control power flow. However, the technique has poor performance in power electronics as sourcing the VAR’s is very expensive and impractical. These techniques were used decades ago to regulate power in electric vehicle charging applications [13]. Practically these systems were realised by storing a bank of fixed reactive components with different values, and each one is switched in or out of the system, when necessary, to change the power transferred to the load. The reactive components take the form of either capacitors or inductors. Four common circuit topologies that can realise this control strategy using the parallel tuned circuit are shown in Figure 2.10. Most of these circuit topologies are hard to implement as the AC switches in every topology require many semiconductor devices to realise. Here, only two reactive components are drawn for the reactance bank for simplicity. The most common type is to use a capacitor bank as they are usually cheaper and easier to design.
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Figure 2.10. Fixed Reactance Bank Topologies showing (a) Parallel Capacitor Bank, (b) Parallel Inductor Bank, (c) Series Capacitor Bank and (d) Series Inductor Bank.

Figure 2.10(a) shows a variable tuning system where extra tuning capacitance can be switched into the resonant tank. Referring to Figure 2.9, the operating range is on the right hand side of $\xi=1$ as the capacitance can only get larger. The equivalent tuning capacitance for this system would be:

$$C_e = C_2 + \Delta C$$  \hspace{1cm} (2-10)

Figure 2.10(b) shows a variable tuning system with inductors in parallel. Any extra inductance switched into the resonant network can be considered to reduce the effective $C_2$ so that the operating range is on the left hand side of $\xi=1$ in Figure 2.9 as the equivalent tuning capacitor will be perceived smaller as a result of the cancellation in reactance. The equivalent capacitance would be given by:

$$C_e = \frac{1}{\omega} \left( \frac{1}{\omega C_2} - \omega \Delta L \right)$$  \hspace{1cm} (2-11)
Similarly, for the series capacitor bank circuit as shown in Figure 2.10(c) and Figure 2.10(d), the equivalent pickup inductance is reduced and increased, respectively. Similarly, the output voltage is also controlled.

### 2.2.5 Variable Tuning Inductor using Magnetic Amplifier Circuit

A magnetic amplifier is basically a saturable inductor whose inductance value can change as the DC bias current on the secondary windings is controlled. This method is usually preferred over making a bank of inductors as the cost of making this magnetic amplifier is usually cheaper. A typical saturable inductor is shown in Figure 2.11.

![Saturable Inductor Diagram](image)

**Figure 2.11. A Saturable inductor.**

The DC bias current in the secondary winding directly controls the inductance of the two external windings. This is because the DC bias current generates a constant field in the core and indirectly shifts its operating point on the B-H curve. As an example, the inductance versus the DC bias current is shown in Figure 2.12. It can be seen as the DC bias current increases, the inductor value changes over a wide range. Using this saturable inductor in the circuit topologies in Figure 2.10(b) and Figure 2.10(d) will also result in a tuning/detuning controller.
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A composite LCL topology has been previously reported using a saturable inductor to tune and detune the pickup to control power output [92]. This specific LCL topology, has the advantage of producing an ideal current source under perfectly tuned conditions, which is desirable for many types of loads.

One disadvantage of using a saturable inductor is the slow transient response of the overall pickup when a step change in load has occurred. This is because the saturable inductor has a dominant first order pole due to the filtering of the secondary winding for the DC bias current. As such, the typical transient response times reported in [92] was thousands of resonant cycles. One other critical disadvantage is the cost of making this saturable inductor for high power IPT applications. To vary the AC power directly, the use of a variable inductor in the non-linear region of the B-H curve can limit the efficiency of the overall system. In addition, the variable inductor is expensive to manufacture because it has to manage the high resonant current without fully saturating and also take into account sensitivity issues in the system.

### 2.2.5.1 Variable Tuning Technique using a Switch Controlled Reactance

One possible technique to achieve a variable reactance component is to switch one fixed reactance component with an AC switch [25]. This idea is widely used in power systems and forms the fundamental theory for Flexible Alternating Current Transmission Systems (FACTS) [93]. The AC switch operates with a relative phase delay against a reference voltage.
or current to control the equivalent reactance. Soft switching conditions such as ZVS and ZCS are both achieved under normal circumstances depending on whether an inductor or a capacitor is used as the reactance component. There are many possible circuit topologies that exist for this technique and the most common forms are shown in Figure 2.13. Figure 2.13(a) and Figure 2.13(b) shows a capacitor used as the reactance in the switching network. For the topology using an AC switch in series with the capacitor, the equivalent capacitance value is controlled from 0 to $C$ as the switch is controlled. Likewise, for the topology using an AC switch in parallel, the equivalent capacitance value is controlled from $C$ to $\infty$. Similarly, Figure 2.13(c) and Figure 2.13(d) shows an inductor used as the reactance in the switching network. The series switch topology would have the inductance value vary from $\infty$ to $L$. Likewise, the parallel switch topology would have the inductance value vary from $L$ to 0. The composite tuned topologies shown in Figure 2.13, is usually used to reduce the voltage and current ratings of the AC switch. However, this is achieved at the cost of a narrower control range. Such networks are usually used in power systems to reduce the semiconductor device cost for 220kV-800kV transmission systems.

![Figure 2.13. Switched Reactance Networks showing (a) Capacitor with Series Switch, (b) Capacitor with Parallel Switch, (c) Inductor with Series Switch, (d) Inductor with Parallel Switch and (e) Composite Reactance with Switch.](image)

The derivation of equivalent reactance for each of the topologies would require considerable mathematical detail, and in previous literature the equivalent capacitance value for Figure 2.13(a) has been derived in [25] and the equivalent inductance value for Figure 2.13(c) has been derived in [94]. As such, the only detailed derivation shown is for the
2.2.5.1.1  **Equivalent Capacitance of a Series Switched Capacitor**

One possible method to control the series switched capacitor illustrated in [25] has the waveforms shown in Figure 2.14. From top to bottom, the waveforms are, the resonant capacitor voltage, the series capacitor voltage, and the two gate control waveforms for driving the equivalent AC switch (This AC switch is formed by two back to back MOSFET’s). The duty cycle range for both the gate control waveforms ranges from 0% to 100% as the capacitance value is varied.

Since the electric charge flowing through capacitor $C$ during a period must be equal to that of using an equivalent capacitor $C_{eq}$, the equivalent capacitance can be shown to be:

---

**Figure 2.14. Voltage and Control Signal Waveforms for Switched Series Capacitor.**
\[ C_{eq} = \frac{2C[1-\cos(\theta)] + C \sin(\theta)(\pi - 2\theta)}{2} \] (2-12)

By plotting (2-12), it can be seen that the equivalent capacitance changes from 0 to \( C \) as the switching angle changes from \( \theta = 0 \) to \( \theta = \pi / 2 \).

![Figure 2.15. Equivalent Capacitor \( C_{eq} \) versus Switching Angle \( \theta \).](image)

The control of the equivalent reactance component can be done very quickly over one resonant cycle as the duty cycle of the gate waveforms can be adjusted in the next cycle. This gives these variable reactance networks very fast transient response times. Since the AC switch is soft switched, the switching losses are also negligible. The key design procedure for this system is rating the switches for their peak voltage and current stresses.

### 2.2.6 Primary Side Controllers

Primary side controllers have been one of the oldest forms of power flow control back dating all the way to the early 1970’s [45, 58, 95]. For primary control to be realized, feedback signals such as the output voltage of the secondary pickup must be fed back to the primary converter via a wireless communication channel. Generally, primary side control has two possible methods of realization – primary current control [67] or operating frequency control [57], and each of these are discussed below.
2.2.6.1 Primary Track Current Controllers

This type of controller relies on the fundamental principle that the power delivered to the secondary pickup is dependent on the magnetic field or the primary track current. By controlling the primary current, the power to the secondary pickup can be controlled. A simple example of how it is related was shown in chapter 1 when the power delivered to the pickup was proportional to the square of the primary track current. There are many different ways to control track current, depending on the topology of the primary converter used. For the conventional current fed parallel tuned push pull converter, the simplest way is to control the input voltage to the converter to control the track current [96]. Similarly, it is also favourable to use the input voltage to control the track current in a Class-E converter [28]. Input voltage control is usually implemented using an extra buck converter and adjusting its duty cycle. For a voltage fed series tuned converter, the more common option is to use phase control [97, 98]. For LCL converters, it is more suitable to use duty cycle control due to the nature of the resonant tuning network [54]. Although these two later primary current regulation methods do not require an additional DC-DC converter stage, they will lose their ZVS condition once the primary track current is regulated if they were originally designed for ZVS operation. This would result in more switching losses, EMI and switching stress, but saves the cost of an extra DC-DC conversion stage.

Primary track current control is an electrically efficient method of controlling power. When little or no power is required by the load, the controller can reduce almost all the current flowing in the IPT system which includes the primary converter and the secondary pickup. The reduction in current reduces the $I^2R$ losses as less current is flowing in the overall system. These results in a higher overall operating efficiency compared to most other controllers, especially when the load requires a small amount of power.

2.2.6.2 Operating Frequency Controllers

One other type of primary side controller is an operating frequency controller which changes the overall operating frequency of the whole IPT system to control power flow [57].
Chapter 2. Overview of IPT Power Controllers.

By adjusting the resonant frequency on the primary converter, the secondary resonant tank will vary in tuning. This type of control method is very similar to the previously mentioned secondary tuning/detuning controllers, but here the variations in reactance are on the primary side. Similarly, the methods used to obtain a variable reactance to adjust the resonant frequency for a push pull converter uses the exact circuit topology outlined in Figure 2.13(a) [57]. Many of the concepts that were deduced in the prior section can be directly applied to this controller as changing the operating frequency can be linked directly to a variation in tuning components when the frequency is constant.

Although primary side control has many advantages, they suffer from their relatively slower transient response as the output voltage cannot be regulated in sub-millisecond time intervals due to the limited bandwidth of resonant circuits and delays in the wireless communication channel necessary for control purposes. In addition, due to the large amount of other wireless communication devices used, the interference generated in many circumstances may lead to loss of data packets which causes output voltage variations that are not acceptable. Moreover, since the wireless communication system is operating very close to the IPT system, any switching noise from the high power electronics devices can easily generate enough interference. As a result, a recent publication has used a clever approach of using an observer to predict and construct the output voltage of the secondary pickup by measuring the reflected impedance on the primary track [99]. However, this technique requires the coupling to be fixed as it is directly used in the observer. In addition, this technique can only be used in applications where there is one secondary pickup. As a result, this estimation technique can only be used in single secondary pickups that have fixed coupling.

2.2.7 Limitations with Existing IPT Controllers

The voltage regulator controller is a simple method. Although this controller can operate over a wide load range, its poor operating efficiency makes it obsolete in most applications these days. Similarly, coupling controllers have also become obsolete as changing the coupling or shielding the magnetic field linkage is not mechanically acceptable in many
Chapter 2. Overview of IPT Power Controllers.

The decoupling controller has been shown to be simple, robust, and widely used in high power industrial applications with multiple pickups. These advantages make it an excellent controller to be used in materials handling systems or other industrial applications [10, 11, 22, 90, 100-103]. However, when this controller is used in applications with large variations in coupling such as EV charging or biomedical implants, an inefficient method of overrating circuit devices is required. Moreover, since the shorting current of the parallel resonant tank is constantly flowing through the switch, this constant current flow causes on-state losses in the switch which lowers efficiency. This is particularly evident in systems whose coupling coefficient can change by over 100%. To design such a system, the minimum coupling coefficient must be used to ensure that the minimum current is enough to satisfy the power transfer required. However, under maximum coupling conditions the short circuit current is 100% larger than required which means all the components have to be rated 200% in current which reduces efficiency and increases cost.

The tuning/detuning controller using a saturable inductor in the AC compensation path is required to realize a variable reactance component [92]. Since this topology can directly control the resonant tuning on the secondary, it has the advantage of being able to operate with variable coupling and tuning conditions. Moreover, since it can decouple the power before the AC-DC rectifier stage, none of the components have to be overrated for large changes in coupling. However such use of a variable inductor in the non-linear region of the B-H curve limits the efficiency of the overall system. In addition, the variable inductor is expensive to manufacture because it has to manage the high resonant current without fully saturating. Moreover, the transient response of this circuit is rather poor as the saturable inductor provides significant filtering and the fastest transient response reported in [92] takes seconds.

Primary current control can handle a wide load range, a varying coupling condition and obtain excellent efficiency due to its ability to wind down the system. Operating frequency controllers which change the operating frequency of the primary converter, can operate well
with changes in tuning components on both the primary and secondary as finely adjusting the operating frequency can bring the system back into resonance condition. However, both primary side control techniques suffer from their relatively slower transient responses as the output voltage cannot be regulated in sub-millisecond time intervals due to wireless communication channel having delays. Moreover, due to the large amount of other wireless communication devices used in applications, the interference generated in many circumstances may lead to loss of data packets which lead to output voltage variations that are not acceptable.

With all the above literature review mentioned, this thesis proposes a new AC processing pickup controller for controlling the power flow in an IPT system on the secondary side. This controller has been found to have advantages over existing controllers and is a new way to control power flow. This pickup can be useful in a wide range of other applications including Electric Vehicle (EV) inductive charging, wireless powering of biomedical implants and industrial applications with multiple secondary pickups on one long primary track.

### 2.3 Summary

A general overview on IPT controllers has been undertaken in this chapter. The purpose of power flow controllers in an IPT system is to compensate for parameters changes in the system. Two of the most commonly varied parameters include load and coupling variations. As a result of the requirement to compensate for the change in these parameters, many controllers including voltage regulators, decoupling controllers, tuning/detuning controllers, coupling controllers and primary side controllers has been presented in the past. Each of these controllers has its individual advantages and limitations. The motivations of this thesis are to propose a novel IPT controller solving some of the aforementioned issues.
# The AC Processing Pickup

This chapter introduces a new AC processing pickup capable of producing an AC controllable current source with a wide variety of applications including lighting, EV charging and many other industrial applications.

## 3.1 Fundamentals of the AC Processing Pickup

A schematic circuit for an AC processing pickup is shown in Figure 3.1. Capacitor $C_2$ is tuned to pickup inductor $L_2$ at the frequency of a primary track current $I_1$ to form a resonant tank. The diodes ($D_1$, $D_2$) and switches ($S_1$, $S_2$) form an AC switch. From standard IPT theory, a pickup coil placed on the primary track would induce an open circuit voltage $V_{oc}$ source given by:

$$V_{oc} = j\omega MI_1$$

where $\omega$ is the operating angular frequency, $M$ is the mutual inductance and $I_1$ is the primary track current.

![Figure 3.1. The AC Processing Pickup.](image)

To illustrate the circuit operation, Figure 3.2 shows the waveforms when the switches are used to clamp parts of the resonant capacitor voltage. $V_{g1}$ and $V_{g2}$ are the PWM gate signals
used to drive switches $S_1$ and $S_2$ at 50% duty cycle with the same switching frequency as the IPT track frequency. Consider the situation where waveforms $V_{g1}$ and $V_{g2}$ are controlled with a phase delay $\phi$ relative to the phase of $V_{oc}$ as shown in Figure 3.2. At time $t=0$, $S_1$ is turned off and $S_2$ is turned on. However, the series diode $D_2$ blocks any current flowing through $S_2$ as it is reverse biased. This causes the capacitor $C_2$ to resonate with the pickup inductance $L_2$ like a parallel resonant tank. The capacitor voltage rings to a peak value and returns back to zero. When the capacitor voltage reaches zero, $D_2$ terminates the resonant cycle and prevents the capacitor voltage building up in the negative direction as it begins to conduct at zero volts, hence clamping the output voltage to zero. This causes $S_2$ to clamp $V_c$ for a time known as the clamp time ($t_c$) at the point where $V_c$ changes from a positive to a negative voltage. In summary, the clamping action from the AC switch generates a phase shift between the open circuit voltage and the capacitor voltage waveform. A single cycle operation for the AC processing pickup is composed of a sequence of linear circuit stages with each corresponding to a particular switching interval as illustrated to Figure 3.3. These can be grouped into the following modes:
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Figure 3.2. Typical Single Cycle Operating Waveforms when the Switches Clamp the Resonant Capacitor Voltage.
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Figure 3.3. Operating Modes of the AC Processing Pickup.

1. Mode 1 (M1): At t=0, S1 is turned off and S2 is turned on. The capacitor voltage \( V_c \) will resonate with the pickup inductor to a positive peak voltage and then decay. Since \( D_2 \) is reversed biased due to the positive \( V_c \), it blocks any current that flows through \( S_2 \). Under this mode, the circuit operates like a parallel resonant tank and current flows into the load resistor.

2. Mode 2 (M2): At \( t=t_1 \), the capacitor voltage naturally crosses zero. Since \( S_2 \) is still turned on, \( D_2 \) becomes forward biased, the resonant cycle is terminated and the capacitor voltage is clamped. In this mode, the inductor current flows through the switch \( S_2 \) and no current flows through the load.

3. Mode 3 (M3): At \( t=t_2 \), S1 is turned on and S2 is turned off. Similar to M1, the circuit operates like a parallel resonant tank and current flows into the load resistor.

4. Mode 4 (M4): At \( t=t_3 \), the capacitor voltage naturally crosses zero. Similar to M3, the resonant cycle is terminated and the capacitor voltage is clamped. In this mode, the inductor current flows through the switch S1 and no current flows through the load. After this mode, the circuit returns back to M1, and the switching process is repeated.

The AC processing pickup achieves approximate soft switching conditions. From Figure
3.2, the resonant inductor current starts to flow through $S_2$ at $t_1$ when there is no voltage across it, hence ZVS is achieved at turn on. When $S_2$ is turned off at $t_2$, the resonant capacitor in parallel with $S_2$ forces the voltage across $S_2$ to increase slowly in the negative direction while the current through it decreases to zero. For most practical switches, the turn off is much faster than the rate of increase of the capacitor voltage, so the $dv/dt$ across the switch is relatively small and ZVS is obtained at the switch off condition. Switch $S_1$ operates in a similar manner and also achieves ZVS at turn on, while achieving a low $dv/dt$ at turn off. Likewise, for diodes $D_1$ and $D_2$, low $dv/dt$ is achieved at turn on, and ZVS is achieved at turn off. In summary, the switches and diodes in the AC processing pickup achieve soft switching. This gives the pickup desirable characteristics such as low switching losses, low switching stress and reduced electromagnetic interference (EMI) levels. The low switching losses gives this pickup controller a very high operating efficiency. Moreover, the low EMI provides little interference on the control circuitry of the pickup and external systems nearby.

### 3.2 Simplified Pickup Analysis

From the previous section, it can be seen that the phase shift between $V_c$ and $V_{oc}$ can be controlled by adjusting the phase delay $\phi$. This concept will be the fundamental basis of controlling power flow in the AC processing pickup. An analysis procedure is presented in this section and it is greatly simplified based on the three following assumptions:

1. The Equivalent Series Resistance (ESR) of both capacitor $C_2$ and Inductor $L_2$ are very small and can be neglected. (This is because the resistive losses dissipated by the load are always much larger)

2. The switching action of the transistors and diodes are instantaneous and lossless.

3. Only the fundamental component of the capacitor voltage and inductor current are considered in the analysis. This is because most secondary pickups operate with high $Q_2$ (typical values range between 3 to 10) in industrial applications.
3.2.1 Steady State Pickup Analysis

Let the voltage across $C_2$ be replaced by a voltage source ($V_{c1}$) with the amplitude and phase of the fundamental component of $V_c$ as shown in Figure 3.4.

![Figure 3.4. Waveform showing Fundamental Component of Capacitor Voltage.](image)

Then using standard power system analysis [104], the power transferred can be approximated by:

$$P = \frac{V_{oc} V_{c1}}{\omega L_2} \cos(\theta_v)$$  \hspace{1cm} (3-2)

where $\theta_v$ is the angle between $V_{oc}$ and $V_{c1}$. This equation describes the amount of power transferred between two active voltage sources with given amplitude and a known phase difference. The power transferred is dependent on the voltage amplitude and the phase between the two voltage sources. For zero phase delay $\phi$, the phase $\theta_v$ between $V_{oc}$ and $V_{c1}$ is $0^\circ$ and maximum power is transferred to the load. This is the equivalent to a perfectly tuned parallel resonant tank with the AC switch not present. For a finite phase delay $\phi$, the phase $\theta_v$ between $V_{oc}$ and $V_{c1}$ is greater than $0^\circ$. Under this condition, less power is transferred - given by (3-2). Consequently, by adjusting this phase delay, the power delivered by the pickup can be controlled. By assuming there are negligible losses in the pickup, an input to output power...
balance equation can be written as

\[
\frac{V_{oc}V_{cl}}{\omega L_2} \cos(\theta_v) = \frac{V_{cl}^2}{R}
\] (3-3)

By simplifying (3-3)

\[|V_{cl}| = Q_{20} V_{oc} \cos(\theta_v) \] (3-4)

\[\angle(V_{cl}) = \theta_v \] (3-5)

where

\[Q_{20} = \frac{R}{(\omega L_2)} \] (3-6)

\(Q_{20}\) is the ideal quality factor of the parallel tuned circuit which is only dependent on circuit parameters. It should not be confused with the actual operating \(Q\) of the circuit mentioned in previous literature. The capacitor voltage given in (3-4) is also the output voltage of the pickup. By noting that the inductor current is lagging the inductor voltage by 90° and the inductor voltage is the difference between \(V_{oc}\) and \(V_{cl}\), the inductor current can be determined as

\[|I_{L1}| = \frac{Q_{20} V_{oc} \cos^2(\theta_v)}{\omega L_2 \cos(\theta_v)} \] (3-7)

\[\angle(I_{L1}) = \theta_i \] (3-8)

where

\[\theta_i = \arctan\left(\frac{1}{Q_{20} \cos^2(\theta_v) + \tan(\theta_v)}\right) \] (3-9)

The output current delivered to the load may be calculated using (3-3) and is

\[I_R = \frac{V_{oc} \cos(\theta_v)}{\omega L_2} \] (3-10)
Similarly, the output power can be written as

\[ P_{\text{out}} = \frac{Q_{20} V_{\text{oc}}^2 \cos^2(\theta_1)}{\omega L_2} \]  

(3-11)

### 3.2.2 Steady State Power Factor Analysis

A major drawback to the AC processing pickup is that the power factor reflected back onto the primary track is non-unity and varies greatly with changing phase delay \( \phi \). An equivalent tuning capacitor value is derived in this section to quantify the reflected power factor. To begin the analysis, both the tuning capacitor \( C_2 \) and the AC switch (\( S_1, S_2, D_1 \) and \( D_2 \)) are replaced with an equivalent tuning capacitor \( C_e \), forming a parallel tuned resonant tank. It has been shown in [4] that the normalized output voltage for a parallel tuned pickup with an equivalent capacitance \( C_e \) is given by

\[
\frac{|V_{\text{cl}}|}{V_{\text{oc}}} = \frac{1}{\sqrt{(1 - \omega^2 L_2 C_e)^2 + \left(\frac{\omega L_2}{R}\right)^2}}
\]

(3-12)

\[
\angle \left(\frac{V_{\text{cl}}}{V_{\text{oc}}}\right) = \arctan \left(\frac{\omega L_2}{1 - \omega^2 L_2 C_e} \right)
\]

(3-13)

By either equating the magnitude or the phase component with (3-4) and (3-5), the equivalent capacitance is

\[
C_e = \frac{1 \pm \left(\tan(\theta_1) / Q_{20}\right)}{\omega^2 L_2}
\]

(3-14)

The reflected power factor on the primary track of the IPT system for a secondary pickup [12] is:

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

(3-15)

where \( Z_2 \) is the impedance seen by the open circuit voltage source by parallel tuned circuit.

For this parallel tuned pickup, \( Z_2 \) is expressed as
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\[ Z_2 = \frac{R}{1 + \omega^2 C_e^2 R^2} + j \left( \omega L_2 - \frac{\omega C_e R^2}{1 + \omega^2 C_e^2 R^2} \right) \]  \hspace{1cm} (3-16)

Substituting (3-15) into (3-16), the reflected impedance is determined as

\[ R_r = \frac{\omega^2 M^2 / R}{(1 - \omega^2 L_2 C_e)^2 + (\omega L_2 / R)^2} \]  \hspace{1cm} (3-17)

\[ X_r = \frac{\omega^2 M^2 \left( \omega C_e \left( 1 - \omega^2 L_2 C_e \right) - \omega L_2 / R^2 \right)}{(1 - \omega^2 L_2 C_e)^2 + (\omega L_2 / R)^2} \]  \hspace{1cm} (3-18)

The reflected impedance can be evaluated using (3-17) and (3-18) by substituting for the equivalent capacitance calculated in (3-14). Both the real and imaginary part of the reflected impedance have to be driven by the primary power supply and an increase in the reactive part may mean that the power supply has to be overrated to source the extra Volt-Amp’s (VA) in the system.

3.2.3 Transient Analysis

The simple power balance analysis used in 3.2.1 can be extended to analyzing the transient operation of the AC processing pickup when the controlled phase delay suddenly changes. Here, the input to output power balance equation is used as before, but in addition, the instantaneous energy stored in the resonant capacitor is also considered. Since a step in the phase delay \( \phi \) will change the input power of the pickup given by (3-2), the output power and instantaneous stored energy in the capacitor will also vary. Note that this analysis only takes into account the envelope of the waveform as time progresses rather than the instantaneous resonant oscillations. This gives a more representative meaning for resonant systems that exhibit predominantly oscillatory behavior [105, 106]. The change in capacitor energy from one period to the next is given by:

\[ E = C_2 \left( \left( \tilde{V}_{c1} \right)_f^2 - \left( \tilde{V}_{c1} \right)_i^2 \right) / 2 \]  \hspace{1cm} (3-19)
where \((\hat{V}_{cl})_f\) is the final peak capacitor voltage and \((\hat{V}_{cl})_i\) is the initial peak capacitor voltage. Considering the input energy, output energy and stored energy in the system, the following power balance equation can be written:

\[
P_{in} = P_{stored} + P_{out} \tag{3-20}
\]

Expanding (3-20), the following equation can be derived

\[
\frac{(\hat{V}_{cl})_f}{\sqrt{2}} \cdot \frac{(\hat{V}_{cl})_i}{\sqrt{2}} \cos(\theta_v) \Delta t = \frac{2}{R} \cdot \left(\frac{(\hat{V}_{cl})_f}{\sqrt{2}}\right)^2 - \left(\frac{(\hat{V}_{cl})_i}{\sqrt{2}}\right)^2 - \frac{C_2}{\Delta t} \cdot \frac{2}{2}\left(\hat{V}_{cl}\right)^2 - \frac{\left(\hat{V}_{cl}\right)^2}{2} \tag{3-21}
\]

For small transient shifts in the phase delay \(\theta_v\), \((\hat{V}_{cl})_f - (\hat{V}_{cl})_i = \Delta \hat{V}_{cl}\) and \((\hat{V}_{cl})_f + (\hat{V}_{cl})_i \approx 2\hat{V}_{cl}\), simplifying (3-21) will give

\[
\frac{\hat{V}_{oc} \cdot \hat{V}_{cl} \cos(\theta_v)}{2\omega L_2} = \frac{\left(\hat{V}_{cl}\right)^2}{2R} + \frac{C_2 \cdot \hat{V}_{cl} \cdot \Delta \hat{V}_{cl}}{\Delta t} \tag{3-22}
\]

Expressing (3-22) as a differential equation results in

\[
\frac{\partial \hat{V}_{cl}}{\partial t} + \frac{1}{2RC_2} \cdot \hat{V}_{cl} = \hat{V}_{oc} \cdot \cos(\theta_v) \tag{3-23}
\]

The solution to (3-23) is

\[
\hat{V}_{cl}(t) = \left(\left(\hat{V}_{cl}\right)_i - (\hat{V}_{cl})_f\right) \exp^{\left(\frac{-at}{2RC_2}\right)} + (\hat{V}_{cl})_f \tag{3-24}
\]

where \((\hat{V}_{cl})_f\) and \((\hat{V}_{cl})_i\) can be calculated using (3-4) by substituting the initial \(\theta_v\) and final \(\theta_v\) values. The capacitor voltage shown in (3-24) only gives the instantaneous envelope of the capacitor voltage as a function of time.

All the preceding analysis assumes the fundamental phase difference between \(V_{oc}\) and \(V_{cl}\) is known. But unfortunately, an analytical relationship between the controlled phase delay \(\phi\) and \(\theta_v\) is extremely difficult to obtain. The reason for such complexity in the analysis is that
although the controlled phase delay $\phi$ is known, the time at which the diode starts to conduct when the circuit resonates back to zero cannot be analytically determined. Although transcendental equations can be formed to solve for the solution, the equations can only be solved by numeric solvers like MATLAB and EXCEL. As such, a normalized graph computed in MATLAB is shown in Figure 3.5, to aid the design and analysis of the pickup. The details of how this graph is produced are outlined in section 3.3.

**Figure 3.5. Fundamental Phase $\theta_v$ vs. Controlled Phase Delay $\phi$.**

### 3.3 Exact Analysis of AC Processing Pickup

It has been shown in the previous section that (3-2) can describe how power is controlled in the proposed AC processing pickup. Although the equation can provide some insight to the circuit operation, it is of little help over most operating conditions as the phase angle $\theta_v$ cannot be easily determined. In this section, an exact analysis in the time domain is proposed to determine the characteristics of the circuit under steady state operation. The basis of the analysis method is that the conditions existing in the circuit at the end of a particular switching period must be the initial conditions for the start of the next switching period, and
these conditions must be identical allowing for changes in polarity caused by the resonant operation. This method includes the harmonic components ignored in the previous analysis technique.

The analysis procedure is simplified using the same assumptions as in section 3.2:

1. The Equivalent Series Resistance (ESR) of both capacitor $C_2$ and Inductor $L_2$ are very small and can be neglected.

2. The switching action of the transistors and diodes are instantaneous and lossless.

### 3.3.1 Basic Equations

Let the tuning capacitance of the circuit be,

$$C_2 = aC_{20} \quad (3-25)$$

where

$$C_{20} = 1/(\omega^2 L_2) \quad (3-26)$$

$$a \in \mathbb{R}$$

With reference to Figure 3.6, the waveform can be separated into two operating modes known as the resonant mode and the clamp mode.

![Figure 3.6. Waveform in Two Modes.](image)
3.3.1.1 Resonant Mode

During the resonant mode, the capacitor voltage may be described as:

\[
\frac{d^2 V_c}{dt^2} + \frac{1}{RC_2} \frac{dV_c}{dt} + \frac{V_c}{L_2C_2} = \frac{V_{oc}}{L_1C_2} \sin(\omega t + \phi) \tag{3-27}
\]

Considering the initial condition \( V_c(t) \big|_{t=0} = 0 \) and \( \frac{dV_c}{dt} \big|_{t=0} = -\frac{i_e(0)}{C_2} \), the complete solution of the above equation is:

\[
V_c(t) = V_{cd}(t) + V_{cu}(t) \tag{3-28}
\]

where

\[
V_{cd}(t) = \frac{\partial_2}{\sin(\gamma_v)} e^{-\sigma t} \sin(\omega_f t - \theta_v) \tag{3-29}
\]

\[
V_{cu}(t) = (1-a)\partial_1 \sin(\omega_f t + \phi) - \frac{\partial_1}{Q_{20}} \cos(\omega_f t + \phi) \tag{3-30}
\]

\[
Q_{20} = \frac{R_2}{(\omega L_2)} \tag{3-31}
\]

\[
\sigma = \frac{1}{2aR_2C_{20}} \tag{3-32}
\]

\[
\omega_f = \omega \sqrt{\frac{1}{a} - \frac{1}{a} / (4a^2Q_{20}^2)} \tag{3-33}
\]

\[
\partial_1 = \frac{V_{oc}}{((1-a)^2 + (1/Q_{20})^2)} \tag{3-34}
\]

\[
\partial_2 = \partial_1 ((1-a) \sin(\phi) - \cos(\phi) / Q_{20}) \tag{3-35}
\]

\[
\partial_3 = \partial_1 ((1-a) \cos(\phi) + \sin(\phi) / Q_{20}) \tag{3-36}
\]

\[
\gamma_v = \tan^{-1} \left( \frac{-\omega_f \partial_2}{i_L(0)/\alpha C_{20} + \sigma \partial_2 + \omega \partial_3} \right) \tag{3-37}
\]
In a similar way, considering the initial condition \( i_L(t)|_{t=0} = i_L(0) \) and
\[
\frac{di_L}{dt} \bigg|_{t=0} = -\frac{V_{oc} \sin \phi}{L_2},
\]
the complete solution to the inductor current is:
\[
i_L(t) = i_{ld}(t) + i_{lw}(t)
\]
(3-38)

where
\[
i_{ld}(t) = -\frac{i_L(0)}{\sin(\gamma)} - \frac{\beta_2}{e^{-\alpha t}} \sin(\omega_L t - \theta_L)
\]
(3-39)
\[
i_{lw}(t) = -\frac{\partial_1}{R_2} \sin(\omega t + \phi) + \frac{\partial_1 Q_{20}}{R_2} (-a(1-a) + \frac{1}{Q_{20}^2}) \cos(\omega t + \phi)
\]
(3-40)
\[
\beta_2 = \frac{\partial_1}{R_2} \left( \sin(\phi) - Q_{20} \cos(\phi)(-a(1-a) + \frac{1}{Q_{20}^2}) \right)
\]
(3-41)
\[
\gamma = \tan^{-1} \left( \frac{-\omega_L (i_L(0) + \beta_L)}{-V_{oc} \sin(\phi) / L_2 + \omega \beta_2 + \sigma (i_L(0) + \beta_2)} \right)
\]
(3-42)
\[
\beta_1 = \frac{\partial_1}{R_2} \left( \cos(\phi) + Q_{20} \sin(\phi)(-a(1-a) + \frac{1}{Q_{20}^2}) \right)
\]
(3-43)

To investigate how long the circuit stays in the resonant mode, \( V_c=0 \) can be substituted in (7), resulting in the following expression:
\[
V_c(t_r) = 0
\]
(3-44)

where \( t_r \) is the time the circuit operates in the resonant mode.

3.3.1.2 Clamp mode

During the clamp mode, the inductor \( L_2 \) is shorted and the current depends on \( V_{oc} \). By Kirchhoff’s Voltage Law, the inductor current equation can now be written as:
\[
i_L(t) = -\frac{V_{oc}}{L} \int_{t_c}^{t} \sin(\omega t + \phi) dt + i_L(t_c)
\]
(3-45)
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Solving (3-45), the inductor current can be expressed as:

\[ i_L(t) = \left( V_{oc} \cos(\omega t + \phi) \right) / (\omega L_2) + i_L^- \]

(3-46)

where

\[ i_L^- = i_L(t_r) - \left( V_{oc} \cos(\omega t_r + \phi) \right) / (\omega L_2) \]

(3-47)

Because the resonant mode and the clamp mode are repeated each half cycle (with only a polarity change), the relationship \( i_L(0) = -i_L(T/2) \) must hold. Hence, the capacitor voltage and inductor current must be given by,

\[
V_c(t) = \begin{cases} 
V_{cd}(t) + V_{cu}(t) & t_0 \leq t < t_1 \\
0 & t_1 \leq t < t_2 \\
-V_{cd}(t) - V_{cu}(t) & t_2 \leq t < t_3 \\
0 & t_3 \leq t < t_4 
\end{cases}
\]

(3-48)

\[
i_L(t) = \begin{cases} 
i_{cd}(t) + i_{cu}(t) & t_0 \leq t < t_1 \\
\left( V_{oc} \cos(\omega t + \phi) \right) / (\omega L_2) + i_L^- & t_1 \leq t < t_2 \\
-i_{cd}(t) - i_{cu}(t) & t_2 \leq t < t_3 \\
-\left( V_{oc} \cos(\omega t + \phi) \right) / (\omega L_2) - i_L^- & t_3 \leq t < t_4 
\end{cases}
\]

(3-49)

3.3.2 Computation Routine

The above analysis is very difficult to solve analytically as the solution of \( t_r \) and \( i_L(0) \) are governed by (3-44) and (3-46) with \( \gamma_c \) and \( \gamma_i \) as interim variables which are associated with the auxiliary equations (3-37) and (3-43). This is in the form of transcendental equations and it can only be solved using numeric solvers such as MATLAB or EXCEL. A computer program based on an iterative computation, shown in Figure 3.7, has been developed to undertake the analysis. The program starts by initializing the circuit parameters such as \( L_2, C_2 \) and \( R_2 \). The initial condition of the inductor current is first set to the solution given by (3-7) and (3-8) in the routine as an educated guess. With \( i_L(0) \) known, \( t_r \) can be calculated solving (3-44) and the inductor current when the resonant mode ends can be calculated using (3-38). With \( t_r \) and known, the inductor current after half a period can be calculated using (3-46). The next step is to check whether \( i_L(0) \) and \(-i_L(T/2)\) have converged to within a given error
If the answer is YES, the program terminates and the correct solution is deemed to be found. Otherwise the iteration repeats itself in the computation loop until a solution is found. The algorithm has proven to be both fast and robust.

Figure 3.7. A Flow Chart of Computation Algorithm.

### 3.3.3 Pickup Characteristics

Using the exact analysis method proposed in section 3.3, normalized pickup characteristics are computed in this section.

The output voltage (or capacitor voltage) characteristics of the pickup are shown in Figure 3.8 for different values of $Q_{20}$ or load condition. The normalized output voltage is defined by the ratio of the output voltage over the open circuit voltage. It can be seen that the output power asymptotically decreases as the controlled phase delay $\phi$ increases from zero. It should be noted that the output voltage can be controlled from the maximum permissible value of the parallel resonant tank ($Q_{20}V_{oc}$) down to zero. The response shown by the normalized graph closely matches the response given by (3-4) once the transformation is made from $\phi$ to $\theta_v$. 

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However, the graphs are necessarily different as harmonics were ignored in the simplified analysis.

The normalized output current is shown in Figure 3.9 for a range of $Q_{20}$ values. It can be seen that the output current of the pickup can be controlled by phase delay as given in (3-10). Figure 3.9 shows that the output current stays approximately constant as the load resistance changes for pickups at high $Q_{20}$ (5-10). This also matches the analytical analysis given in (3-10) where the output current is independent of the load. Hence, this pickup demonstrates controllable current source behaviour. For low $Q_{20}$ values, the simplified expression is no longer true as the harmonics increase. In addition, $\theta_v$ will also be marginally dependent on $Q_{20}$ if the $Q_{20}$ of the circuit is low.

![Figure 3.8. Normalized RMS Output Voltage vs. Controlled Phase Delay $\phi$.](image-url)
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Figure 3.9. Normalized RMS Output Current vs. Controlled Phase Delay $\phi$

The output current-voltage characteristic is shown in Figure 3.10. The current source behaviour is again demonstrated as the output current stays approximately constant for a given phase delay irrespective of output voltage as long as the output voltage is reasonably high.

A plot of the operating $Q_2$ and the nominal $Q_{20}$ is shown Figure 3.11. The current source behaviour is again demonstrated as the slope of output voltage increases as the nominal $Q_{20}$ increases for a given phase delay.

The normalized reflected resistance, reactance and impedance are shown in Figure 3.12, Figure 3.13 and Figure 3.14, respectively. The normalization is performed with respect to:

$$Z_r = \frac{\omega M^2}{L_2} \quad (3-50)$$

As predicted by (3-17) and (3-18), the maximum impedance does not reach the peak when the converter is delivering full power at low $Q_{20}$. This can be seen clearly from Figure 3.14 for $Q_{20}=1$, the maximum reflected impedance is 40% higher than the value of the $Q_{20}$ of the circuit. This shows that if the secondary pickup requires power proportional to $Q_{20}$, the power
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supply has to source more than the real required power. However, for pickups with a $Q_{20}$ above 3, Figure 3.14 shows that the overrating of the normalized impedance of the primary converter is less than 10%. Most commercial systems use IPT pickups with the highest $Q_{20}$ obtainable ($Q_{20}=3-10$) to reduce the primary track required. Under these conditions, the overrating required by the power supply is insignificant. Furthermore, the high $Q_{20}$ also denotes that the simplified analytical analysis would be quite accurate for most practical systems.

Figure 3.15 shows the transient response of the AC processing pickup with a step in the controlled phase $\phi$ from 20-50º at time equal to 0.5ms. Similarly, Figure 3.16 shows the transient response of the AC processing pickup with a step in the controlled phase $\phi$ from 50-20º at time equal to 0.5ms. The simulation results using the expressions discussed in Section 3.2.3 are also added to the same figure for comparison purposes. Since the transient analysis only models the fundamental component, the peaks of the capacitor voltage calculated using (3-24) will be slightly less than the simulation results. Nonetheless, the first order approximation provides a good rule of thumb for the capacitor voltage under transient operation. It can be seen that the transient response of the pickup is always over damped with no overshoot or undershoot, so the components in the pickup only have to be rated for steady state operation.
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Figure 3.10. Pickup Output Voltage Current Characteristics.

Figure 3.11. Operating $Q_2$ vs. Nominal $Q_{20}$. 
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Figure 3.12. Reflected Resistance on Primary Track.

Figure 3.13. Reflected Reactance on Primary Track.
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Figure 3.14. Reflected Impedance on Primary Track.

Figure 3.15. Capacitor Voltage when Controlled Phase Delay $\phi$ Steps from 20-50°.
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3.3.4 Component Tolerance

To determine the circuit operation with variations in component values, the analysis in the former section will be used to produce normalized figures to aid the discussion. The AC processing pickup is simple and only consists of three circuit components \((L_2, C_2, R_2)\). Inductor \(L_2\) can be accurately measured using an LCR meter and its value hardly changes with aging. It may be argued that systems using ferrite or magnetic material may have large changes in the inductance value under movement, however, any variations in inductance can be considered compensated with an equivalent change in \(R_2\) and \(C_2\) parameters using the techniques below and in chapters 4-5. As a result, inductor \(L_2\) will be assumed to be fixed and variations of the components \(R_2\) and \(C_2\) will be discussed below.

3.3.4.1 Variations in \(R_2\)

Variations in \(R_2\) are quite common as the power demand of the load changes. A brief discussion was outlined in the previous section. The value of \(R_2\) is captured by the quality factor \(Q_{20}\) defined by (3-6). To demonstrate how the circuit operation varies with \(Q_{20}\), a normalized plot shown in Figure 3.8 is calculated using (3-28) and (3-38). The diagram shows
Chapter 3. The AC Processing Pickup.

a plot of the normalized output voltage \( V_c/V_{oc} \) versus the phase delay for different quality factors. The quality factor of the circuit can vary anywhere between the maximum load (corresponding to the minimum quality factor) and zero load (corresponding to an infinite quality factor). Under the maximum load condition, the circuit should give the desired output voltage at a phase delay of zero. Referring to Figure 3.8, if the load decreases from the maximum load (increasing \( Q_{20} \)), the controller simply increases the phase delay to regulate (or maintain) the output voltage. If the output voltage is required to be lower than the nominal value, the controller can further increase the phase delay to decrease the output voltage. Thus, any changes in load resistance \( R_2 \) do not sacrifice the controllable output voltage range of the circuit.

3.3.4.2 Variations in \( C_2 \)

Variations in capacitance \( C_2 \) from its nominal value are usually due to component degradation and tolerance. In conventional IPT pickups, the tuning capacitor \( C_2 \) is perfectly tuned to the inductor \( L_2 \), so that in (3-25) \( \alpha = 1 \). Under this condition, the capacitance value is defined to be the nominal value and the maximum output power occurs at zero phase delay.

When \( \alpha > 1 \), it is found that the capacitor voltage oscillation will not have reached zero at the point of switching if a phase delay of zero is used to control the switches. Switching at this instant will cause high shoot-through currents through the switch given by:

\[
I_s = \frac{V_c}{(R_{so})_{on}}
\]  

(3-51)

This current will most likely exceed the current rating of the transistor and cause device failure. However, if a controller is used to adjust the phase delay so that the switch can only turn on after the capacitor voltage has reached zero, the maximum power which can be obtained from the circuit will significantly decrease (dependant on the quality factor). Figure 3.17 shows the output power of the pickup when \( \alpha > 1 \) and a quality factor of 5 is used. Note that, when \( \alpha = 1.2 \) and 1.4, the controller has to adjust the phase delay to no smaller than 50° and 70° respectively, so that no shoot through current is present. As shown, the maximum
Chapter 3. The AC Processing Pickup.

voltage decreases substantially with $\alpha>1$. Thus, it is undesirable for capacitor $C_2$ to be larger than its nominal tuned value as it will decrease the maximum power transferred within the IPT system.

When $\alpha<1$, the characteristics of the circuit operation are also shown in Figure 3.17. For a capacitor that is 80% of the nominal tuned value, the maximum output voltage decreases by a small amount – less than 2%. But there is a significant advantage as it enables cheaper and less accurate capacitors to be used for tuning without compromising the maximum power of the IPT system.

Due to the finite tolerance of the capacitance in a practical IPT system, a capacitor which is smaller than the nominal tuned value (by slightly more than its tolerance level) should be used. This also allows for any slight changes in any other circuit components. The controller must be capable of accommodating small variations in the capacitor value.

![Figure 3.17. Normalised RMS Capacitor Voltage for Different C2 at Q20=5.](image)

Figure 3.17 shows the maximum obtainable normalized output voltage for a wide range of tuning capacitance. The maximum obtainable output refers to the maximum voltage that the controller can produce by varying its phase delay $\phi$. It can be seen that for a smaller than nominal tuning capacitance, the obtainable output voltage always decreases compared to the nominally tuned for all $Q_{20}$ conditions. However, the maximum obtainable voltage is not
sacrificed significantly as the gradient of the graph is low.

There are a few disadvantages that need to be mentioned for lower tuning capacitor values. The semiconductor switch ratings increase when a lower $\alpha$ value is used as the peak capacitor voltage increases as shown in Figure 3.19. Hence, as the $\alpha$ value decreases, the ratio of the maximum obtainable output voltage (or power transferred) in the IPT system over the device rating is reduced at a even greater rate. As such, the overall cost per kilo-watt of the system is increased. Moreover, the RMS current in the switch also increases as the $\alpha$ value decreases as shown in Figure 3.20. Thus there will be more conduction losses in the switch and the IPT system may have a lower efficiency. Furthermore, the reflected resistance and reactance varies for different $\alpha$ values as shown in Figure 3.21. The maximum reflected resistance is reduced for lower $\alpha$ values. This shows that the maximum power that can be transferred decreases - conforming to results shown in Figure 3.18. The reflected reactance also becomes more inductive at lower phase delays for smaller $\alpha$ values.

![Figure 3.18. Maximum Obtainable Normalised Output Voltage for different $\alpha$ values.](image-url)
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Figure 3.19. Peak Normalised Capacitor Voltage for different $\alpha$ values.

Figure 3.20. Normalised RMS Switch Current for different $\alpha$ values.
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Figure 3.21. Reflected Resistance and Reactance for $\alpha=0.9$ in (a), (b), $\alpha=0.8$ in (c), (d), $\alpha=0.6$ in (e), (f) and $\alpha=0.4$ in (g), (h).

Because of this, the $\alpha$ value should be chosen as high as possible, but also satisfying all the tuning component tolerances. In an actual design, a compromise has to be met by weighing
the cost of using accurate expensive tuning components with lower cost ones but with the disadvantages described above.

### 3.3.5 Component Stress

It is important to determine the maximum voltage and current ratings for the semiconductor switches used in an AC processing pickup. The maximum voltage which can occur across the switch and diode is the peak capacitor voltage. To determine this, it is first necessary to calculate the time at which it occurs. This can be found by differentiating (3-28) and setting the result equal to zero.

$$\frac{dV_c}{dt}(t_p) = \delta V_{cd} + \delta V_{cu} = 0$$

$$\delta V_{cd} = \frac{\partial \omega_0 e^{-t_p/T}}{\sin(\theta_v)} \left( -\sin(\omega_0 t_p - \theta_v) + \omega_0 \cos(\omega_0 t_p - \theta_v) \right)$$

$$\delta V_{cu} = \omega_0 \dot{\omega} \left( (1-a) \cos(\omega_0 t_p + \phi) + \sin(\omega_0 t_p + \phi) / Q_0^2 \right)$$

Substituting $t_p$ back into (3-28) gives the peak switch voltage.

$$(V_s)_{max} = V_c(t_p)$$

Since the time at which the switch is closed does not necessarily occur when the peak current in the inductor is at its peak, determining the maximum current through the switch and diode is not trivial. One possible method is to solve the inductor current equation using (3-38) and (3-46), and determine the point of maximum current by observing the waveform. This will be illustrated in Section 3.4.

### 3.4 Discussion and Experimental Results

In this section the design of an AC processing pickup for a 550W lighting system is described. The desired output voltage is 215V and the equivalent load resistance is 84Ω when the bulb has heated up. The pickup inductor has an open circuit voltage ($V_{oc}$) of 44.5V and an inductance of 72.6μH. The IPT system is operating at a resonant frequency of 38.4 kHz.
The first step is to determine the tuning capacitance for the circuit. From (3-26), the nominal tuning capacitance is 239nF. Using the approach described earlier, a smaller capacitance of 227nF is chosen allowing for 4% tolerance. In this case, variable “α” is equal to 0.96. From (3-6), $Q_{20}$ for the circuit is 4.8.

Equations (3-28) and (3-38) are then used to solve for steady state operation. Figure 3.22 and Figure 3.23 show the calculated waveforms for the circuit under both rated and 50% load. The calculated peak and RMS values of the voltage and current for each component are listed in Table 3.1 and Table 3.2. It should be noted that the equivalent load resistance of the light will be smaller at lower output voltages.

![Capacitor Voltage](image1)
![Inductor Current](image2)
![Capacitor Current](image3)
![Switch Current](image4)

Figure 3.22. Operating Waveforms at Rated Load showing (a) Capacitor Voltage, (b) Inductor Current, (c) Capacitor Current and (d) Switch Current.
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Figure 3.23. Operating Waveforms at 50% Load showing (a) Capacitor Voltage, (b) Inductor Current, (c) Capacitor Current and (d) Switch Current.

Table 3.1: Voltage and Current of Components at Rated Load.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Peak</th>
<th>RMS</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_c$</td>
<td>304V</td>
<td>212V</td>
</tr>
<tr>
<td>$I_L$</td>
<td>17.8A</td>
<td>12.6A</td>
</tr>
<tr>
<td>$I_c$</td>
<td>17.7A</td>
<td>12.2A</td>
</tr>
<tr>
<td>$I_s$</td>
<td>17.4A</td>
<td>2.21A</td>
</tr>
</tbody>
</table>

Table 3.2: Voltage and Current of Components at 50% Load.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Peak</th>
<th>RMS</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_c$</td>
<td>224V</td>
<td>150V</td>
</tr>
<tr>
<td>$I_L$</td>
<td>14.5A</td>
<td>10.5A</td>
</tr>
<tr>
<td>$I_c$</td>
<td>14.5A</td>
<td>9.39A</td>
</tr>
<tr>
<td>$I_s$</td>
<td>14.4A</td>
<td>3.27A</td>
</tr>
</tbody>
</table>

The last step is to find the component stress and rate the semiconductor devices for normal
operation. Figure 3.24 shows the calculated component stress using the equations in the above section. As shown, the switches and diodes have to be rated for at least 310V at 18A. It should be noted that the peak values in this system happen close to rated load.

In order to validate the above design procedure, the AC processing pickup determined above was built. The controller for the AC processing pickup is drawn as a block diagram shown in Figure 3.25. The phase of \( V_{oc} \) is measured using a separate phase sense coil \( L_3 \) placed on the primary track to detect the phase of the track current - which is exactly 90° out of phase with the open circuit voltage \( (V_{oc} = j\omega I_1) \). The phase delay \( \phi \) is set by the computer interface and a microcontroller is used to adjust the switch drive waveforms accordingly.

Figure 3.26 and Figure 3.27 show the circuit waveforms for the AC processing pickup at rated and 50% load, respectively. The measurements are within 10% of the values calculated by the equations, except for the switch current. Since the current waveforms have a square wave shape with significant high frequency harmonics, the 100 kHz bandwidth current probe does not have the bandwidth to measure the waveform accurately. Despite this, it is evident that the correlation between the calculated waveform and experimental results is very good.
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Figure 3.25. Block Diagram of Controller.

Figure 3.26. Operating Waveforms at Rated Load showing (a) Capacitor Voltage and Inductor Current, (b) Switch Current and Capacitor Current.
Figure 3.27. Operating Waveforms at 50% Load showing (a) Capacitor Voltage and Inductor Current, (b) Switch Current and Capacitor Current.

Waveforms for different phase delay are also shown in Figure 3.28 to Figure 3.30. The first and second traces are the $V_{g1}$ and $V_{g2}$ PWM gate signals driving switches $S_1$ and $S_2$ respectively. The third trace is the open circuit voltage $V_{oc}$ and the fourth trace is the capacitor voltage $V_c$. For the first set of waveforms, the capacitor voltage is approximately 50V (Figure 3.28 (a)) when the time delay is 6.1µs (Figure 3.28 (b)), corresponding to a phase delay of 87.8°. Similarly, the second set of waveforms in Figure 3.29 shows the output voltage is approximately 100V when the phase delay is 67.8°. Likewise, the third set of waveforms in Figure 3.30 shows the output voltage is approximately 210V when the phase delay is 23°.
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Figure 3.28. Operating Waveforms at 47V showing (a) $V_{oc}$ and $V_c$ waveforms and (b) Magnified Version of the previous Oscillographs

Figure 3.29. Operating Waveforms at 100V showing (a) $V_{oc}$ and $V_c$ waveforms and (b) Magnified Version of the previous Oscillographs

Figure 3.30. Operating Waveforms at 217V showing (a) $V_{oc}$ and $V_c$ waveforms and (b) Magnified Version of the previous Oscillographs
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The AC processing pickup was applied to an AC load comprising a 550W incandescent light bulb bank. Figure 3.31 shows that the output power can be controlled by adjusting $\phi$. Note that although the control range is from 0-180º, the phase controller only needs to change between 0-120º to regulate power over the entire operation range. The analytical results are also plotted on the same figure for comparison purposes. Since the analytical analysis ignores the ESR losses in both the pickup inductor and tuning capacitor, and the losses in the switches and diodes, the output power is higher than that obtained from experimental measurements. Despite this, the difference between the two is not greater than 10% as the controller is quite efficient.

An efficiency vs. output power plot is shown in Figure 3.32 for the processing circuit. This efficiency is determined using measurements of the input power to the primary converter and the output power from the secondary pickup after subtracting 40W of primary converter no-load loss. However, the primary converter loss is likely to be higher at higher load in practice, so the efficiency measurements are conservative. The efficiency of the system remains above 90% when an output power of more than 100W is delivered to the load. With a 550W load, the efficiency can reach as high as 96%.

![Output Power vs. Controlled Phase Delay](image)

**Figure 3.31. Output Power vs. Controlled Phase Delay $\phi$.**
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![Efficiency vs. Output Power](image)

Figure 3.32. Efficiency vs. Output Power.

### 3.5 Summary

This chapter presented a new IPT pickup which has significant advantages compared to traditional pickups that use AC-DC-AC conversion topologies for producing a controllable AC output voltage. This simple AC processing pickup controller can provide controlled AC power over a wide resistive load range. In addition, the simple pickup circuitry has advantages in terms of increasing system efficiency and reducing the pickup size. The fundamental properties and design equations of the AC processing pickup have been investigated.

The AC processing pickup has some disadvantages. It has a poor reflected power factor on the primary track and consequently the primary converter must be overrated to source the extra VA. In addition, the output voltage always has harmonics present in it due to the clamping action of the switches and diodes. Moreover, it requires knowledge of the phase of the open circuit voltage which can be obtained by placing an additional phase sense coil onto the primary track. In chapter 4 of this thesis, a new method for detecting the phase indirectly is presented.
In this chapter, a simple power balance analysis was presented to demonstrate the operation of the pickup. In addition, normalized plots were produced to aid the understanding of the pickup operating characteristics. It was found that the pickup demonstrates desirable current source properties which may be useful in battery charging applications. Moreover, an exact analysis obtained switch and component ratings for the pickup to use in actual designs. Finally, a 550W prototype circuit was designed to verify the validity of the theoretical analysis. The measurements are within 10% of the values calculated by the equations. A maximum efficiency of 96% was measured with the new AC processing pickup driving a 550W lighting fixture.
4 Practical Implementation of AC Processing Pickups

In this chapter, some practical implementation aspects for parallel AC processing pickups are outlined. Firstly, the normalised characteristics of the pickup operation under steady state are captured in a graphical format. If the circuit is required to operate with feed-forward control to improve its transient response, a microcontroller is required to store each control phase delay $\phi$ corresponding to their output voltage value. If the load or $Q_{20}$ of the circuit changes during operation, then the controller has to store a range of output voltage values for all the $Q_{20}$ values and phase delays $\phi$. However the large number of figures required to show these characteristics results in many operating points that would normally be needed to be stored within a microcontroller to properly represent these two dimensional surfaces for control purposes. Here, a fourth order polynomial approximation is used to produce a simple look-up table and reduce the number of stored points required so that direct implementation is possible.

Secondly, the ratings required for the AC switch may be high in a high power AC processing pickup. As such, the ratings of the best practical switches available may not meet the required conditions. In consequence, series and parallel connection of switching devices must be used in order to achieve the necessary ratings of the AC switch in the proposed AC processing pickups, especially when high power applications are considered. Thirdly, a new PWM control strategy is outlined to decrease the conduction losses in the switches of a parallel AC processing pickup, and at the same time reduce the cost and size of such systems. Lastly, a 1.2kW lighting system is built to illustrate the operation of a closed-loop parallel AC processing pickup controller.
4.1 Normalised Characteristics of the AC Processing Pickup

A limited set of normalised graphs were used in chapter 3 relating to the operation of the parallel AC processing pickup. This section undertakes a more detailed analysis of each pickup and presents a full set of normalised graphs under various practical operating conditions. Using a similar set of analytical techniques described in chapter 3, the normalised output characteristics are computed and shown in Figure 4.1 for the parallel AC processing pickup. The y-axis shows either the normalised output voltage or output current and the x-axis shows the phase delay angle. These output voltages and currents are normalised with respect to the open circuit voltage or the short circuit current. The calculations are done for a wide range of $Q_{20}$ conditions. Different tuning capacitance values are also shown in each subplot where $\alpha$ denotes the ratio of the actual tuning capacitance against the nominal ideally tuned capacitance ($C_2 = 1/(\omega L_2)$) as described in chapter 3.

As shown, the output voltage can be fully controlled over a wide range of load conditions ($Q_{20}$) by simply controlling the phase delay angle. When the tuning capacitor is lower than its nominal value (i.e. $\alpha < 1$), the maximum point is no longer at zero phase delay. The reason behind this is quite simple. For the parallel AC processing pickup, an AC switch in parallel with the resonant capacitor directly controls the fundamental phase of the capacitor voltage relative to its current. A greater phase delay will result in the capacitor voltage lagging its current, hence increasing its equivalent capacitance. Because the pickup uses a lower than nominal tuning capacitance when $\alpha < 1$, this increase in the “equivalent tuning capacitance” with $\phi$ causes the AC processing controller to retune the parallel resonant network and eventually reach its maximum power point. However, for selected $C_2$ values that are much lower than nominal, e.g. $\alpha = 0.2-0.1$, the maximum output voltage obtainable is substantially lower than theoretically expected because the AC switch generates significant harmonics in the resonant circuit so that the operation of the pickup can no longer be simply approximated by its fundamental components.
For control purposes, the normalised characteristics are very useful without requiring a large number of graphs and tables to define all possible operating conditions. This is especially true if more operating waveforms such as the peak switch voltage, the RMS and peak of the pickup inductor current and the RMS and peak switch current are also of concern. Each one of these waveforms requires a number of these normalised graphs to illustrate any variations with each of the three common variables, namely, phase delay $\phi$, $Q_{20}$ and $\alpha$. Consequently, it is difficult to handle so much data when considering direct implementation in a microcontroller, and it is also difficult for designers to use.

A linear regression method using least squares may be used to approximate the normalised curves computed by these analytical routines and thereby reduce the number of values or graphs required for control purposes. A multiple regression analysis technique was initially performed on the normalised surface of the two variables, phase delay $\phi$ and $Q_{20}$. However, this technique was found to be inaccurate especially when the variation in output voltage can be ten times or more across the nominal range of $Q_{20}$ values ($1 < Q_{20} < 10$). The error for smaller $Q_{20}$ values on the 2D surface can be as high as 20% because larger $Q_{20}$ values dominate the approximation.

The accuracy was greatly improved when a 4th order least squares polynomial was used to approximate the solution using only one input variable, phase delay $\phi$. All other variables are considered as constants and inherently incorporated into the approximation. In this way the approximation error can be kept as low as 1.5% (as shown in Figure 4.2). As such, many polynomial coefficients for each individual $Q_{20}$ and $\alpha$ value are required. If finer resolution is required, interpolation can be used between two sets of polynomials.
Figure 4.1. Normalised Plots for Parallel AC Processing Pickup when (a) $\alpha=1$, (b) $\alpha=0.8$, (c) $\alpha=0.6$, (d) $\alpha=0.4$, (e) $\alpha=0.2$, (f) $\alpha=0.1$. 
Chapter 4. Practical Implementation of AC Processing Pickups.

Figure 4.2. Approximation Error using 4th Order Least Squares Polynomial for α=1.

To approximate the output characteristics, the approximation function is in the form of:

\[ f(V_o, V_s, I_s, V_L, I_L) = a_4 \phi^4 + a_3 \phi^3 + a_2 \phi^2 + a_1 \phi + a_0 \]  \hspace{1cm} (4-1)

And the values to the polynomial can be obtained from:

\[
A = \begin{bmatrix} a_4 \\ a_3 \\ a_2 \\ a_1 \\ a_0 \end{bmatrix} = (X^T X)^{-1} X^T Y \hspace{1cm} (4-2)
\]

Where \( X \) is the data matrix for the phase delay and \( Y \) is the data matrix for the normalised output data of concern such as the normalised output voltage. Because the \( X \) matrix is most likely not square, a direct matrix inversion is not possible. As such, a pseudo-inverse process is usually required as illustrated in (4-2), where the transpose of matrix \( X \) must be added to convert it into a square matrix so that matrix inversion can take place. A thorough computation process is undertaken to generate a normalised table shown in the Appendix B and C of this thesis. This not only includes the normalised output voltages but it also captures many other useful waveforms of concern for both the series and the parallel AC processing pickups. In addition, these polynomials can be easily stored on a microcontroller,
feed-forward control can be implemented.

### 4.1.1 Normalised Characteristics using Clamp Time Control

Clamp time control is another feedback mechanism that can be used to synchronise and control the parallel AC processing pickup. Rather than synchronising the gate control waveforms to the phase of the open circuit voltage, clamp time synchronisation can be used for control purposes instead. The clamp time in parallel AC processing pickups always has a relationship with the phase delay and hence can be used as the controller input. The key advantage of this technique is in applications where an open circuit voltage cannot be easily measured using a secondary pickup (as discussed in Section 3.4). An example of this is in battery charging applications in which ferrite is used in both the primary and secondary. Such systems are highly coupled with lumped coils on both the primary and secondary so that it is almost impossible to get an independent “separately coupled” phase sense coil that only senses the primary track current without interference from the secondary pickup. Another common application is biomedical implants. Consequently clamp time control is essential in these applications.

The fundamental relationship between clamp time $t_c$ and controlled phase delay $\phi$ is non-linear, extremely complex and dependent on many circuit operating parameters. However, some basic principles do exist between these two variables. Firstly, the clamp time is always changing in the same direction as phase delay. Although the rate of change between the two variables is rather complex, the gradient between the two variables is always positive. Secondly, the relationship between the two variables is time-invariant. This only refers to the case when an AC processing pickup is operating under steady state operation so that the relationship that governs the two variables never changes with time.

#### 4.1.1.1 Practical Realisation of Controller for Clamp Time Control

A typical $Q_{20}$ dependent operating waveform for a parallel AC processing pickup is shown in Figure 4.3. To realise a clamp time controller practically, some feedback signal is required to detect when the circuit enters the clamp state. The two waveforms that show when the
circuit has entered the clamp state are the resonant capacitor voltage and the AC switch current as shown in Figure 4.3. The capacitor voltage changes polarity when the circuit enters the clamp state. Similarly, the AC switch current starts to flow when the circuit enters the clamp state. Each of these detection methods will be discussed below.

**Figure 4.3. Typical Operating Waveform for a Parallel AC Processing Pickup.**

### 4.1.1.1 Capacitor Voltage Detection

When the circuit enters the clamp state, the body diodes will start to conduct and stop the resonant capacitor voltage building up in the opposite direction. For the body diode to turn on, the voltage across the capacitor must change in polarity and be biased at the forward voltage drop of the body diode. This polarity change is the essence for detection of the clamp state. A
Chapter 4. Practical Implementation of AC Processing Pickups.

A comparator can be used as shown in Figure 4.4, to detect this change. However, this technique is highly sensitive to noise, especially under typical switching environments. This is because the circuit is trying to detect a voltage polarity change of approximately zero when the resonant capacitor voltage can be as high as several hundred volts under normal operation. The precision required in the comparator is very high and very careful PCB layout is required to get such systems to operate correctly. Consequently, a more practical technique that measures the switch current is preferred in a real circuit.

![Comparator Measuring Resonant Capacitor Voltage](image)

**Figure 4.4. Comparator Measuring Resonant Capacitor Voltage.**

### 4.1.1.1.2 Switch Current Detection

When the circuit enters the clamp state, the body diodes will immediately start to conduct the resonant pickup current. Detecting this change allows the start of the clamp state to be precisely determined. A Current Transformer (CT) can be used to detect the switch current and transform this into a voltage signal with lower amplitude. A comparator is used to generate a 50% duty cycle square wave which may be used for the microcontroller to show when the circuit entered the clamped state as shown in Figure 4.5. Because this technique measures the current through the switch, it is much less prone to noise. In addition, during startup, the parasitic current flowing through the body diodes is sufficient to accurately trigger this detection circuit. This technique is now widely adopted in the design of AC processing pickups used in EV battery charging applications due to its robustness and simple circuitry.
4.1.2 Normalised Characteristics of Clamp Time Controllers

To obtain the normalised characteristics of a clamp time controller, a full set of simulations is required for every clamp time condition. However, there is an alternative method to obtain these characteristics. Since the normalised characteristics outlined in the previous section based on the phase delay are already calculated, the relationship between clamp time and phase delay is all that is required. It is quite easy to deduce the clamp time, as the calculation routine in Figure 3.7 inherently calculates the clamp time. Figure 4.6 shows a plot of the calculated results for the clamp time against different phase delays. Both axes are normalised to the IPT system period, shown in degrees.
To calculate the normalised characteristics for the clamp time, a very simple mapping routine can be used to combine the normalised characteristics outlined in section 3.3.3 and Figure 4.6. The result is shown in Figure 4.7.

**Figure 4.6. Clamp Phase vs. Phase Delay for $\alpha=1$.**

Although the clamp time controller has the advantage of controlling the circuit without a phase sense coil, the sensitivity of the control circuit is much higher in high $Q_{20}$ circuits as shown in Figure 4.7. Despite this, clamp time controllers may be the *only* practical option in IPT systems where both the primary and the secondary inductor coils are lumped.
4.2 Series and Parallel Connection for Switching Devices in AC Processing Pickups

Series and parallel connection of semiconductor switches has been widely used in high voltage or high current power electronic systems [107-123]. In some applications, there is an attractive proposition that for a given power rating, the operating voltage increase corresponds to a reduction in current. For high voltage applications, a series connected topology is attractive because this is lower cost and has lesser complexity in circuit design and control compared to multi-level circuit topologies [122]. The main concern of the series connection is unequal voltage distribution between the devices under both transient and steady state conditions. The operating voltage of the series connected device is higher than the allowable individual device operating voltage and the voltage rating must be shared ideally equally between the series connected devices so that no device has its ratings exceeded. However, due to the spread of device inherent parameters and PCB layouts, unequal voltage sharing between the devices exists in practice and if the voltage unbalance causes an individual device voltage rating to be exceeded as shown in Figure 4.8, failure of the device will occur. Section 4.2.1 will discuss the solutions to these issues in detail.

![Figure 4.8. Voltage Unbalance between Switches.](image)

In high current applications, paralleling of devices can be used. Static and dynamic current

- 89 -
sharing must also be guaranteed. The main concern for asymmetric current sharing is the tolerance of inherent device parameters, propagation delay times of the various gate drive circuits, different stray inductances, and poor circuit layout. Current imbalance can cause some devices to carry higher currents than their individual ratings and subsequent failure will occur. Section 4.2.2 discusses how these issues are resolved.

In AC processing pickups, the AC switch is usually exposed to both the full resonant voltage and current. For example in a parallel tuned AC processing pickup, the voltage across the AC switch and current through it has to be rated for:

\[
V_s = Q_s V_{oc} \sqrt{2} \tag{4-3}
\]

\[
I_s = \sqrt{Q_s^2 + I_{oc}^2} \sqrt{2} \tag{4-4}
\]

From (4-3) and (4-4), the AC switch must be rated for the maximum VA's in the resonant tank – not just the real power that is delivered to the load. The peak switch voltage may easily exceed 600V in high power applications. Choosing a single semiconductor module that can handle these operating conditions in a high power application may be difficult. There is a rapid growth in silicon carbide semiconductor devices that have higher ratings, however these devices are in their infancy and not available in mass production. As a result, a discussion on the use of series and parallel connected silicon devices to allow for high operating conditions is presented for the AC processing pickup.

### 4.2.1 Series Switch Connection

With a series connection of power semiconductor devices, the important issue is to maintain voltage sharing among all members of the series string during both the steady state and transient states. Different methods are used for different types of devices. Since voltage imbalance is mainly a result of device parameter spread and gate drive delay, careful selection of power semiconductor devices that have low parameter spread whilst ensuring synchronisation of gate drive signals will minimise the problems associated with series connection. However these techniques suffer from reliability issues as parasitic elements in
Chapter 4. Practical Implementation of AC Processing Pickups.

the device still cause voltage imbalances. In addition, most of the parameters are temperature dependent which can contribute to thermal run-away or localised losses in one component in the string [118]. Hence practical design techniques are introduced below for series connection of devices.

4.2.1.1 Static Voltage Balance

Static voltage balance refers to a uniform spread of voltage across all switching elements in a string of devices under steady state conditions. A very simple method of achieving such a condition is to match the output impedance of all switching devices using a simple voltage divider as shown in Figure 4.9. The resistances within the potential divider should be chosen to be 10 times lower than the off state output impedances of each semiconductor device. If smaller resistances are chosen for the voltage divider, static voltage sharing between the devices can be achieved with greater accuracy, however, the power losses as a result of the lower resistance may be large. Hence, balancing the power losses by the resistor divider and the error in voltage between series devices is a trade-off. With the rapid advancement of semiconductor devices, these trade-offs are usually easily met even at kilo-volt levels due to the high blocking resistance in most modern day semiconductor switches such as MOSFET’s and IGBT’s. An alternative solution is to use power semiconductor switches manufactured with the ability to withstand any avalanche breakdown, and can operate at IPT frequencies. This would eliminate the need for a voltage divider network. Using the avalanche handling capability, if there is uneven voltage sharing between devices, the device that has higher voltage will enter the avalanche breakdown mode where the voltage across it, is inherently limited to its maximum voltage rating, with negligible loss, as the current through the switch during an avalanche state must be very low.
4.2.1.2 Dynamic Voltage Balance

For dynamic voltage balance, the voltage across the switches must transiently balance during turn on and turn off. One key switching condition for the parallel AC processing pickup outlined in chapter 3 is that it achieves ZVS while the AC switch is operating. As such, the voltage across the switching devices must be close to zero when the switch transiently changes its condition. Hence, dynamic voltage balance is achieved inherently in the parallel AC processing pickup as the voltage across a series string of devices will be ideally zero.

Although inherent voltage balance is achieved, the propagation delay of the extra isolated gate drive stage necessary to drive the high side switches in any string of devices may pose issues for dynamic voltage balance. If there is a significant delay in the gate drive signal for the upper switch in a series switch pair as shown in Figure 4.10(a), then all the voltage will appear across the bottom switch which may cause device failure. The waveforms in Figure 4.10(b) are: gate drive for the top switch, gate drive for the bottom switch, the voltage across the bottom switch, the voltage across the top switch and the resonant capacitor voltage. As can be seen the voltages across the two devices are no longer transiently balanced under this condition.
Chapter 4. Practical Implementation of AC Processing Pickups.

![Diagram](image)

**Figure 4.10. (a) Series Switch Connection and (b) Operating Waveforms due to Delay in Gate Drive Signal.**

If it is assumed that the voltage across the resonant capacitor are half sinusoids ($Q_{20}$ above 3 will satisfy this approximation), then the maximum delay time allowable between each gate drive signal in the top and bottom switch of Figure 4.10(a) (or the time the circuit is in State 1 in Figure 4.10) is given by:

$$
t_{\Delta g} = \arcsin \left( \frac{V_{s_{\max}}}{V_{c_{2_{\max}}}} \right) / \omega
$$

(4-5)

where $t_{\Delta g}$ is the delay in gate signal between the two switches, $V_{s_{\max}}$ is the switch voltage rating, $V_{c_{2_{\max}}}$ is the peak resonant capacitor voltage and $\omega$ is the operating frequency.

State 2 is the state when voltage across the both switches starts to rise, however the rate of change of voltage as shown is higher for the top switch. This is because the parasitic output capacitance of the switch charges faster when its output voltage is lower. In State 3, the bottom switch enters avalanche mode where the voltage across it is clamped. The losses of the switch in this state is usually very low (<0.1W for 1kW, 1kV converters) and can be neglected in most loss analysis.

### 4.2.2 Parallel Switch Connection

When power semiconductor devices are connected in parallel, the important issue is to
maintain current sharing among each parallel path during both steady state and transient switching conditions.

4.2.2.1 Static Current Balance

In practice, static current sharing is mostly achieved using switches that have a positive temperature coefficient. Devices such as MOSFET’s can inherently achieve current sharing as the resistance in the device will rise when its temperature is higher than any other device. As a result, it will carry less current and the temperature will be shared among all parallel devices.

4.2.2.2 Dynamic Current Balance

For dynamic current balance, the current through the switches must transiently balance during turn on and turn off. To achieve dynamic current sharing for paralleled devices in parallel AC processing pickups, more complex methods must be used. Dynamic current imbalance is mainly due to the spread of transfer characteristics and the external gate drive circuit. Hence, circuit layout and gate drive play an important role in the parallel operation of switching devices. There are many possible techniques for achieving dynamic current sharing such as de-rating of paralleled devices [116], closed-loop gate control of the switching devices [112] and impedance balancing [108]. Assuming a sensible layout is used, the most common method is applied to de-rate each of the devices to ensure reliable option. The basic principle behind this technique is that some mis-sharing is allowed between parallel devices based on expected variations, whilst ensuring that none of the devices exceed their maximum current and temperature rating. This requires the device current rating to be de-rated by a de-rating factor, which is a function of the number of devices in parallel and their mismatch factor [116]. The literature outlined in [122] evaluated appropriate de-rating factors for dynamic current sharing and thermal stability. However, external circuit conditions are assumed to be ideal. An investigation described in [116], showed that parameter spread has less influence on rate of change of current when compared to the possible values of emitter ground parasitic inductance. For an optimum switching behaviour of parallel devices, it is necessary to drive the gates with only one common resistance and to balance emitter-ground wiring. It is usually
good to assume a sufficiently large de-rating factor of more than 30% for prototype systems as the cost of the devices is not very important. For mass manufactured systems where cost is important, thorough simulations and practical measurements of device packaging and PCB layout must be undertaken to carefully choose a de-rating factor such that the reliability is not compromised with lower de-rating factors.

### 4.3 Improved PWM Control for the Parallel AC Processing Pickup

In high power applications, the conduction losses in the parallel AC processing pickup may become significant in the AC switch. The effects are more severe when the body diode of the MOSFET is used in the conduction path. It may be argued that these losses may only be a very small portion of the overall losses in the pickup and the overall efficiency is only decreased marginally. However, it is always desirable to reduce the losses. As such the cost of the heatsink and the size of the PCB board can be reduced. In this section, a new duty cycle control method is proposed to reduce the conduction losses of the body diodes in the switches without substantially increasing the complexity of the control.

#### 4.3.1 Circuit Operation using Greater than 50% Duty Cycle Control

The key idea to reducing conduction losses is to use the MOSFET on-state resistance to carry the current through the switch rather than the body diode. If the on resistance is low enough, the current will preferentially conduct through the MOSFET channel instead. If this method is used, the voltage drop across the on resistance of the MOSFET must be lower than the body diode voltage drop, in order for the MOSFET channel to carry the current. Hence, this technique may require paralleling many MOSFET switches for high current applications, so that the on resistance is low enough. Figure 4.11 shows a circuit diagram of a parallel AC processing pickup using a back to back MOSFET switch pair to achieve the AC switch.
To achieve the condition that the MOSFET channel carries the current, the gate drive waveforms must also be modified as shown in Figure 4.12. Since the gate control waveforms are greater than a 50% duty cycle, the name for the controller implementation chosen is “greater than 50% duty cycle control”. It can be seen that the circuit waveforms under normal operation is identical to the waveforms shown in Chapter 3. It should be noted here that, the feedback detection to produce a controlled phase delay to adjust the power transfer is now using the switch current feedback signal described in Section 4.1.1.1.2 rather than that achieved using $V_{oc}$ reference. This is because the new control technique uses the information when the circuit enters the clamped state, and this information is only provided using clamp time control.
4.3.2 **Comparison of Loss between Original PWM Control and Greater than 50% Duty Cycle Control**

An example of the loss reduction for a parallel AC processing pickup using the new control strategy is presented in this section. A comparison of power loss for a 100kW DC output AC processing pickup with a rectifier is shown in Figure 4.13. A more thorough discussion of cascading a rectifier with an AC processing pickup will be outlined in chapter 5. Similar to before, the diode will carry the resonant current during the clamp state if the original phase delay control method is used. For the greater than 50% duty cycle control method, the current preferentially flows through the MOSFET channel rather than the body diode. The diagram
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shown in Figure 4.14 shows the new current conduction path for the pickup using greater than 50% duty cycle control. There are four modes of operation as stated below:

a. $V_c$ positive and not clamping – the DC current is equivalent to the difference between the inductor current and the capacitor current.

b. $V_c$ negative and clamping – the capacitor current is zero. The resonant current now preferentially conducts through the switch resistive channel rather than the body diode.

c. $V_c$ negative and not clamping.

d. $V_c$ negative and clamping.

To generate the gate control signals stated above, the controller diagram shown in Figure
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4.15 is used. The controller is very similar to the one outlined in the Section 4.1.1.1.2 where the switch current is used to detect the clamp time. Only two extra OR gates are added after the microcontroller to realise greater than 50% duty cycle control.

![Controller Diagram]

**Figure 4.15. Controller for Greater than 50% Duty Cycle Control.**

The simulation results of the power losses in the AC switches for a 100kW AC processing pickup are shown in Figure 4.16 using parameters from Table 4.1. Power losses are plotted against clamp time control and the red and blue traces are a single switch loss in a pair of switches using greater than 50% duty cycle control and the original phase delay control, respectively. It can be seen that for this particular example the maximum losses using greater than 50% duty cycle control are more than halved compared with the original control technique. Consequently in high power applications where a large number of MOSFET’s is paralleled to handle the current ratings, it may be desirable to use greater than 50% duty cycle control to reduce the losses in the AC switch. As such, the cost of the heatsink required can be greatly reduced, at the cost of marginally increasing the complexity of the controller. In addition, the reduction in losses may lead to smaller and lighter pickup systems.
Table 4.1. Parameters for 100kW AC Processing Pickup.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{oc}$</td>
<td>300V</td>
</tr>
<tr>
<td>$L_{dc}$</td>
<td>2mH</td>
</tr>
<tr>
<td>$R_{on}$</td>
<td>0.001Ω</td>
</tr>
<tr>
<td>$V_f$</td>
<td>2V</td>
</tr>
<tr>
<td>$L_2$</td>
<td>7.96uH</td>
</tr>
<tr>
<td>$C_{dc}$</td>
<td>100uF</td>
</tr>
<tr>
<td>$R_L$</td>
<td>0.9Ω</td>
</tr>
<tr>
<td>$\omega$</td>
<td>$1.257 \times 10^5$ rad/s</td>
</tr>
<tr>
<td>$R_{Ldc}$</td>
<td>0.01Ω</td>
</tr>
</tbody>
</table>

Figure 4.16. Simulation Results of Power Loss Comparison between Original PWM Control and Greater than 50% Duty Cycle Control.

4.4 Practical Lighting System Design

A design example of a 1.2kW AC processing pickup used in stage lighting applications is presented in this section. A block diagram of the lighting system is shown in Figure 4.17. The switch current detection method outlined in section 4.1.1.1.2 is used. The RMS voltage is first rectified and then filtered to DC inside the microcontroller. Next, the microcontroller samples this DC voltage using an ADC and adjusts the phase of the gate drive waveforms for the AC switch accordingly. The design specifications and system parameters are shown in Table 4.2. A primary LCL power supply is used operating at 20kHz and producing a track current of
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125A. An asymmetrical shaped S pickup is used with an open circuit voltage of 84.9V and a short circuit current of 5.86A. The tuning capacitor is chosen lower than the nominal with an α value of 0.9. The equivalent resistance of the OSRAM 1.2kW 220V lamp changes from 3-40Ω over the full output current range. This is because the bulb resistance is temperature dependent and the current flowing through it directly determines its temperature.

![Block Diagram of Lighting System](image)

**Figure 4.17. Block Diagram of Lighting System.**

<table>
<thead>
<tr>
<th>Spec. Parameters</th>
<th>Values</th>
<th>Design Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_1$</td>
<td>125A</td>
<td>$L_2$</td>
<td>115.2uH</td>
</tr>
<tr>
<td>$P_{out}$</td>
<td>1.2kW</td>
<td>$V_{oc}$</td>
<td>84.9V</td>
</tr>
<tr>
<td>$(V_c)_{rms}$</td>
<td>220V</td>
<td>$I_{sc}$</td>
<td>5.86A</td>
</tr>
<tr>
<td>$R_2$</td>
<td>3Ω - 40Ω</td>
<td>$\alpha$</td>
<td>0.9</td>
</tr>
<tr>
<td>$f$</td>
<td>20kHz</td>
<td>$C_2$</td>
<td>495nF</td>
</tr>
</tbody>
</table>

Table 4.2. Specifications and System Parameters.

For the lighting system, the output voltage vs. clamp time is shown in Figure 4.18. It can be seen that the majority of the output voltage range can be achieved by using a clamp time range of 2.7-15 μs. A clamp time that is lower than 2.7 μs will result in higher output voltages than the rated voltage of the lamp and these characteristics are not measured. By observing Figure 4.18, the phase angle sensitivity for this system is not a practical problem, as the microcontroller (PSOC29466) can easily achieve clamp time increments of around 100ns.
A more detailed block diagram of the microcontroller is shown in Figure 4.19. A second order low-pass filter with a cut off frequency of 2kHz is used to remove the high frequency components of the rectified output voltage. The cut off frequency should be chosen as high as possible to allow fast response of the system without compromising the filtering action to remove the 40kHz component and its harmonics produced by the rectifier. The filtered output is sampled at 3.2 kilo-samples per second (ksp/s) which is the maximum the PSOC29466 can handle for this application. This is then compared against the reference set by the user. Note that there is a relationship between the RMS and ADC value and a conversion block is required. A simple PI controller is implemented with “anti-windup”. The output of the PI controller will drive the phase modulation block which generates the gate drive signals for the AC switch.
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Figure 4.19. Block Diagram of Microcontroller.

The measured relationship between the RMS capacitor voltage vs. the ADC value is shown in Figure 4.20 by the blue trace. The ADC values shown are represented in the form of an 8 bit output with a value from 0-255 representing 0-5V. A similar polynomial approximation method to section 4.1 is used here to determine the ADC value for a RMS input. The approximated function is shown in the red trace and the correlation between the two is very good. The approximation function is given by:

\[
\left( V_{e_{-ref}} \right)_{adc} = 1.4489 e^{-7} \left( V_{e_{-ref}} \right)_{rms}^4 - 7.4644 e^{-5} \left( V_{e_{-ref}} \right)_{rms}^3 + 0.012348 \left( V_{e_{-ref}} \right)_{rms}^2
\]

\[
-0.14037 \left( V_{e_{-ref}} \right)_{rms} + 4.2905
\]  

(4-6)
Figure 4.20. RMS Voltage vs. ADC Value.

Figure 4.21 shows the circuit waveforms when the output voltage is regulated to 220V, 140V and 50V respectively. These correspond to 100%, 50% and 10% of rated power. The waveforms show the resonant capacitor voltages or output voltages (upper trace) and the pickup inductor currents (lower trace).

Figure 4.21. Operating Waveforms of Lighting System for (a) 100%, (b) 50% and (c) 10% Output Power.

For the closed loop system, the output RMS output voltage is measured using a differential voltage probe and is compared to the user input. The error between the two is plotted in Figure 4.22. It can be seen that the absolute error in output voltage does not exceed 5V. The two red traces show the quantization error of the 8 bit ADC in the PSOC. Only when the
voltage drops near 30V does the measured error exceed the quantization error, as the signal diodes in Figure 4.17 start to introduce non-linearities near its forward voltage.

![Figure 4.22. Error in Output Voltage.](image)

The transient response of the lighting system was measured. Figure 4.23 shows the step response of the output voltage from 30V – 220V. The top trace represents the lamp voltage and the bottom trace represents the lamp current. The controller is set with a PI gain of $K_p=1$, $K_i=375$. The anti-windup limit of 2.7us clamp time. It can be seen that the rise time for the output current is 178ms and the rise time for the output voltage is 668ms. Because the bulb is a non-linear resistor that changes with temperature, the output voltage and current response times are not the same. Figure 4.24 shows the fall time for the output current and voltage, and the response times are 424ms and 570ms, respectively.

![Figure 4.23. 30V - 220V Step Response, $K_p=1$, $K_i=375$, $tc_{min}=2.7us$ showing (a) Rise Time of Bulb Current and (b) Rise Time of Bulb Voltage.](image)
Figure 4.24. 220V - 30V Step Response, $K_p=1$, $K_i=375$, $t_{c\_min}=2.7\mu s$ showing (a) Fall Time of Bulb Current and (b) Fall Time of Bulb Voltage.

The output current step response is largely dominated by the AC processing pickup controller because of its controlled current source property. The slow response time for the output voltage is caused by the low dynamic resistance (or $dR/dt$) of the bulb which significantly slows down the transient response. The low dynamic resistance of the bulb adds an extra 490ms (or more than 200%) to the response time. In order to overcome this issue, more current can be driven into the light so that it heats up more quickly. Figure 4.25 shows the step response of the output voltage from 30V – 220V for a controller with similar PI gains but reduced anti-windup limit of 1.5us. The minimum clamp is reduced to increase the output current during transient start up. Because there is a PI controller, the steady state clamp time will not go below 2.7$\mu$s at 220V, hence the lamp is not overdriven under these conditions. The current rise time is 160ms which is similar to the system before. However, the voltage rise time is reduced to 530ms (or by 130ms) by using higher peak currents to drive the bulb.

Figure 4.25. 30V - 220V Step Response, $K_p=1$, $K_i=375$, $t_{c\_min}=1.5\mu s$ showing (a) Rise Time of Bulb Current and (b) Rise Time of Bulb Voltage.
Lastly, the PI gains are increased for the pickup to yield a faster transient response. The $K_p$ and $K_I$ gains are increased to 5 and 750, respectively. Figure 4.26 shows a step response of this system. It can be seen that both the current rise time and voltage rise time are much shorter – 92ms and 402ms, respectively. Figure 4.27 shows the step response from 220V – 30V. The fall times for both the current and voltage are also shorter as expected. However, even if the gain is increased significantly, the output voltage response does not have a considerable change as it is dominated by the low dynamic resistance of the bulb. The best method for providing high currents during start up to heat up the bulb quickly is to use a controllable voltage source. A detailed description of realising an AC controllable voltage source is outlined in chapter 6 using a series AC processing pickup.

![Figure 4.26](image1.png)

Figure 4.26. 30V - 220V Step Response, $K_p=5$, $K_I=750$, $t_{c\_min}=1.5us$ showing (a) Rise Time of Bulb Current and (b) Rise Time of Bulb Voltage.

![Figure 4.27](image2.png)

Figure 4.27. 220V - 30V Step Response, $K_p=5$, $K_I=750$, $t_{c\_min}=1.5us$ showing (a) Fall Time of Bulb Current and (b) Fall Time of Bulb Voltage.
4.5 Summary

In this chapter, some practical implementation aspects of the AC processing pickups are discussed. Firstly, a fourth order polynomial approximation is used to produce a look-up table so that the controller can be directly implemented in a microcontroller. Secondly, series and parallel connection of switching devices are outlined in order to achieve higher ratings for practical AC processing pickups for high power applications. Thirdly, a new PWM control strategy is outlined which can significantly decrease the losses in the AC switch of a parallel AC processing pickup, enabling reduced heatsink and pickup size. Lastly, a 1.2kW lighting system is built to illustrate the operation of a closed-loop parallel AC processing pickup controller.
In the previous chapters, the AC processing pickup has been investigated and designed to produce a high frequency AC output for driving suitable AC loads such as lights. In this chapter, a new design which incorporates an AC processing pickup with cascaded rectifier is investigated to produce a controlled DC output. Although this circuit performs a similar task (producing a controllable DC) to that of a traditional pickup that utilizes decoupling (or shorting) control [12] - it has several advantages. If the decoupling controller, shown in Figure 5.1, is used in applications with large variations in coupling, an inefficient method of overrating circuit devices is required. If the coupling can change over a wide range, the short circuit current also changes significantly. Components like the DC inductor, the shorting switch and the rectifier, shown in Figure 5.1, have to be rated for the highest short circuit current condition while normally operating at lower currents. Moreover, since the short circuit current of the parallel resonant tank is constantly flowing through the switch, this causes constant on state losses in the switch which lowers efficiency. This is particularly evident in systems whose coupling coefficient can change by over 100%. To design such a system, the minimum coupling coefficient must be assumed in order to ensure that the minimum current is enough to satisfy the power transfer required. However, under maximum coupling conditions the short circuit current may be as much as 100% larger than that required which means all the components have to be chosen such that they will operate safely at 200% of “rated” current. This inevitably reduces efficiency and increases cost. On the other hand, the AC processing pickup directly controls the output current from the resonant circuit to the rectifier, consequently, no overrating is required in applications where coupling varies. This is particularly useful in wireless powered battery charging applications such as required in EV’s or biomedical implants.
5.1 Circuit Operation

The AC processing pickup cascaded with a rectifier is shown in Figure 5.2. The DC inductor and capacitor acts as a filter network to produce a DC output. An additional purpose of the DC inductor is to provide continuous current flow through the rectifier.

An alternative circuit topology that is more practical or has fewer components is shown in Figure 5.3. The rectification is performed in this new circuit structure using the body diodes in the back to back MOSFET’s forming the AC switch. This has the advantage of eliminating the need for two additional diodes in the rectifier. In addition, this new circuit has a common ground reference on both the AC and DC side. This makes measurement of any voltage signal on the AC or DC side easier as the measurement circuitry does not need to be isolated.
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To understand the operation of the rectifier with the AC processing pickup operation, an equivalent load model is developed.

### 5.2 Load Model of Rectifier

To model the rectifier exactly without any approximations is very difficult as both the AC processing pickup and the rectifier are non-linear. The analytic equations that describe the AC processing pickup alone are already in the form of transcendental equations which can only be computed using numeric solvers. The addition of the rectifier makes the computation even more difficult to analyze. Here a simplified approximate analysis procedure is undertaken to derive an approximate load model for the rectifier such that the normalised graphs in chapters 3-4 can be reused.

The AC voltage and current at the input of the rectifier are shown in Figure 5.4. Here, it is assumed that the DC inductor current is continuous and reasonably large so that the inductor current stays constant with respect to the high frequency waveforms. Consequently, the input current ($i_2$) is approximately rectangular with a magnitude of the DC load current ($i_{Ldc}$). The fundamental component of this rectangular current is:

$$ i_2 = \frac{2\sqrt{2}I_{Ldc}}{\pi} $$

(5-1)

It is assumed that the voltage waveform can be approximated by half sinusoids with a section of the waveform clamped. This approximation is good for pickups with relatively high $Q_2$ (3-10). If the inductor current is continuous, the average voltage across the inductor must be zero and the relationship between the peak of the resonant capacitor voltage and the DC output voltage must be:

$$ V_{dc} = 2f \int_{V_c}^{1/2f} \hat{V}_c \sin(\omega t) dt $$

(5-2)

Solving (5-2) results in:
Chapter 5. The AC Processing Pickup with DC Output.

$$\vec{V}_{c2} = \frac{\pi}{2(1 - 2t_c/T)} V_{dc} \quad (5-3)$$

The fundamental component of the capacitor voltage transfers the power and is given by:

$$(V_{c2})_l = \int_{t_c}^{T/2} \vec{V}_{c2} \sin \left( \frac{2\pi}{1/f - 2t_c} - \frac{2\pi t_c}{1/f - 2t_c} \right) \sin(\omega t) dt \quad (5-4)$$

Solving (5-4) and substituting (5-3) into the solution results in:

$$(V_{c2})_l = \frac{\sin \left( \frac{2\pi f t_c}{1/f - t_c} \right)}{2\pi f^2 \left( \frac{1}{f} - t_c \right) t_c} V_{dc} \quad (5-5)$$

The equivalent AC resistance can be computed using (5-1) and (5-5), and is given by:

$$R_{ac} = \frac{(V_{c2})_l}{I_2} = \frac{\pi \sin (\omega t_c)}{2\sqrt{2} \left( 1 - ft_c \right) \omega t_c} R_{dc} \quad (5-6)$$

![Figure 5.4. Voltage and Current Waveforms at Input of Rectifier.](image)

Equation (5-6) shows that the equivalent AC resistance decreases with clamp time. Using this developed load model for the rectifier, many of the results derived in the chapter 3 and 4 can be directly used to analyse this rectifier load. To illustrate the accuracy of this derived model, a comparison of the simulation results of a rectifier load and the normalised graphs in
Chapter 5. The AC Processing Pickup with DC Output.

chapter 4 are shown in Figure 5.5 to Figure 5.7. The red traces in each case indicate the normalised graphs of Figure 4.7. The blue traces are the actual simulated results of the rectifier circuit with the scaling factor (5-6) included. Figure 5.5 shows the comparison of output voltage. It can be seen that the accuracy between the two plots is remarkably similar. In addition, the error between the two is not more than 3%. Figure 5.6 shows the comparison of the predicted output current which again is well matched. Notably, the approximation is better at high $Q_2$, whereas the graphs at lower $Q_2$ values have higher error. Figure 5.7 shows the comparison of the reflected impedance. The correlation between the normalised reflected resistance is reasonably accurate, however the correlation between the reflected reactance is poorer. The rectifier circuit reflect greater capacitive VAR's. This is because the rectifier adds harmonics to the resonant capacitor voltage such that the fundamental capacitor voltage gets further delayed against the current. This is equivalent to a small capacitive component $C_{ac}$ in series with an equivalent load resistor $R_{ac}$. This capacitive component seen by the open circuit voltage source, shown in Figure 5.8, would be inductive, resulting in higher capacitive loading once it is reflected onto the primary track. With this said, the resistive component of the reflected impedance stays largely the same, showing that the equivalent resistance model correlates well. The capacitive VAR's at the maximum power condition is larger, however this is not a big issue for the primary power supply as it must be rated to source the peak VAR's which is only marginally increased.
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Figure 5.5. Comparison of the Output Voltage of the Simulated Results of Rectifier Load and the Normalised Graphs using AC Resistor.

Figure 5.6. Comparison of the Output Current of the Simulated Results of Rectifier Load and the Normalised Graphs using AC Resistor.
5.3 Load Model with Output Voltage Regulation

In section 5.2, the load model of the rectifier was analyzed with a DC resistor as the load. Although this load model can be used in many applications, it cannot be directly used for battery type loads that exhibit a voltage source behavior. In this section, a procedure for modeling a battery type load is presented. Since the voltage across the load is constant, the
current flowing into the load is the only controllable parameter of interest.

### 5.3.1 Output Current Characteristics

To determine the output current characteristics, the output voltage characteristic must be used to determine the $Q_2$ of the circuit. Figure 5.9 shows the normalized output voltage vs. the quality factor. It should be noted that the output voltage $V_{dc}$ will have a ratio given by (5-5) with respect to the fundamental component of the resonant capacitor voltage, because the normalized graphs are all computed with respect to AC values rather than DC values. As such, a quality factor of 10 does not directly refer to a normalized output voltage boost of 10 times. Using Figure 5.9, $Q_{20}$ can be determined given any output voltage and clamp phase condition. Next, this $Q_{20}$ quality factor and clamp phase value can be used to evaluate the output current from Figure 5.10. Note this graph shows that the DC output current will have a ratio given by (5-1) with respect to the AC output current. Using these two figures, any output current condition can be obtained using measurements of the normalized output voltage and any known clamp phase condition with an error less than 5%.

![Figure 5.9. Output Voltage Characteristic of AC Processing Pickup with Rectifier Load.](image-url)
5.3.2 Peak Switch Voltage Characteristics

An important aspect in practice is to keep the peak voltages across both the AC switch and the rectifier diode constrained during operation. The peak voltage that appears across the switch is directly proportional to $V_{dc}$ if the DC inductor current is continuous. However, it is also proportional to the clamp phase of the circuit. Assuming that the inductor current is continuous during operation, the peak switch voltage can be calculated using (5-3). A graph of the peak switch voltage vs. the clamp phase is shown in Figure 5.11. The peak switch voltage rapidly increases without sensible limit as the clamp phase increases to 180º. This is a practical concern as switches with infinite ratings cannot be physically realised. Consequently, a solution for reducing the peak switch voltage by allowing discontinuous currents through the DC inductor is proposed.
Discontinuous current operation has previously been avoided in IPT pickup rectifiers as the maximum power the pickup can deliver reduces [124]. In addition, it causes higher ripple currents in the rectifier. Despite these disadvantages, one advantage of discontinuous operation is that the non-linearities introduced by the rectifier lower the peak switch voltage. This advantage is much more important for AC processing pickups with rectifiers as their peak capacitor voltage increases (ideally to infinity) as clamp time increases. This is a practical problem that traditional IPT pickups do not have. Consequently, discontinuous current operation can be useful here.

It is important to calculate the minimum DC inductance value necessary to keep the current continuous because forcing continuous current conduction is essential at zero clamp time to ensure that the peak power of the IPT system is not compromised. Although the minimum DC inductance has been analyzed previously [124], its derivation will be briefly mentioned for completeness. Figure 5.12 shows the circuit operation of the rectifier operating at the minimum DC inductance to force continuous conduction in the inductor. The following differential equation can be written as,

$$L_{dc} \frac{dI_{dc}}{dt} = V_{ac} \left| \sin(\omega t) \right| - V_{dc}$$  \hspace{1cm} (5-7)
Chapter 5. The AC Processing Pickup with DC Output.

Solving (5-7) results in,

$$i_{L_{dc}} = \frac{-V_{ac} \cos(\omega t) - V_{dc, ot} + \sqrt{V_{ac}^2 - V_{dc}^2} + V_{dc} \arcsin \left( \frac{V_{dc}}{V_{ac}} \right)}{\omega L_{dc}}$$

(5-8)

Since the average of the DC inductor current in (5-8) must be equal to the output current to the load, the minimum DC inductance can be given by,

$$L_{dc, min} = 0.3307 \frac{R_{dc}}{\omega}$$

(5-9)

Figure 5.12. Operating Waveforms of Rectifier.

In this work, the minimum DC inductance is defined as that required to maintain continuous current flow when the clamp phase is zero. Any clamp phase higher than zero will result in discontinuous currents and a decrease in power delivery. However, the reduction in power delivery is not a problem as the primary reason for increasing the clamp time is to reduce the output current below nominal. The only practical difference is that the output current now decreases at a faster rate than is normally expected when operating with an ideal rectifier and continuous current flow. Figure 5.13 shows the peak switch voltage rating vs. the clamp phase for different DC inductance values as a ratio of the minimum inductance. It can be seen that using an inductor value set to the minimum DC inductance allowable, significantly reduces the peak switch voltage rating compared with the earlier ideal operating
condition. It should also be noted that choosing an $L_{dc}=10L_{dc\_min}$ only marginally increases the peak voltage compared to $L_{dc}=L_{dc\_min}$. Using this knowledge, lower switch ratings can be realised when using the AC processing pickup to output controlled DC.

![Figure 5.13. Peak Switch Voltage vs. Clamp Time.](image)

**5.4 Practical EV Charging System Design**

For IPT systems used in EV charging applications, a power flow controller is necessary to control the charging voltage and current to the on-board battery. In some stringent requirements, a typical output voltage error of less than 1% is desired. However, there are many factors like variations in load, coupling condition or system parameters in an IPT system that can cause the power transfer or the output voltage to vary [92]. Almost all IPT power flow controllers handle variations in load to some degree, however, not all of them can handle large variations in either coupling or component values. One example of this is the decoupling/shorting controller used in materials handling applications [12].

To overcome these variations, an AC processing pickup is used to regulate the output voltage to the load. A brief description of how the AC controller can achieve wide load control is described below.

- 120 -
5.4.1 The AC Processing Equivalent Circuit

From the analysis in chapter 3, equation (3-14) demonstrates that the AC processing controller can be modeled as a controller that varies the pickup’s equivalent tuning capacitance from its existing value to a much larger value. This behavior can be redrawn in its Norton equivalent circuit as shown in Figure 5.14. This Norton equivalent circuit, has a fully tuned parallel resonant tank (consisting of $L_2$ and $C_2$) which reduces to an infinite impedance and can be neglected. By adjusting the phase delay $\phi$, the impedance of the equivalent capacitor $C_e$ in the current divider is controlled. As the impedance of the capacitance branch changes, the current delivered to the load can be directly controlled in the current divider. The magnitude of the output voltage delivered to the load is then given by:

$$
|V'| = \frac{\omega M I Q_1}{\sqrt{1 + \omega^2 R_2^2 (C_e - C_2)^2}}
$$

(5-10)

Figure 5.14. Equivalent Circuit of AC Processing Pickup.

Using the equation in (5-10), variations of system parameters can be quantified. Each of these is discussed below.

5.4.2 Variations in an EV Charging System

To investigate variations in the IPT system, some definitions relating to the nominal operating condition is defined in Table 5.1. The nominal conditions here, refers to the ideal operating conditions for the IPT system without any variations.
Table 5.1. Definitions of Nominal Conditions.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{20}$</td>
<td>Nominal Pickup Inductance</td>
</tr>
<tr>
<td>$C_{20}$</td>
<td>Nominal Tuning Capacitance</td>
</tr>
<tr>
<td>$k_0$</td>
<td>Nominal Coupling Condition</td>
</tr>
</tbody>
</table>

5.4.2.1 Variations in Pickup Inductance and Tuning Capacitance

Any variation in the value of the tuning capacitor or the pickup inductor on the secondary is classified as a tuning variation. Capacitors usually have tolerances because of manufacturing precision or component aging. The pickup inductor on the other hand varies with the alignment of the charging pads [125] as they usually have ferrite to localize the magnetic field. From Figure 5.15, a change in tuning capacitance will change the reactance of the equivalent capacitance in the AC processing pickup, and thereby change the controllability of the pickup. Figure 5.15 shows a few trends of the output voltage against different tuning capacitances for a $Q_{20}=5$ and a $k=100\%k_0$. It can be seen that if the tuning capacitor is lower than nominal, the full control range is still obtained by increasing the phase delay angle to reach the maximum. However, when the actual tuning capacitance is higher than nominal, the full control range (or peak voltage obtained) is reduced and the normalized output voltage is lower than 5. This means that the full control of output voltage from 0-5 can no longer be achieved. For a change in inductance, a similar effect is observed. However, a variation in the pickup inductance changes the short circuit current, and thereby changes the normalized output voltage. Despite this change, it can be seen that a smaller pickup inductance than nominal would retain the full control range of 0-5.
Figure 5.15. Normalised Output Voltage vs. Normalised Capacitance for Different Tuning Capacitance.

Figure 5.16. Normalised Output Voltage vs. Normalised Capacitance for Different Pickup Inductance.

5.4.2.2 Variations in Load Resistance

Figure 5.17 shows a plot when the load resistance (or $Q_{20}$) of the circuit varies from a $Q_{20}=3-7$ while $C_2=C_{20}$, $L_2=L_{20}$ and $k=100\%k_0$. Notably, the output voltage varies with variations in the load resistance. By adjusting the phase delay angle or the equivalent capacitor, the output voltage can be varied anywhere from the maximum nominal tuned value.
down to zero. As a result, the controller can only reduce output voltage from the maximum tuned value to zero. However, a circuit cannot have a maximum normalized output voltage higher than the operating \(Q_{20}\) of the circuit by operating in this manner. This denotes that there must be an \(R_{\text{min}}\) condition so that the output load cannot draw more power than its maximum power condition when the output voltage is regulated at its nominal value.

![Figure 5.17. Normalised Output Voltage vs. Normalised Capacitance for Different \(Q_z\) values.](image)

### 5.4.2.3 Variations in Coupling Condition

From Figure 5.18, the output voltage is proportional to the coupling condition of the system. Since the coupling coefficient \(k\) is also directly proportional to the mutual inductance, the output voltage is proportional to the coupling coefficient. Figure 5.18 shows a plot for a range of coupling coefficients when \(C_z=C_{20}\), \(L_z=L_{20}\) and \(Q_{20}=5\). It can be seen as the coupling varies, the output voltage varies by the same proportion. By controlling the equivalent capacitor, the output voltage can be controlled at a fixed reference. For example, if the desired reference is a normalized output voltage of 5, then the output voltage can be regulated, except for the one when \(k<100\%k_0\), as the coupling coefficient or short circuit current is too low to maintain the output voltage condition.
5.4.3 Steady State Design

The objective of any practical design is to ensure that practical system variations are considered. In consequence it is important to define some key specifications. Here \((V_c)_{\text{min}}\) is referred to as the minimum capacitor voltage the pickup must supply to the load and indirectly defines the power required to be transferred by the IPT system.

A deviation in pickup inductance will change the natural resonance, but also change the output voltage given this is inversely proportional to \(Q_{20}\) of the circuit. Similar to the above, a smaller inductance value does not lose control range, as such the pickup inductance must be chosen using:

\[
L_2 = L_{20} - L_{2e} \tag{5-11}
\]

where \(L_{20}\) and \(L_{2e}\) are the nominal inductance and the error or deviation in pickup inductance values, respectively. Note that (5-11) indicates that an inductance smaller \(L_{20}\) should be used. Since the inductance also changes \(Q_{20}\), an output voltage gain factor must be included, defined here as:

\[
g_L = \frac{1}{(1 - L_{2e}/L_{20})} \tag{5-12}
\]
Any variation in tuning capacitance will change the control range of the AC processing controller from its existing value to infinity. Since any tolerance in tuning capacitor directly changes the equivalent capacitor value, the error can be compensated by choosing a smaller capacitor:

\[ C_2 = C_{20} - C_{2e} \]  \hspace{1cm} (5-13)

where \( C_{20} \) and \( C_{2e} \) are the nominal and error in the tuning capacitance, respectively.

It can be seen that the pickup inductance and tuning capacitance can both be chosen to be smaller than their nominal values to allow the AC processing pickup to maintain its control range. However, it is complicated to look at each one independently. An alternative and easier approach is to simply normalize each of these variations against one variable, namely the tuning capacitance. Thus, in order to select a tuning capacitance that is smaller than nominal, the following equations are used:

\[ C_2 = \alpha C_{20} \]  \hspace{1cm} (5-14)

\[ \alpha = \left(1 - \frac{L_{2e}}{L_{20}}\right) \left(1 - \frac{C_{2e}}{C_{20}}\right) \]  \hspace{1cm} (5-15)

where \( \alpha \) can change from 0 to 1, and as shown in (5-15), it is desirable to keep it as low as possible. However, in reality there are several practical costs to choosing a lower \( \alpha \) value. Figure 5.19 shows the maximum output voltage obtainable under any \( \phi \) from the AC processing pickup against \( \alpha \) values for different \( Q_{20} \). This is a normalized plot which captures the non-linearities and harmonics in the AC processing pickup and is obtained by solving the equations relating to the pickup tuning circuit using the exact analysis technique outlined in chapter 3. The voltage gain is given by a complex non-linear function as:

\[ g_c = f(Q_{20}, \alpha) \]  \hspace{1cm} (5-16)

where the \( g_c \) factor is given by the normalized output voltage divided by \( Q_{20} \) as shown in
Figure 5.19. When $\alpha=1$, the maximum voltage obtainable is $V_{oc}Q_2$, which conforms to traditional parallel LC resonant tank theory [12]. However, as the $\alpha$ value decreases, the maximum obtainable voltage also decreases which reduces the power transfer capacity of the pickup. In addition, as shown in Figure 5.20, the semiconductor switch rating also increases as lower $\alpha$ values are used. Thus with lower $\alpha$ values, the power transferred within the IPT system compared to the device rating reduces thereby increasing the cost per watt of the IPT system. Because of this, the $\alpha$ value should be chosen as high as possible, while also satisfying all the component tolerances.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure519.png}
\caption{Maximum Output Voltage vs. $\alpha$.}
\end{figure}
Chapter 5. The AC Processing Pickup with DC Output.

Figure 5.20. Maximum Peak Switch Voltage vs. $\alpha$.

By noting that there is a power reduction resulting from the complex non-linear behavior of the AC processing pickup, the following relationship must be satisfied to meet the necessary design specifications:

$$\left( V_{c} \right)_{\text{min}} = \omega M I_{1} \left( Q_{2} \right)_{\text{min}} g_{L} g_{C}$$  \hspace{1cm} (5-17)

Rearranging (5-17), the minimum open circuit voltage of the pickup to ensure power transfer is:

$$\left( V_{oc} \right)_{\text{min}} \geq \frac{2\sqrt{2} \left( V_{dc} \right)_{\text{min}} \omega L_{0}}{\pi \left( R_{dc} \right)_{\text{min}} B_{L} B_{C}}$$  \hspace{1cm} (5-18)

Using this, the designer can derive the minimum coupling required in the chosen IPT system.

5.4.4 Design Example and Experimental Results

A design example for a 1kW AC processing pickup is presented in Figure 5.21. Here, a block diagram is used to illustrate the AC processing pickup configuration where the switch current detection method outlined in chapter 4 is used. The microcontroller measures the output voltage and adjusts the phase of the switch gate drive waveforms accordingly. Some
key specifications are listed on the left side of Table 5.2 and the designed component values are presented on the right. The pickup must regulate the output voltage $V_{dc}$ to a nominal 200V when the load varies from 40Ω to open circuit. In addition, the coupling coefficient can change by more than 100% when the height between the primary and secondary pads changes from 45mm to 90mm which is the nominal operating range for the chosen pad [125]. The primary and secondary IPT pad has a circular structure shown in Figure 5.22. Dimensions of both pads are exactly identical and each pad has 8 pieces of ferrite with 12 turns. The variation in coupling with respect to height is shown in Figure 5.23. It can be seen that the coupling factor changes from 0.16 – 0.37 as the distance between them varies from 45mm - 90mm. In addition, the variation in the inductance of the pads is shown in Figure 5.24. The inductance of the pickup is 72.4uH and its value changes by 2.7% as the pad alignment changes. The $\alpha$ value calculated from the specifications using (5-15) is 0.87, however the exact capacitance value of 206.5nF was not available. To be conservative, an $\alpha$ value of 0.81 was chosen which gives a capacitance value of 192nF. The minimum $V_{oc}$, calculated using (5-18), is 77.12V and the measured value is 78V when $k=0.16$. As discussed in section 5.3.2 DC inductance is chosen to be 10 times higher than the minimum inductance of (5-9), while still maintaining appropriate switch ratings.

![Figure 5.21. Block Diagram of the AC Processing Pickup.](image-url)
Chapter 5. The AC Processing Pickup with DC Output.

Table 5.2. Steady State Design Parameters.

<table>
<thead>
<tr>
<th>Spec. Parameters</th>
<th>Values</th>
<th>Design Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k$</td>
<td>$0.16 - 0.37$</td>
<td>$(V_{oc})_{min}$</td>
<td>78V</td>
</tr>
<tr>
<td>$P_{out}$</td>
<td>1kW</td>
<td>$(I_{sc})_{min}$</td>
<td>4.5A</td>
</tr>
<tr>
<td>$V_{dc}$</td>
<td>200V</td>
<td>$\alpha$</td>
<td>0.87</td>
</tr>
<tr>
<td>$R_{dc}$</td>
<td>$40\Omega - \infty \Omega$</td>
<td>$g_r$</td>
<td>1.03</td>
</tr>
<tr>
<td>$I_1$</td>
<td>29.2A</td>
<td>$g_c$</td>
<td>0.99</td>
</tr>
<tr>
<td>$f$</td>
<td>38.4kHz</td>
<td>$C_2$</td>
<td>192nF</td>
</tr>
<tr>
<td>$L_0$</td>
<td>72.4uH</td>
<td>$L_{dc}$</td>
<td>540uF</td>
</tr>
<tr>
<td>$C_0$</td>
<td>237.3nF</td>
<td>$C_{dc}$</td>
<td>1mF</td>
</tr>
<tr>
<td>$L_{2e}/L_0$</td>
<td>0.02695</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$C_{2e}/C_0$</td>
<td>0.1</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

All dimensions in mm

A 30 Ferrite Width
B 118 Ferrite Length
C 10 Ring Thickness
D 238 Coil Diameter
E 420 Pad Diameter
Pad Thickness 25
Coil 12 Turns

Figure 5.22. Charger Pad Layout.

![Figure 5.22. Charger Pad Layout.](image)

Figure 5.23. Coupling Factor vs. Distance Between Pads.

![Figure 5.23. Coupling Factor vs. Distance Between Pads.](image)
Chapter 5. The AC Processing Pickup with DC Output.

Figure 5.24. Pickup Inductance vs. Distance Between Pads.

For the practical system, the output voltage vs. clamp time is shown in Figure 5.25. The blue trends shows the output voltages when operating at a distance $d=90\text{mm}$ with a coupling coefficient $k=0.16$. Similarly, the red trends shows the output voltages when operating at a coupling coefficient $k=0.37$ with an air gap distance $d=45\text{mm}$. As shown, the majority of the output voltage range can be achieved using a clamp time range of between 1-6 $\mu\text{s}$. Since the theoretical control range for this 38.4kHz IPT system would be 0-13us, this control range is feasible and not too sensitive. To regulate the output voltage at 200V, the controller is required to measure the output voltage and vary the clamp time accordingly.

Figure 5.25. Output Voltage Characteristics for Open Loop Controller.
A closed loop system was built based on the block diagram shown in Figure 5.21. The output voltage is regulated to a desired reference using a simple PI controller to adjust the clamp time as shown in Figure 5.26. The reference voltage can be set by a PC using serial communication. The capacitor voltage phase is measured using a detection block to compute when the current starts to flow through the AC switch. The phase modulator block controls the phase between the output gate drive signals with respect to the phase of the capacitor voltage. Ideally this system provides a constant output voltage irrespective of coupling and load conditions. Since the change in pad inductance is inherently linked with coupling, only coupling variations are of interest as they include inductance variations.

**Figure 5.26. Diagram of Controller.**

Figure 5.27 shows the circuit waveforms when the output voltage is regulated to 200V at $k=0.16$ (or $d=90\text{mm}$) for rated, 50% and 10% load, respectively. The waveforms are the resonant capacitor voltage, inductor current, DC inductor current and synchronization signal from the switch current detection block. Note that, the switch current detection inverts the logic which is inverted again at the output of the gate driver. Similarly, Figure 5.28 shows the circuit waveforms when the output voltage is regulated to 200V at $k=0.37$ (or $d=45\text{mm}$) for rated, 50% and 10% load, respectively. For the more tightly coupled system of $k=0.37$, the resonant inductor current is higher as the short circuit current in the pickup inductor is higher. In addition, the peak capacitor (or switch) voltage is also higher. This is because a higher
clamp time is needed for the higher coupling condition to regulate the output at 200V. The measured result matches with the predicted results of Figure 5.25. Consequently, a higher clamp time for the same regulated output voltage results in higher peak voltages across the switches as identified in section 5.3.2. However, the peak switch voltage does not exceed 450V which is half of the ratings of the physical devices chosen.
Figure 5.27. Measured Waveforms at 200V output with $k=0.16$ for (a) Rated Load, (b) 50% Load and (c) 10% Load.
Chapter 5. The AC Processing Pickup with DC Output.

Figure 5.28. Measured Waveforms at 200V output with k=0.37 for (a) Rated Load, (b) 50% Load and (c) 10% Load.
Chapter 5. The AC Processing Pickup with DC Output.

The measured output voltage under different operating conditions is shown in Figure 5.29. Here, the circuit is tested over a wide range of coupling and load conditions. It can be seen that when the coupling and load resistance are higher than the designed minimum values specified (i.e. \( k > 0.16 \) and \( R_L > 40\Omega \)), the output voltage can be regulated at 200V. However, if a higher load is used (i.e. \( R_L < 40\Omega \)), the output voltage drops below 200V as the minimum clamp time will not be able to produce enough output current to sustain the voltage. Similarly, when \( k < 0.14 \), the output voltage also drops below 200V.

![Figure 5.29. Output Voltage under Different Circuit Conditions.](image)

The measured efficiency of only the secondary AC processing pickup is shown in Figure 5.30. This measurement includes all the losses on the secondary pickup and neglects both the losses in the primary converter and the pickup pads. The efficiency of the secondary pickup electronics is as high as 97% when outputting 1kW of controlled output power. However, it decreases at lighter loads. It should be noted that there is a point where the efficiency drops to a local minimum. This is mainly because the conduction losses in the AC switch are increased as the RMS current in these switches gets higher with increasing clamp time. At much lighter loads, the switch loss decreases because the overall resonant current in the parallel LC network is much lower.
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The DC-DC efficiency for the IPT system was measured and shown in Figure 5.31. The efficiency is taken by measuring and comparing the DC input power to the primary LCL power supply and the DC output power to the load. The efficiency drops off at higher coupling conditions as the AC processing pickup reflects higher reactive VAR’s back into the primary converter causing more conduction losses in both the LCL network and the H bridge.

Figure 5.31. Efficiency of IPT System.
5.5 Conclusions

This chapter presented a study and a method of design when a rectifier is cascaded at the output of the conventional parallel AC processing pickup introduced in chapters 3 and 4. Firstly, an equivalent resistive model of the rectifier loading condition is developed. This model was shown to be accurate when comparison results of the normalized graphs closely correlates with an accurate circuit simulation. In consequence the normalized graphs illustrated in the chapter 4 can be directly applied to the design of an AC processing pickup with a rectifier. Secondly, a loading model is developed for a system that has a regulated output voltage. The problem of excessive high peak voltage across the AC switches was solved by deliberately using a lower DC inductance value. Finally, a design example of a 1kW EV charging system using an AC processing pickup with rectifier was constructed and shown to operate as expected with a peak measured efficiency of 97%.
6 A Series AC Processing Pickup

In the previous chapter, a parallel AC processing pickup was proposed. These circuits were the first type of AC processing circuits developed specifically for lighting applications although recently they have been adopted for Electric Vehicle recharging as discussed in chapter 5. However, there is another class of pickups, similar to the parallel AC processing pickup which is also suitable for various applications. This is introduced here as a series AC processing pickup.

The series AC processing pickup is similar to its parallel AC processing controller and most of its properties are duals or opposites. As expected it operates as a controllable voltage source and similar to other series IPT pickup designs [101], it can be very efficient when powering high current low-voltage loads. This chapter introduces the use of an AC switch operating under Zero-Current-Switching (ZCS) conditions to regulate the output voltage directly for a series AC processing pickup. This topology can produce a high frequency controlled AC voltage source which is ideal for incandescent lamps where the resistance rises as it warms up. When a rectifier is added to the AC output, the circuit can control the output DC voltage. This modified topology eliminates the need for an extra buck converter stage, improves efficiency and reduces cost. In addition, the circuit has the ability to control and eliminate the transient inrush current during start up. Preliminary experimental results show that the pickup regulator operates with high efficiency.

Figure 6.1 shows a traditional series tuned pickup with a buck converter output stage. The pickup stage comprises a series resonant tank, a bridge rectifier and a DC filter capacitor. The buck converter stage includes the switch, the filter inductor and capacitor, and the flywheel diode. The buck converter is used to regulate the output voltage to a desired voltage equal to or less than the rectified open circuit voltage of the pickup. It also protects the circuit if the load is a short circuit. Although there are many advantages to the series compensated pickup with a buck converter, it is not ideal because of the large number of components required in
the circuit. In addition, the hard switching operations in the buck converter reduce efficiency and generate more EMI.

![Figure 6.1. A Series Compensated Pickup with a Buck Converter.](image)

The new AC processing pickup proposed in this chapter can also generate a controlled DC voltage, like the series compensated pickup with a buck converter, however the overall circuit requires less components and the efficiency can be higher than that of the traditional circuit as it has soft switching operation, with only one conversion stage.

### 6.1 Circuit Operation

The series AC processing pickup is shown in Figure 6.2 with an AC output voltage $V_{R2}$. Capacitor $C_2$ is tuned to inductor $L_2$ at the frequency of the primary track current $I_1$ to form a series resonant tank. For simplicity, switch $S_1$ is drawn as an ideal AC switch and is the basis for controlling the output voltage. The open circuit voltage source $V_{oc}$ represents the induced voltage of the pickup and is given by:

$$V_{oc} = j\omega MI_1$$  \hspace{1cm} (6-1)

where $\omega$, $M$ and $I_1$ are the primary power supply angular frequency, the pickup mutual inductance and the track primary current, respectively.
Chapter 6. A Series AC Processing Pickup.

![Diagram of a Series Tuned AC Processing Pickup](image)

**Figure 6.2.** The Series Tuned AC Processing Pickup.

By controlling the switching action of the AC switch, a phase shift $\phi$ will be generated between the open circuit voltage and the inductor current waveform as shown in Figure 6.3. Since the power sourced from the $V_{oc}$ source is directly proportional to the phase shift $\phi$ between the voltage and the current, the output power will be directly controlled if the conversion network is assumed to have negligible losses. Hence, by controlling the phase shift $\phi$, the output power is controlled.

![Waveform showing Two Operating States](image)

**Figure 6.3.** Waveform showing Two Operating States.

To illustrate the circuit’s operation, Figure 6.4 shows the operating waveforms of the series AC processing pickup. $V_{g1}$ is the PWM control signal which turns $S_1$ on and off. Consider the situation where $V_{g1}$ is controlled with a phase delay $\phi$ relative to the phase of $V_{oc}$ as shown in Figure 6.4. In Mode 1 ($M_1$, $0 < t \leq t_1$), $S_1$ is turned on and the capacitor $C_2$ resonates with pickup inductance $L_2$ like a series resonant tank. During this time, the inductor current reaches a peak value and then returns back to zero at which time $S_1$ is turned off and the circuit enters...
Mode 2 \((M_2, t_1 < t \leq t_2)\). In this mode, no current flows through any device and the inductor current is discontinuous for a phase known as the discontinuous phase \((\omega t_d)\). At the beginning of Mode 3 \((M_3, t_2 < t \leq t_3)\), \(S_1\) is turned back on. Similar to \(M_1\), the circuit operates like a series resonant tank and current flows into the load resistor. In Mode 4 \((M_4, t_3 < t \leq T)\), similar to \(M_2\), the resonant cycle is terminated and the inductor current is discontinuous. After this mode, the circuit returns back to \(M_1\), repeating the switching process.

![Figure 6.4. Operating Waveforms of Series AC Processing Pickup.](image)

The AC switch in the series AC processing pickup achieves ZCS conditions. From Figure
6.4, at $t_1$, the voltage across $S_1$ increases from zero to a positive voltage while the current through it is at zero. Because there is no current flow, ZCS is achieved at turn off. When $S_1$ is turned on at $t_2$, the pickup inductor in series with $S_1$ forces the current through it to increase slowly in the negative direction while the voltage across it decreases rapidly to zero. For most practical switches, the turn on is much faster than the rate of increase of the inductor current, as such the $\frac{di}{dt}$ through the switch is relatively small and a ZCS condition is obtained at turn on. In summary, if the timing of the gate drive signal for the AC switch is accurate, the AC processing pickup achieves ZCS conditions. In section 6.4.1, a more practical method of driving the AC switch which does not rely on accurate timing is described. The soft switching condition described above gives the pickup desirable characteristics such as low switching losses, low switching stress and reduced electromagnetic interference (EMI) levels.

A rectifier can also be cascaded onto the series AC processing pickup to produce a controlled DC output as shown in Figure 6.5. The small signal characteristic of the AC processing pickup is modeled and a phase shift generator can be used to control the phase of the inductor current with respect to the open circuit voltage. A Proportional Integral (PI) controller can be used in the inner current loop to regulate the average inductor current magnitude within a set limit. An outer voltage loop is used to regulate the output voltage to a reference using a PI controller as well. By using the AC switch before the rectifier, the transient startup inrush current can be fully controlled as the inductor current can be regulated. In addition, this topology also removes the need for a buck converter after the pickup as the output voltage can be controlled directly.

![Figure 6.5. A Series AC Processing Pickup with Rectifier.](image-url)
Chapter 6. A Series AC Processing Pickup.

6.2 Pickup Analysis

From the previous section, it can be seen that the phase shift between $V_{oc}$ and $I_L$ can be controlled by adjusting the phase delay $\phi$. In this section, an exact time domain analytical analysis approach by solving differential equations in each of the circuit’s operating states is presented. This procedure follows closely the work presented in [126]. With reference to Figure 6.3, the waveform can be separated into two operating states known as the resonant state and the discontinuous state. It is assumed that Capacitor $C_2$ and inductor $L_2$ are perfectly tuned forming a series resonant tank.

6.2.1 Resonant State

During the resonant state, the inductor current may be described as:

$$i_{Lr}(t) = \frac{-Q_{20}V_{oc}}{\omega L_2 \sin(\theta_1)} e^{-\sigma t} \sin(\omega_J t - \theta_1) + \frac{Q_{20}V_{oc}}{\omega L_2} \sin(\omega t + \phi)$$

(6-2)

where

$$Q_{20} = (\omega L_2) / R_2$$

(6-3)

$$\sigma = R_2 / (2L_2)$$

(6-4)

$$\omega_J = \omega \sqrt{1 - \frac{1}{4Q_{20}^2}}$$

(6-5)

$$\theta_1 = \tan^{-1}\left(-\frac{\omega_J V_{oc} \sin(\phi) / \omega}{V_c(0) - V_{oc} \sin(\phi) + Q_{20} V_{oc} \cos(\phi) + \sigma Q_{20} V_{oc} \sin(\phi) / \omega}\right)$$

(6-6)

$$\theta_c = \tan^{-1}\left(-\frac{\omega_J (V_c(0) + Q_{20} V_{oc} \cos(\phi)) / \sigma}{-V_c(0) - Q_{20} V_{oc} \cos(\phi) + \omega Q_{20} V_{oc} \sin(\phi) / \sigma}\right)$$

(6-7)

In a similar way, the complete solution to the capacitor voltage is:

$$V_{cr}(t) = -\frac{V_c(0) + Q_{20} V_{oc} \cos(\phi)}{\sin(\theta_c)} e^{-\sigma t} \sin(\omega_J t - \theta_c) - Q_{20} V_{oc} \cos(\omega t + \phi)$$

(6-8)
To find the time $t_r$ that the circuit stays in the resonant state, $i_{L}(t)=0$ can be substituted in (6-2).

### 6.2.2 Discontinuous State

During the discontinuous state ($t_d$), the series resonant circuit becomes an open circuit and the capacitor voltage remains constant while the inductor current is zero.

\[ V_{cd}(t)|_{t=t_1} = V_{c}(t_r) \]

\[ i_{LD}(t) = 0 \]  

(6-9)  

(6-10)

Because the resonant state and the discontinuous state are repeated each half cycle (with only a polarity change), the relationship $V_c(0)=-V_c(T/2)$ must hold. Hence, the capacitor voltage and inductor current are given by,

\[
V_c(t) = \begin{cases} 
V_{cr}(t) & 0 \leq t < t_1 \\
V_{cd}(t) & t_1 \leq t < t_2 \\
-V_{cr}(t) & t_2 \leq t < t_3 \\
-V_{cd}(t) & t_3 \leq t < T 
\end{cases} 
\]

(6-11)

\[
i_L(t) = \begin{cases} 
i_{LD}(t) & 0 \leq t < t_1 \\
i_{rd}(t) & t_1 \leq t < t_2 \\
-i_{cr}(t) & t_2 \leq t < t_3 \\
-i_{cd}(t) & t_3 \leq t < T 
\end{cases} 
\]

(6-12)

Fourier analysis can be performed on the inductor current waveform to compute the harmonics. The in-phase and quadrature components of both the fundamental and harmonics are given by:

\[
I_{L_{ew}} = \frac{2\omega}{\pi} \int_{0}^{\pi/\omega} i_L(t) \cos(n \omega t) \, dt 
\]

(6-13)

\[
I_{L_{eqn}} = \frac{2\omega}{\pi} \int_{0}^{\pi/\omega} i_L(t) \sin(n \omega t) \, dt 
\]

(6-14)

It is important to determine the amount of power sourced from the primary IPT power
supply for the pickup to operate. If harmonics are ignored, the real and reactive power sourced from the primary supply is given by:

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

where

\[ Z_t = \frac{-j\omega M^2 I_{q1} + jI_{f_q}}{I_{sc}} \]  

and \( I_{sc} \) is defined as the short circuit current of the pickup inductor.

### 6.2.3 Computation Routine

The above analytical analysis is very difficult as the solution of \( t_r \) and \( V_c(0) \) are governed by (6-11) and (6-12) with \( \theta_v \) and \( \theta_i \) as interim variables which are associated with the auxiliary equations (6-6) and (6-7). This is in the form of transcendental equations and it can only be solved using numeric solvers such as MATLAB or EXCEL. A computer program based on an iterative computation, shown in Figure 6.6, has been developed to undertake the analysis. The program starts by initializing the circuit parameters such as \( L_2, C_2 \) and \( R_2 \). The initial capacitor voltage is set as an initial condition. With \( V_c(0) \) known, \( t_r \) can be calculated by solving (6-9). With \( t_r \) known, the capacitor voltage at half a period can be calculated using (6-11). The next step is to check whether \( V_c(0) \) and \(-V_c(T/2)\) have converged to a given error (\( \varepsilon < 0.01\% \)). If the answer is YES, the program terminates when the correct solution is found. Otherwise the iteration repeats itself in the computation loop until a solution is found. The algorithm has proven to be both fast and robust.
6.2.4 Rectifier Load Modelling

The series AC processing pickup can output a controlled DC voltage by adding a rectifier with a large DC filter capacitor. The output voltage is maintained at a DC level with the high frequency AC component removed. As a result, the AC voltage at the input of the rectifier becomes a rectangular waveform with an ideal amplitude equal to the DC output voltage (assuming the two diode forward voltage drops are ignored). If only the fundamental component is modelled and the harmonic components are ignored, the RMS of the fundamental component of the rectangular voltage is given by:

\[
V_{R2} = \frac{2\sqrt{2}V_{dc}}{\pi}
\]  

(6-18)

If it is assumed that the input current to the rectifier can be approximated by half sinusoids with discontinuous sections in between, then the RMS of the fundamental component of the
AC current is related to the rectified DC current by:

\[ I_L = \frac{\pi \sin(\omega t_d) I_{dc}}{2\sqrt{2} (1 - t_d / T) \omega t_d} \]  

(6-19)

Therefore the equivalent AC load is:

\[ R_2 = \frac{8(1 - t_d / T) \omega t_d R_{dc}}{\pi^2 \sin(\omega t_d)} \]  

(6-20)

Note that, when \( t_d \) approaches zero, \( R_2 \) does not tend towards infinity, but rather \( 8R_{dc}/\pi^2 \).

The equivalent AC resistor was used in simulation and shown to give reasonably accurate results. This derived AC resistor can be used in (6-3) for the AC analysis described in section 6.2 to compute the operating waveforms of the series AC processing pickup with a rectifier load.

### 6.3 Pickup Characteristics

In this section, the operating characteristics of the series AC processing pickup are presented in comparison to the parallel AC processing pickup previously reported in [126]. The output current (or inductor current) characteristics of the series topology are shown in Figure 6.7(a) for different values of \( Q_{20} \) or load conditions. The normalized output current is defined as the ratio of the output current over the short circuit current. For the series AC processing pickup, the output current is dependent on both phase delay \( \phi \) and \( Q_{20} \). The normalized output current can be controlled from a maximum value (or \( Q_{20} \)) to zero as \( \phi \) changes for all load conditions. In comparison, for the parallel AC processing pickup mentioned in section 3, the output current is mostly dependent on \( \phi \) and marginally dependent on \( Q_{20} \).
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Figure 6.7. Normalized RMS Output Current vs. Phase Delay $\phi$ for (a) Series and (b) Parallel.

The output voltage-current characteristic is shown in Figure 6.8(a) for the series topology. The voltage source behaviour is demonstrated as the output voltage stays approximately constant for a given phase delay irrespective of output current as long as the output current is reasonably high. Similarly, the parallel AC processing pickup current-voltage characteristic, shown in Figure 6.8(b), appears as a controlled current source behaviour. In applications where high currents are required for a short amount of time, the voltage source characteristic is particularly useful. If a parallel topology is used in such applications, it is usually more costly and inefficient to overrate the inductor coil in the parallel AC processing pickup to meet this demand. As previously noted, the series AC processing pickup will give rise to significant advantages in the overall system, as the output voltage can be controlled to any value below the open circuit voltage without the need of an extra buck converter after the pickup (shown in Figure 6.1). This eliminates both the losses of an extra DC-DC conversion stage and the need for a costly DC inductor used in the buck converter.
Figure 6.8. Pickup Output Characteristics showing (a) Series and (b) Parallel.

Figure 6.9 (a) shows the first four harmonics of a series AC processing pickup operating at a $Q_{20}=5$ obtained from Fourier analysis. In this figure, the amplitude of the harmonics is expressed as a fraction of the fundamental component under full load conditions. It can be seen that the amplitudes of the harmonic components are all relatively low compared to the fundamental. The highest harmonic component for the inductor current does not exceed 6.3% of the maximum fundamental component. Likewise, the harmonics of the parallel topology are also plotted in Figure 6.9(b). In relative comparison, the parallel topology has smaller harmonic components because of the extra filtering stage of the pickup inductor. As such the winding losses due to harmonics in the pickup inductor for the series topology will always be higher than the parallel topology. Although the same can be said for the harmonics reflected back onto the primary track, in practice the low coupling of the secondary and the natural filtering in the resonant circuit on the primary significantly reduces these harmonics and the associated losses are therefore negligible.
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![Graph](image)

\[ Z_{t} = \frac{\omega M^2}{L_2} \]  \hspace{1cm} (6-21)

Figure 6.9. Harmonic Components of Inductor Current as Percentage of the Maximum Fundamental Value at \( Q_{20} = 5 \) for (a) Series and (b) Parallel.

The normalized reflected impedance characteristic for the series topology is shown Figure 6.10(a) at different values of \( Q_{20} \). Both the resistive and reactive components are normalized against the factor:

The factor \( Z_t \) is the transfer impedance of the pickup and it is a direct measure of the magnetic coupling and power transfer capability. A phase delay \( \phi \) of zero is plotted as points on the right hand side of the semicircle on Figure 6.10(a) for each \( Q_{20} \) value. As \( \phi \) is increased in 10º steps, the impedance points move along the semicircle towards the left hand side. The real and imaginary power sourced by the primary power supply can be easily determined after the impedance values are denormalised and substituted into (6-15) and (6-16). When \( \phi \) is zero, only a resistive load is reflected back on the track and the real power is supplied by the power supply to drive the pickup. As \( \phi \) increases (thereby decreasing the output power), both a real and a capacitive load are reflected on the primary track, and the power supply has to source both the real power and the extra capacitive VAR’s. When \( \phi \) increases towards 180º, both the reactive and resistive load decrease to zero. For the parallel topology, the reflected impedance is similar to the series topology with only one difference in that the initial impedance is offset at negative one.
6.4 Design Procedure

In this section, the design of a 1.2kW series AC processing pickup is described. The maximum desired output voltage is 90V and the AC load resistance is 6Ω. An asymmetrical S-shaped magnetic inductor was chosen for the prototype pickup because of its higher output power with the same ferrite volume/length compared to traditional magnetic pickup structures [20]. This pickup has a measured $V_{oc}$ of 90V and an inductance value of 115μH. The primary IPT converter uses an LCL topology operating at a fixed frequency of 20 kHz with 125A in the primary track [127]. For a perfectly tuned resonant tank, the nominal tuning capacitance is 551nF and this value was used in the prototype. Using (6-3), $Q_{20}$ of the circuit is 2.4 at maximum power. Equations (6-11) and (6-12) are then used to solve for steady state operation. Figure 6.11 shows the calculated waveforms for the circuit with a phase delay $\phi$ of 0°, 58° and 85° corresponding to 100%, 50% and 20% power, respectively.
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Figure 6.11. Calculated Waveforms when Phase Delay \( \phi \) is (a) 0°, (b) 58° and (c) 85°.

6.4.1 Implementation of the AC switch

The AC switch, shown in Figure 6.2, has to conduct current and block voltage in both directions. Figure 6.12 shows three possible ways of forming a bidirectional AC switch using MOSFET's. Type-1 is not ideal for the series AC processing pickup as it is primarily used in low voltage high current applications and the forward voltage drop of two diodes compared to one diode drop in other configurations is less efficient. In the Type-2 configuration, current flows in either direction taking the top or bottom path. One of the switches in this set of devices has to have an isolated gate drive (or high side gate drive). The Type-3 configuration has advantages of common ground between the transistors, so it is easier to drive the gates. Furthermore, the body diodes of the switching devices can be used without the need for external diodes, provided the body diodes have sufficiently low loss.

Figure 6.12. Bi-directional Switches.

One key requirement for realizing the AC switch in practice is to allow the switches to operate on simple gate drive waveforms. In section 6.1, a very precise timed gate drive waveform at 40kHz is required for the series AC processing pickup used in 20kHz IPT
systems. However, in practice it is very difficult to generate this waveform with sufficient accuracy using microcontrollers. Any error in timing forces the circuit into the discontinuous state when the inductor current is non zero and the resulting overshoot in voltage across the switches may cause device failure. In consequence, it is more desirable to use the diode’s turn-off behaviour to block the current in the reverse direction and allow the circuit to naturally enter the discontinuous state due to diode commutation. Hence, the PWM gate signals are only required to be operating at 50% duty cycle at 20kHz with a phase delay $\phi$ relative to the phase of $V_{oc}$ to allow the circuit to enter the resonant state, while the natural diode turn off characteristic allows the circuit to enter the discontinuous state. Figure 6.13 shows the gate drive waveforms while the circuit diagram is shown in Figure 6.14.

If the diode was to ideally turn off at zero current, a reverse recovery charge of zero would be required. However, a diode with a reverse recovery charge of zero does not exist in practice, so a suitable switch that has a diode with a low reverse recovery charge is needed. Here the Type-3 AC switch configuration from Figure 6.12 was used, with IGBT’s (IRGP20B60PDPbF) rather than MOSFET’s because this switch has an extra internal diode.
with a low reverse recovery charge of 80nC. A simple RC snubber was designed to damp the high frequency resonant oscillations caused by the pickup inductance and parasitic output capacitance of the IGBT at the instant the body diode naturally turns off. The oscillation arises because the reverse recovery current of 3A becomes the initial inductor current of an LC resonant circuit comprising the pickup inductance of 115\(\mu\)H and the switch output capacitance of 130pF. With reference to [128], a snubber is designed with a damping factor of 1. It was experimentally measured that the maximum power loss in the snubber is less than 4W under any phase delay \(\phi\).

### 6.5 Component Stress

To calculate the maximum ratings of the circuit components, the phase delay \(\phi\) has to be set slightly above zero degrees in order to observe the peak voltage across the switches and maximum RMS rating for the capacitor and inductor. The calculated peak and RMS value of the voltage and current for the capacitor, inductor and switch are listed in Table 6.1. It can be concluded from Table I that the switches have to be rated for both 310V and 22A at normal operation. However, an overshoot of 10% is still possible from the snubber design despite a damping factor of 1, so the switch rating should be greater than 350V. In practice a 600V device is used.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Peak</th>
<th>RMS</th>
</tr>
</thead>
<tbody>
<tr>
<td>(V_c)</td>
<td>307V</td>
<td>217V</td>
</tr>
<tr>
<td>(V_L)</td>
<td>309V</td>
<td>218V</td>
</tr>
<tr>
<td>(V_s)</td>
<td>309V</td>
<td></td>
</tr>
<tr>
<td>(I_L)</td>
<td>21.2A</td>
<td>15A</td>
</tr>
<tr>
<td>(I_s)</td>
<td>21.2A</td>
<td>15A</td>
</tr>
</tbody>
</table>

### 6.6 Controller

A practical system setup with controller for the AC processing pickup is shown as a block diagram in Figure 6.14. The phase of \(V_{oc}\) is measured using a separate phase sense coil \(L_3\) placed on the primary track to detect the phase of the track current, which is exactly 90° out of phase with the open circuit voltage (\(V_{oc} \propto I_1 \angle 90°\)). The phase sense coil cannot be placed too
close to the main pickup inductor as the mutual coupling between them may cause phase distortion. However, it was found in practice that if the two were placed more than 300mm apart, the distortions are negligible. The phase delay $\phi$ is set by a computer interface while a microcontroller adjusts the switch gate drive waveforms accordingly.

![Diagram](image)

**Figure 6.14. Block Diagram for the Controller.**

### 6.7 Series and Parallel Connection for Switching Devices in Series AC Pickups

Series and parallel semiconductor switches have been widely used in high voltage or high current power electronic systems. The use of series and parallel connections for the Series AC processing pickup is outlined in this section.

#### 6.7.1 Series Switch Connection

In a series connection of power semiconductor devices, the important issue is to maintain voltage sharing among each member of the series string during both steady state and transient
states. Practical design techniques are introduced below for series connection of devices.

6.7.1.1 Static Voltage Balance

Static voltage balance refers to a uniform spread of the total voltage across all switching elements in a string of devices under steady state conditions. A very simple method of achieving such a condition is to match the output impedance of all switching devices using a simple voltage divider as shown in Figure 6.15. The resistances within the potential divider should be chosen to be 10 times lower than the off state output impedances of each semiconductor device.

![Figure 6.15. Series Switch with Voltage Divider.](image)

6.7.1.2 Dynamic Voltage Balance

To achieve dynamic voltage sharing between a series string of devices in a series AC processing pickup, more complex methods must be used. It was previously noted that the series AC processing pickup also achieves soft switching conditions, however, the AC switch only achieves ZCS. The voltage across each device rises very rapidly as the switch changes state. There are many techniques for achieving dynamic voltage sharing such as passive snubbers [113], active gate control [107, 109, 111, 114, 115, 118, 120, 123] and voltage clamping [109, 110, 117, 119, 121]. Here only the most common type of voltage sharing technique using a passive snubber will be briefly outlined.

A fully forward biased snubber comprising of a resistor, a capacitor and a diode is shown in Figure 6.16. For the series AC processing pickup of Figure 6.17, the IGBT body diode self commutates its turn off. Under this state, the capacitor $C_s$ will be charged. The operating
waveforms are shown in Figure 6.18.

During the turn-off state the voltage across the diodes is naturally balanced as the reverse voltage across both switches is applied exactly at the same time during diode turn off.

During the turn on state of the switches, the voltage across both switches cannot be guaranteed to be the same as the gate drive signal for the top switch may be slower than the bottom switch. If the bottom switch turns on more quickly, the voltage across the top switch
builds at a rate primarily controlled by the snubber capacitance $C_{s4}$. It is complex and difficult to predict the exact snubber capacitor voltage (switch voltage) as the ODE that models the phenomenon is in the form of a transcendental equation and cannot be solved analytically. Here an approximate analysis procedure on the simplified equivalent circuit shown in Figure 6.19 is used. Several assumptions are listed below to reduce complexity without introducing any significant error.

1. The $V_{oc}$ contribution to the snubber capacitor voltage is small and is ignored. Since $V_c$ is at its peak and $V_{oc}$ is leading by approximately 90 degrees, the $V_{oc}$ contribution is marginal.

2. The resonant capacitor voltage can be assumed to be a constant voltage source. This assumption can be satisfied if the snubber capacitance is much lower than the resonant tuning capacitance which in most practical circuits will be true.

3. The damping of the load resistor can be ignored. Since the snubber capacitance is much smaller than the resonant capacitance, the frequency that the snubber capacitance resonates at with the pickup inductor will be much higher than the system frequency. If the frequency is higher, then the equivalent $Q_2$ of the circuit in an already high $Q_2$ pickup will be even higher and the damping is negligible.

With the above assumptions, a simple ODE for the snubber capacitor voltage can be written as:

$$\frac{d^2 V_{c_{s4}}}{dt^2} + \frac{V_{c_{s4}}}{L_2 C_{s4}} = \frac{V_{c2_{max}}}{L_2 C_{s4}}$$

(6-22)
Note the initial conditions: the switch is in the off state and assuming steady state voltage balance is achieved, the initial condition for the snubber capacitor voltage across it is $V_{c_{\text{max}}}/2$ and the initial inductor current is zero. By solving (6-22) and noting that $V_{cs4}(0)=V_{c_{\text{max}}}/2$ and $i_{L2}(0)=0$, the solution to the capacitor voltage can be obtained:

$$V_{Cs4}(t) = -\frac{V_{c_{\text{max}}}}{2} \cos(\omega_{f}t) + V_{c_{\text{max}}}$$  \hspace{1cm} (6-23)

where

$$\omega_{f} = \frac{1}{\sqrt{L_{2}C_{s4}}}$$  \hspace{1cm} (6-24)

By equating the snubber capacitor voltage to the maximum ratings of the switch, the time delay between the gate drive signals is:

$$t_{\text{ag}} \leq \sqrt{L_{2}C_{s4}} \arccos \left( 2 - \frac{2V_{s_{\text{max}}}}{V_{c_{\text{max}}}} \right)$$  \hspace{1cm} (6-25)

It can be seen from (6-25) that the maximum gate propagation delay in the top switch that can be allowed without exceeding the device voltage rating is directly controlled by the snubber capacitance.

To verify the accuracy of the above assumptions, analytical calculated results using (6-25) and those obtained from simulation are compared in Figure 6.20. The red trace shows the simulation results and the blue shows the theoretically predicted results. It can be seen that the results are a close match for different $V_{s_{\text{max}}}$ and snubber capacitance values. The circuit parameters used to obtain the simulation results are shown in Table 6.2.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{2}$</td>
<td>100μH</td>
</tr>
<tr>
<td>$C_{2}$</td>
<td>158nH</td>
</tr>
<tr>
<td>$V_{oc}$</td>
<td>50V</td>
</tr>
<tr>
<td>$Q_{2}$</td>
<td>5</td>
</tr>
<tr>
<td>$\omega$</td>
<td>$2.513 \times 10^{5}$ rad/s</td>
</tr>
</tbody>
</table>
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The choice of the snubber resistance is determined by the minimum on-time of the switches. The minimum on-time can be designed to be at least five times that of the time constant of the fully forward biased snubber. Theoretically the minimum on-time is zero, however, when the on-time is zero, no power is transferred in a series AC processing pickup and no voltage appears across the switches. It was found that using the normalised graphs in Figure 6.7, the power drops to negligible levels when the conduction time is one sixth of the IPT system period. Hence,

\[ R_s \leq \frac{1/6 f}{5C_s} \]  \hspace{1cm} (6-26)

6.7.2 Parallel Switch Connection

When power semiconductor devices are connected in parallel, the important issue is to maintain current sharing among each parallel path during both steady state and transient switching conditions.

6.7.2.1 Static Current Balance

In practice, static current sharing is mostly achieved using switches that have a positive temperature coefficient. Nominally devices such as MOSFET’s inherently achieve current sharing as the resistance in the device will rise when its temperature is higher than any other
device. As a result, it will carry less current and the temperature will be shared among all parallel devices.

6.7.2.2 Dynamic Current Balance

One key switching condition for the series AC processing pickup is that it achieves ZCS when the AC switch is operating. As such, the current through the switching devices must be close to zero when each switch transiently changes condition. Hence, dynamic current balance is achieved inherently in the series AC processing pickup as the current through each parallel device during switching will be ideally zero. The propagation delay between the gate drive signals is normally negligible as one gate drive signal usually controls all devices in parallel and providing the gate drive circuits one near each gate, negligible propagation delays are guaranteed.

6.8 Discussion and Experimental Results

6.8.1 Controlled AC Output

The series AC processing pickup as described above was coupled to a small section of track (Figure 6.14) and used to drive 1.2kW into a 6Ω AC resistor. Figure 6.21 shows the circuit waveforms for the series AC processing pickup at 100%, 50% and 20% power when \( \phi \) is set to 0°, 58°, and 85°, respectively. It shows different waveforms: inductor current, capacitor voltage, inductor voltage and switch voltage. The inductor current and capacitor voltage are both sinusoidal having low distortion at 100% power. The small voltage overshoots across the switch voltage is due to the snubber oscillations discussed in section 6.4.1. The measurements as shown have excellent correlation with the calculated waveforms in Figure 6.11. The amplitudes of the measured waveforms are within 10% of the values calculated by (6-11) and (6-12). Since the series AC processing pickup has an AC voltage source property, it can be used to drive incandescent lights.
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6.8.2 Controlled DC Output

A series AC processing pickup outputting DC, shown in Figure 6.2, was designed using the parameters shown in section 6.4 with a low level of output voltage ripple (<0.1%) utilizing a 2.2mF DC capacitor. The pickup was operated with a closed-loop controller where the output voltage was set to a desired value in the microcontroller. The microcontroller was configured to maintain the desired load voltage by adjusting $\phi$ in accordance to feedback from the output voltage. Figure 6.22 shows the circuit waveforms when the output voltage is regulated to a DC voltage of 80V for 8$\Omega$, 12$\Omega$ and 16$\Omega$ load, respectively. The waveforms are gate drive $V_{g1}$ (Figure 6.14), inductor current, input voltage to the rectifier and DC output voltage. It can be seen that the output voltage is nearly constant with an error of less than 1% around the desired value. Due to the relatively low $Q_2$ value in this circuit, the inductor current is represented as an exponentially damped sinusoid. The maximum voltage that appears across the switch is 340V operating with the rectifier. This conforms with the results deduced in section 6.4.

Figure 6.21. Measured Waveforms for (a) 100% Power, (b) 50% Power and (c) 20% Power with Light Bulb Load.

A series AC processing pickup outputting DC, shown in Figure 6.2, was designed using the parameters shown in section 6.4 with a low level of output voltage ripple (<0.1%) utilizing a 2.2mF DC capacitor. The pickup was operated with a closed-loop controller where the output voltage was set to a desired value in the microcontroller. The microcontroller was configured to maintain the desired load voltage by adjusting $\phi$ in accordance to feedback from the output voltage. Figure 6.22 shows the circuit waveforms when the output voltage is regulated to a DC voltage of 80V for 8$\Omega$, 12$\Omega$ and 16$\Omega$ load, respectively. The waveforms are gate drive $V_{g1}$ (Figure 6.14), inductor current, input voltage to the rectifier and DC output voltage. It can be seen that the output voltage is nearly constant with an error of less than 1% around the desired value. Due to the relatively low $Q_2$ value in this circuit, the inductor current is represented as an exponentially damped sinusoid. The maximum voltage that appears across the switch is 340V operating with the rectifier. This conforms with the results deduced in section 6.4.

Figure 6.22. Measured Waveforms for (a) 8$\Omega$, (b) 12$\Omega$ and (c) 16$\Omega$ load with constant 80V DC voltage.
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The simulation results of a series AC processing pickup during a startup transient are shown in Figure 6.23 with parameters in section 6.4. It can be seen that the inrush current is completely eliminated by means of controlling the inductor current with the AC switch. For comparison purposes, a traditional series IPT pickup startup response is also simulated with similar component parameters. This circuit has very large inrush currents which are 12 times higher than its steady state values. Hence, it can be seen that the series AC processing pickup can control the inductor current very effectively and minimize any inrush currents. In practice the component cost can be reduced by minimizing the ratings of the pickup inductor and tuning capacitor to only their steady state operation ratings.

![Figure 6.23. Inductor Current, Resonant Capacitor Voltage, DC Output Voltage Waveforms for (a) Series AC Processing Pickup and (b) Conventional Series IPT Pickup.](image)

**6.9 Efficiency measurement**

An efficiency vs. output power plot is shown in Figure 6.24(a) for the series AC processing pickup outputting a controlled AC voltage to a 6Ω load. In addition, the efficiency of the overall system including the AC processing pickup with rectifier outputting controlled DC to an 8Ω load is also plotted. Referring to Figure 6.14, the overall IPT system efficiency is determined using measurements of the DC input power to the LCL primary converter and the AC output power from the secondary pickup. Similarly, the pickup efficiency is calculated
using the AC input power delivered by the open circuit voltage source \( V_{oc} \) and the AC output power from the secondary pickup. The \( V_{oc} \) source is only present in a transformer coupling model and cannot be directly measured while the circuit is in operation. As such, an extra coil \( L_4 \) (shown in Figure 6.14) is used to estimate the actual phase and magnitude of \( V_{oc} \). This method was found to be an accurate technique for measuring the input power given by:

\[
P_{in} = \frac{\omega}{2\pi} \int_0^{2\pi/\omega} V_{oc}(t)I_L(t) dt
\]  

(6-27)

The pickup efficiency measurement neglects the supply (LCL converter) power losses and gives a more meaningful measure of the conversion efficiency of the pickup itself. The DC efficiency is measured with the losses in the rectifier taken into account. The efficiency of the pickup remains above 91% when the output power is more than 100W to the AC load. With a 1.2kW load, the efficiency of the series AC processing pickup and the overall IPT system can reach as high as 93% and 84%, respectively when outputting either controlled AC or DC voltage. A breakdown of the power loss measurements for the series AC processing pickup are shown in Figure 6.24(b). The switch losses, pickup inductor coil and core losses, capacitor ESR losses, snubber losses, rectifier losses and stray losses are all accounted for individually. The stray loss includes copper losses of the PCB tracks and connectors. At full power there are 35W of loss in the switch giving a 3% reduction in efficiency compared to a traditional series tuned pickup without a buck converter. In a cascaded buck converter, the efficiency would be lower due to additional losses in the DC inductor. At lower power levels, the long discontinuous parts in the resonant waveforms as a result of the controlled switch operation (Figure 6.13) decrease the losses in all components and the efficiency of the overall pickup is maintained relatively high. For the harmonic components generated by the discontinuous control of the AC processing method shown in Figure 6.9, there are pickup inductor losses due to skin effect. However, the magnitude of the skin effect losses is only 10% of the maximum losses of the fundamental component for this 20 kHz IPT system.
6.10 Summary

This chapter presents a new series AC processing pickup which can produce a variable AC voltage source suitable for driving incandescent lights. By cascading a rectifier after the pickup, it can also output a controlled DC voltage. This modified topology eliminates the need to use a buck converter in the traditional series tuned controllers with the added advantages of enabling a controlled transient response, lower steady state losses and reduced cost compared to traditional series tuned controllers. The capability of the circuit to control the resonant current and voltage directly to achieve either an AC or DC output, eliminates the need to overrate components at start up. Moreover, the pickup operates under near ideal soft switching conditions which results in high efficiency. The AC processing pickup can control the output power over a wide load range for a 1.2kW system and a maximum efficiency of 93% was measured.
7 Investigation of an AC Controlled Cyclo-converter

In previous chapters, the different topologies of the AC processing pickup are outlined along with its potential applications. In this chapter, the AC processing concept is applied to IPT systems to develop new applications. This chapter investigates a direct AC-AC IPT system which takes 50Hz mains input and has 50Hz mains output without requiring a DC link in conventional systems. This technique is based on the AC processing concept with matrix converters giving advantages of low energy storage. In addition, this system can potentially have a higher efficiency due to its ability to directly convert from AC to AC without any intermediate DC stage.

7.1 AC-AC IPT Systems

One conventional method to achieve an AC-AC IPT system is shown in Figure 7.1. The compensation capacitors and the power flow controller on the secondary are assumed to be present and not shown for simplicity. The 50Hz mains input voltage is first rectified to DC with large electrolytic capacitors to smooth the 100Hz input ripple. Power factor correction circuitry is usually employed along with the rectification stage for high power applications. After the DC link, a DC-AC converter is used to generate a high frequency AC current. The primary track current induces a voltage in the secondary inductor coil which is then converted into DC via a rectifier and filter stage. The DC is then converted back into 50Hz AC via a PWM converter and a filter network.

Figure 7.1. A Conventional AC-AC IPT System.
One disadvantage in such systems is the requirement of large DC electrolytic capacitors. These capacitors are very expensive and contribute to a significant portion of the overall IPT system cost. In addition, the capacitors are also bulky in size and heavy in weight which makes the overall system cumbersome. Moreover, if there is a fault in the DC-AC primary converter and cause a short across the DC capacitor, the large energy storage onboard may cause catastrophic failures.

Another option to design an AC-AC IPT system is to use a low energy storage power supply. A Barracuda power supply has been recently proposed that outlines a technique to directly generate high frequency track currents without the use of a DC voltage link [129, 130]. This technique uses a small filter capacitor after the rectifier to reduce the energy storage of the power supply. The small filter capacitor cannot filter out the rectified 100Hz ripple and these ripples become the envelope of the track current. In such systems, the primary track current waveforms are similar to an Amplitude Modulated (AM) signal where its high frequency carrier will have a low frequency envelope. The current induces a voltage in the secondary inductor coil which is then compensated with capacitors. A rectifier and a large filter network are used to produce the DC link required on the secondary. A similar DC-AC stage can be used to generate the final 50Hz output waveform. However, it should be noted that this technique does not actually reduce the required energy storage in the system, but moves the DC energy storage from the primary side to the secondary side. Hence, the disadvantages stated above still apply to such systems.

In this section, a novel AC-AC IPT system is represented without significant energy storage. This new systems combines the Barracuda power supply with a secondary AC-AC cyclo-converter that converts the high frequency AC directly to a low frequency AC output. In addition, the AC processing control approach described in chapters 3-4 is employed in the AC-AC cyclo-converter to accurately generate the output AC waveform despite distortions in the Barracuda track current under light load conditions.
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

7.1.1 Circuit Operation

The new AC-AC IPT system along with its basic operating waveforms is shown in Figure 7.2. The primary side is assumed to be driven from a Barracuda power supply [129, 130]. The AC input voltage is first rectified, however a small capacitor is used to produce a DC voltage with large half sinusoidal ripples. Note that, this DC is very different from the DC link produced in traditional systems. An H-bridge and compensation network is used to generate a high frequency track current. The resulting track current takes the form of an AM signal which has a low frequency envelope with a high frequency carrier. A cyclo-converter in the form of a reversible rectifier is used to generate 50Hz AC directly. The AC processing control method is combined in the cyclo-converter along with a simple Proportional Integral (PI) controller to realise a high quality AC output waveform.

7.1.1.1 The Barracuta Power Supply

The detailed operation and design procedure of the Barracuda power supply has been thoroughly discussed in [129]. Here, a brief review on the operation and design of this power supply is presented for completeness. A Barracuda power supply is shown in Figure 7.3.
A small filter capacitor $C_f$ is used after the rectifier to smooth the waveform at the input of the H-bridge. The capacitor is chosen so that under rated load, the voltage across it closely follows the absolute value of the input voltage during the complete period. If the capacitor is too large, the voltage will discharge too slowly and cause “distortions” in the track current at the desired 100Hz modulation. On the other hand, if the capacitor is too small, the voltage will drop to zero before the AC input reaches zero. In both situations, Power Factor (PF) at the input of the converter is reduced because of the “distortion”. Hence, a trade-off must be made. The switches in the H-bridge are operated with phase control as shown in Figure 7.4. Here in the $V_i$ waveform, the 50Hz modulation envelope is ignored for simplicity but necessarily must be present. The gate driving waveforms of $S_{1p}$, $S_{3p}$ and $S_{2p}$, $S_{4p}$ are complementary. The gate signals of $S_{1p}$, $S_{3p}$ and $S_{2p}$, $S_{4p}$ are at a controlled phase angle $\sigma$.

![Figure 7.4. Operating Waveform of LCL Converter.](image)

The output voltage relationship with the phase angle is given by:

$$V_i = \frac{2\sqrt{2}}{\pi} V_d \sin\left(\frac{\sigma}{2}\right)$$  \hspace{1cm} (7-1)

The LCL (Inductor Capacitor Inductor) network is designed such that all the reactances in each branch are identical and equal to $X_1$. The first inductor is produced by matching the equivalent reactance of the leakage inductance $L_b$ in the transformer and DC blocking capacitor $C_b$, to the desired value. The capacitor $C_1$ and primary track inductor $L_1$ must also be
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

tuned at $X_1$.

$$X_1 = \left( \omega L_b - \frac{1}{\omega C_b} \right) = \frac{1}{\omega C_i} = \frac{1}{\omega L_i}$$ (7-2)

The primary track current in the Barracuda power supply is given by rearranging (7-1) and (7-2):

$$I_1 = n \frac{V_i}{X_1} = n \frac{2\sqrt{2} V_d}{\pi X_i} \sin \left( \frac{\sigma}{2} \right)$$ (7-3)

It can be seen from (7-3) that the primary track current is dependent on several system parameters such as input voltage, the reactance $X_1$ of the LCL network, the turns ratio of the transformer and the phase angle $\sigma$. To regulate the primary track current, phase angle control is used. Under normal design criteria, it is desirable to design the phase angle at 120º as the harmonics in the primary track current can be minimized [129]. Slight variations in conduction angle are used by the primary converter controller to form a closed loop controller to regulate the track current to a fixed reference irrespective of load variations.

7.1.1.2 The Cyclo-Converter

A secondary pickup with a cyclo-converter is shown in Figure 7.5. The secondary pickup consists of a parallel AC processing circuit. In this application, it will be assumed that nominal tuning is used so that the resonant capacitor is fully tuned with the pickup inductor. The main purpose of the AC processing circuit is to regulate the output voltage to the nominal desired output value irrespective of load changes. In addition, its secondary purpose is to produce a pure sinusoidal envelope in the resonant capacitor voltage $V_c$ if the primary track current envelope has distortion at light load conditions not enabling the small DC filter capacitor $C_f$ to discharge fully [129]. Since the control of the AC processing circuit has a fast transient response and has the ability to directly control the amplitude of the resonant capacitor voltage, it is sufficient to correct for any 50Hz distortions within the track current. The reversible rectifier and output filter is used to convert the high frequency resonant capacitor voltage with a sinusoidal envelope directly to a 50Hz output. The reversible rectifier
shown uses four AC switches which are each formed using a pair of back to back MOSFET’s. Each switch can block voltage and conduct current in both directions in the reversible rectifier. The objective is to convert the resonant capacitor voltage back into a 50Hz waveform. The polarity of $V_c$ alternates every 100Hz but rectified to produce a $V_{ac}$ output that has a 50Hz fundamental component with harmonics that are more than two times the IPT operating frequency as shown in Figure 7.6. The output filter removes all the harmonics and the desired 50Hz component is retained. Note the size of the output LC filter is much smaller than the one required to generate DC mentioned in chapters 5-6 as it only has to reject the high frequency harmonics while maintaining both the amplitude and phase component of the fundamental 50Hz component. Hence the filter components are chosen such that the cut frequency is much higher than the 50Hz and much lower than the two times the IPT resonant frequency:

$$f_{\text{main}} \ll \frac{1}{2\pi \sqrt{L_{ac}C_{ac}}} \ll 2f_0$$

(7-4)

Figure 7.5. A Secondary Pickup using AC Processing with Reversible Rectifier.

The operating waveforms of the secondary pickup over two 50Hz cycles are shown in Figure 7.6. The AC processing circuit controls the envelope of the resonant capacitor voltage to closely follow an amplitude modulated sine wave. The amplitude of the resonant capacitor voltage is controlled using phase delay control by driving the switches $S_a$ and $S_b$. Note here, in this example, no distortion in the track current is assumed. The polarity of the voltage after the rectifier is switched either positive or negative every 100Hz. The inductor capacitor filter
removes the high frequency component at two times the IPT frequency and the 50Hz output voltage is generated.

Figure 7.6 Operating Waveforms of Secondary Pickup showing 50Hz Output.

Figure 7.7 shows the operating waveforms of the secondary pickup at high frequency. It should be noted that there are many possible methods to control the switches in the reversible rectifier. Here, a simple gate drive waveform is used although more complex optimized gate drive waveforms will be discussed in Section 7.1.2. It is assumed that the output voltage to the load is positive and changes slowly. The circuit operation presented as a sequence of linear circuit stages corresponding to each mode is shown in Figure 7.8. $V_{gsa}$ is the PWM gate signal driving switch $S_a$ at 50% duty cycle, at the same frequency as the IPT track frequency. The gate signal for switch $S_b$ is exactly the inverted version of $V_{gsa}$. Consider the situation where waveforms $V_{gsa}$ and $V_{gsh}$ are controlled with a phase delay $\phi$ relative to the phase of $V_{oc}$ to clamp parts of the resonant capacitor voltage as shown in Figure 7.7. In Mode 1 ($M_1$), $S_a$, $S_2$, $S_3$, $S_4$, $S_5$ are turned on and $S_b$, $S_1$, $S_2$, $S_7$, $S_8$ are turned off. The resonant capacitor voltage is clamped and negligible energy is transferred across the reversible rectifier. The switch in the reversible rectifier is operated to provide a conduction path for the filter inductor current. In Mode 2 ($M_2$), $S_a$, $S_2$, $S_3$, $S_4$, $S_5$ are turned off and $S_b$, $S_1$, $S_2$, $S_7$, $S_8$ are turned on. The parallel resonant circuit behaves like a parallel resonant tank and energy is transferred across the reversible rectifier. In Mode 3 ($M_3$), the operation is very similar to Mode 1, with the
difference that the resonant capacitor voltage is now negative. In Mode 4 ($M_4$), similar to $M_2$, the resonant circuit delivers energy. After this mode, the circuit returns back to $M_1$, repeating the switching process.

In summary, switches $S_a$ and $S_b$ operate using the same basic approach for the AC processing pickup as outlined in chapter 3. In addition, the reversible rectifier operates very similar to a conventional rectifier. The only difference is that the switching gate waveforms changes by 180º or reverses polarity when the output voltage alternates.

Figure 7.7. Waveforms of Secondary Pickup showing High Frequency Operation when the Output Voltage ($V_{ac}$) is Positive.
7.1.2 Cyclo-converter Modulation Strategies

A simple modulation strategy was used to control the reversible rectifier in Section 7.1.1.2. In this section, a more thorough discussion on the possible modulation strategies is outlined. The reversible rectifier and the AC switch for the controller can be integrated to form one cyclo-converter with exactly the same structure as the reversible rectifier as shown in Figure 7.9. This reduces the number of semiconductor switches required from 10 to 8, giving a reduction in cost and component count. However, new modulation techniques are required to control the cyclo-converter.

In its new structure, the cyclo-converter has 8 switches. Each switch has two possible
states, namely on and off. This generates 256 \((2^8)\) possible switching states for the cyclo-converter. Although there are a vast number of switching states, many can be grouped together in individual operating modes describing their equivalent operating condition. Below is a description of all the operating modes.

7.1.2.1 Not Permissible (NP) Mode

The Not Permissible (NP) mode describes an operating condition when the circuit cannot physically operate in. These modes are defined to be the condition when a device component gets damaged as a result of operation. In this cyclo-converter, this can be simply regarded as the cyclo-converter no longer providing a path for the filter inductor current \(I_{Lac}\) to follow. An example of this is when all the switches are turned off and there is inductor current \(I_{Lac}\) present. The effect of opening an inductor when current is flowing will directly damage the cyclo-converter.

7.1.2.2 Energy Injection (EI) Mode

The Energy Injection (EI) mode describes an operating condition when the resonant tank is not connected with the filter network or load, and allows the energy to build up in the resonant network by energy injection. A typical example of the circuit operating in the EI mode is shown in Figure 7.10. For a perfectly tuned parallel resonant circuit, the voltage across the capacitor will build up towards infinity. If the circuit is left uncontrolled in this state, the voltage and current in the resonant network may be too high and result in component failure. Unfortunately, direct generation of a 50Hz output using this technique without significant energy storage, is not very useful. The main reason for this is that energy will continue to be injected from the track in to the tuned circuit at the input of the rectifier, but as the output load is decoupled from the input, the resonant capacitor voltage will increase as the circuit operation is maintained inside the EI mode, and thereby increasing the voltage ratings of the switches in the cyclo-converter and resonant network. This leads to an increase in both cost and size of the main pickup components. As such, the EI mode cannot be used without careful design and a thorough discussion of the operation of the circuit in this EI mode is
outside the scope of this thesis.

7.1.2.3 Energy Transfer (ET) Mode

During the Energy Transfer (ET) mode, power is delivered from the resonant tank directly to the filter network and the load. A typical example of the circuit operating in the ET mode is shown in Figure 7.11. It should be noted that the ET mode also includes operating conditions when energy is transferred from the load to the resonant tank. Hence, bi-directional power flow is permitted in such systems.

7.1.2.4 Clamp Mode

During the Clamp mode, the resonant tank is shorted to provide no power while a conduction path is provided for the inductor current $I_{Lac}$ to flow. A typical example of the circuit operating in the Clamp mode is shown in Figure 7.12. The parallel resonant tank is clamped using $S_1$, $S_2$, $S_3$, $S_4$, hence no energy is transferred. $S_8$ is turned on to allow the inductor current to circulate through the load. This mode is similar in operation to that which
arises in the AC processing pickup when the parallel resonant tank is shorted. Using this mode, the amount of energy delivered to the load can be fully controlled from the maximum output power down to zero.

![Figure 7.12. A Typical Clamp Mode.](image)

### 7.1.2.5 Energy Injection and Energy Transfer (FOET) Mode

During the Energy Injection and Energy Transfer (EIET) mode, the circuit operates in either the EI mode or the ET mode depending on the voltage polarity of the resonant capacitor voltage $V_c$ and this operation repeats each period of the track frequency. A typical example of the circuit operating in the EIET mode arises when $V_c$ is negative and is shown in Figure 7.13(a). In this state, switches $S_2$, $S_4$, $S_8$ are turned on. The body diode in $S_1$ will block the current conduction path and the resonant circuit is permitted to inject energy, as previously defined for the EI mode. When $V_c$ becomes positive as shown in Figure 7.13(b), the inductor current flows through the resonant circuit enabling energy transfer. Although it may seem that the inductor current can flow through either the resonant tank or the body diode of $S_3$, in practice the current must flow through the resonant tank. The reason behind this statement is very simple. The resonant capacitor voltage can be modeled as a voltage source and the inductor current can be modeled as a current source. The equivalent circuit under this condition will be a voltage source in parallel with a body diode and a current source. Because any current source in parallel with the voltage source is fundamentally a voltage source, the inductor current will preferentially conduct through the voltage source (namely the capacitor voltage) rather than the short circuit path created by the body diode in $S_3$. 

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7.1.2.6 Energy Injection and Clamp (EIC) Mode

During the Energy Injection and Clamp (EIC) mode, the circuit can operate in either the EI mode or the Clamp mode depending on the voltage polarity of the resonant capacitor voltage $V_c$ and this operation repeats each period of the track frequency. A typical example of the circuit operating in the EIC mode when $V_c$ is positive is shown in Figure 7.14(a). In this state, switches $S_1$, $S_4$, $S_8$ are turned on. The body diodes in $S_2$, $S_3$ will block the current conduction path and the resonant circuit is permitted to freely oscillate, defining the EI mode. When $V_c$ becomes negative as shown in Figure 7.14(b), the circuit is in the Clamp mode.

7.1.2.7 Energy Transfer and Clamp (ETC) Mode

During the Energy Transfer and Clamp (ETC) mode, the circuit operates in either the ET mode or the Clamp mode depending on the voltage polarity of the resonant capacitor voltage
$V_c$, repeating each period of the track frequency. A typical example of the circuit operating in the ETC mode arises when $V_c$ is positive and is shown in Figure 7.15(a). In this state, switches $S_2$, $S_3$, $S_8$ are turned on. The body diodes in $S_1, S_4$ will short the resonant circuit (defining the Clamp mode). Although it may seem that the circuit is also in the ET mode, the short circuited resonant tank cannot transfer any power. If $V_c$ is negative as shown in Figure 7.15(b), the circuit is in the ET mode where energy is transferred from the LC filter to the resonant circuit.

![Figure 7.15. A Typical Energy Transfer and Clamp (ETC) Mode showing (a) Clamp Mode and (b) ET Mode.](image)

### 7.1.2.8 Energy Injection, Energy Transfer and Clamp (EIETC) Mode

During the Energy Injection, Energy Transfer and Clamp (EIETC) mode, the circuit can operate in either the EI mode, ET mode or the Clamp mode depending on the polarity of both the resonant capacitor voltage $V_c$ and the inductor current $i_{Lac}$. A typical example of the circuit operating in the EIETC state when $V_c$ is positive is shown in Figure 7.16(a). In this state, switches $S_2$, $S_3$, $S_6$, $S_7$, $S_8$ are turned on. The body diodes in $S_1, S_4$ will short the resonant circuit hence defined as the Clamp mode. Although it may seem that the circuit is also in the ET mode, the short circuited resonant tank cannot transfer any power. If $V_c$ is negative and $i_{Lac}$ is positive, as shown in Figure 7.16(b), the circuit is in the ET mode where energy is transferred from the LC filter to the resonant circuit. If $V_c$ is negative and $i_{Lac}$ is negative, as shown in Figure 7.16(c) the circuit is in the EI mode.
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Figure 7.16. A Typical Energy Injection, Energy Transfer and Clamp (EIETC) Mode showing (a) Clamp Mode, (b) ET Mode and (c) EI Mode.

7.1.2.9 Discussion of Operating State

There are a total of 8 possible operating modes mentioned above. However, there are 256 possible switching states. To group all the number of states into the 8 modes, a time consuming and tedious process is required. As such, logic expressions are derived by the controller to compute the operating modes. The Boolean expression that defines the NP modes are given by:

\[
Q_{NP} = (S_1S_5 + S_1S_7 + S_3S_5 + S_3S_7 + S_4S_6 + S_4S_8 + S_6S_8 + S_2S_6) \tag{7-5}
\]

where \(Q_{NP}\) denotes the NP mode, \(S_1\) denotes that switch \(S_1\) is turned on and the bar on top of the expression denotes a logic inversion. The expression of (7-5) defines the NP mode as any switching state that does not allow inductor current to flow.

The Clamp, EI and ET mode are computed by:

\[
Q_C = \overline{Q_{NP}} \left( S_1S_3S_4 + S_2S_6S_8 + S_2S_5S_6S_8S_9S_1S_3S_8 + S_1S_4S_5S_8S_2S_3S_6S_9 \right) \tag{7-6}
\]
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

\[ Q_{EI} = \overline{Q_{PEI}} \overline{Q_{PET}} \overline{Q_{PC}} \]  
\[ (7-7) \]

\[ Q_{ET} = \overline{Q_{PEI}} \overline{Q_{PET}} \overline{Q_{PC}} \left( s_1 s_3 s_5 s_7 s_2 s_4 s_6 s_8 + \overline{s_1 s_3 s_5 s_7} s_2 s_4 s_6 s_8 \right) \]  
\[ (7-8) \]

where

\[ Q_{PEI} = s_1 s_3 + s_3 s_7 + s_2 s_6 + s_4 s_8 \]  
\[ (7-9) \]

\[ Q_{PET} = s_1 s_7 + s_2 s_8 + s_3 s_5 + s_4 s_6 \]  
\[ (7-10) \]

\[ Q_{PC} = \overline{Q_{NP}} \left( s_1 s_4 + s_2 s_3 + s_5 s_6 + s_6 s_7 \right) \]  
\[ (7-11) \]

\( Q_{PEI}, Q_{PET}, Q_{PC} \) are defined to be the partial modes of the EI, ET, Clamp modes. The definition of the partial mode is quite simple. For example, the partial EI mode define that if any of the 256 possible switching states meet the conditions of the EI mode outlined in the previous section, then it must exhibit the behaviour of the EI mode. However, that switching state may exhibit other operating modes, namely an ET or a Clamp state.

The first two terms in the Clamp mode computation in \( (7-6) \) describe when four switches form a short circuit, hence, the system must be clamped. The last two terms in \( (7-6) \) accounts for clamp modes that are formed using two pairs of back to back body diodes in the switches which short the resonant tank in both directions. The two terms in the ET mode computation in \( (7-8) \) accounts for the condition when all four body diodes are configured like a full bridge rectifier.

The expression below describes the calculation for the other modes:

\[ Q_{EET} = Q_{PEI} Q_{PET} \overline{Q_{PC}} \overline{Q_{ET}} \]  
\[ (7-12) \]

\[ Q_{ETC} = \overline{Q_{PEI}} Q_{PET} \overline{Q_{PC}} \overline{Q_{C}} \]  
\[ (7-13) \]

\[ Q_{ESC} = Q_{PET} \overline{Q_{PC}} \overline{Q_{C}} Q_{1} + Q_{PEI} Q_{PET} \overline{Q_{PC}} \overline{Q_{C}} Q_{1} \]  
\[ (7-14) \]

\[ Q_{EEETC} = Q_{PEI} Q_{PET} \overline{Q_{PC}} \overline{Q_{C}} Q_{1} \]  
\[ (7-15) \]
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where

\[ Q_i = Q_{NP} \left( S_2S_6 + S_4S_8 + S_3S_6 + S_2S_8 \right) \left( S_3S_5 + S_1S_7 + S_3S_7 + S_1S_5 \right) \]  \hspace{1cm} (7-16)

\( Q_i \) defines whether bidirectional inductor current flow is permitted. It should be noted that the second term in the EIC mode in (7-14), may describe a mode where it is actually a EIETC state. However, if the inductor current is unidirectional, then the circuit must exhibit only the EIC state. The inductor current is required to be bidirectional to be classified as the EIETC state.

A computation algorithm can be written using the equations above to quantify each of the 256 possible switching states into individual operating modes. Figure 7.17 shows the number of switching states within each of the operating modes using the computation routine. Different switching states can be used to realise the modulation technique used in any cyclo-converter.

![Diagram](image)

**Figure 7.17. Number of Switching States in Each Operating Mode.**

### 7.1.3 Design Procedure

#### 7.1.3.1 The Primary Barracuda Power Supply

In this section, a design procedure for a 1kW AC-AC IPT system is described. The Barracuda input mains voltage is at 230V rms and 50Hz. The operating IPT frequency is set at 38.4kHz. The system parameters for the Barracuda circuit, shown in Figure 7.3, are listed in Table 7.1. The series components \( L_b \) and \( C_b \) are designed to have a reactance of \( X_1 \) at the
operating frequency \(f_0\). In addition, \(C_1\) and \(L_1\) also have a reactance of \(X_1\). Using (7-3), the calculated primary track current would be 21.2A. However, at lighter load the loading on the track current may be slightly higher than this value as the voltage across \(C_f\) is not fully discharged.

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</tr>
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</tr>
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</table>

### 7.1.3.2 The Secondary Cyclo-converter

The cyclo-converter which combines the AC processing switches and the reversible rectifier shown in Figure 7.9 is used. As discussed in section 7.1.2.9, there are many possible operating modes in such a converter. However, after some practical consideration, many of the possible modes cannot be easily used without sophisticated modulation and control techniques. Firstly, all operating modes that include the Energy Injection (EI) behaviour are discarded from this argument, as the original idea of combining the AC processing technique with the reversible rectifier is to incorporate just the ET and the Clamp mode, and this new state does not classify under any of them. In addition, it also has the disadvantage of increasing the device rating and cost of the system as described earlier. Secondly, in order to realise a simple controller, pure ET and Clamp states are not very useful as they require complex detection circuitry to accurately detect the voltage polarity change across the resonant capacitor voltage to allow the transition between the ET and Clamp mode for output power control. Any delays in such circuits will result in shorting of the resonant capacitor when there is finite voltage across it, damaging the switches. In consequence, the only operating mode that is left is the ETC mode. This mode allow the body diodes in the switches to naturally clamp the resonant capacitor voltage as it changes polarity which is very similar to the AC processing pickup outlined in chapter 3. The four chosen switching states are shown in Table 7.2 for the operating modes in Figure 7.18 and the operating waveforms in Figure
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

7.19. It can be seen from Figure 7.18 that the cyclo-converter essentially has the combined operation of both clamping the resonant capacitor voltage and rectifying the high frequency AC.

<table>
<thead>
<tr>
<th>Conditions</th>
<th>Switching State</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{ac}&gt;0$, $M_1$ or $M_2$</td>
<td>$S_1, S_2, S_3, S_5, S_6, S_7, S_8$</td>
</tr>
<tr>
<td>$V_{ac}&gt;0$, $M_3$ or $M_4$</td>
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</tr>
<tr>
<td>$V_{ac}&lt;0$, $M_1$ or $M_2$</td>
<td>$S_1, S_2, S_3, S_5, S_7, S_8$</td>
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<tr>
<td>$V_{ac}&lt;0$, $M_3$ or $M_4$</td>
<td>$S_1, S_2, S_3, S_4, S_6, S_7, S_8$</td>
</tr>
</tbody>
</table>

Figure 7.18. Operating Modes when the Output Voltage ($V_{ac}$) is Positive.
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

Figure 7.19. Waveforms of Secondary Pickup when the Output Voltage ($V_{ac}$) is Positive.

The pickup inductor $L_2$ has an open circuit voltage of 60V with an inductance value of 47μH. The system parameters for the cyclo-converter circuit, shown in Figure 7.9, are listed in Table 7.3. The resonant capacitor $C_2$ is fully tuned with the pickup inductor $L_2$. The filter network components are chosen using (7-4).

### Table 7.3. Parameters for Cyclo-converter

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{oc}$</td>
<td>60V rms</td>
<td>$L_{ac}$</td>
<td>1mH</td>
</tr>
<tr>
<td>$I_{sc}$</td>
<td>5A rms</td>
<td>$C_{ac}$</td>
<td>1μF</td>
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<tr>
<td>$L_2$</td>
<td>47μH</td>
<td>$V_{ac}$</td>
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</tr>
<tr>
<td>$C_2$</td>
<td>365nH</td>
<td></td>
<td></td>
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</tbody>
</table>

7.1.3.3 Controller

A controller block diagram for the AC-AC IPT system is shown in Figure 7.20. Firstly, the mains input is used as a 50Hz AC reference signal to generate the cyclo-converter output voltage $V_{ac}$ to drive the load. The reference signal $V_{in}^*$ is obtained using a Phase Lock Loop (PLL) locking onto the input mains voltage. Mathematical absolute functions are used to remove the negative signs in both the $V_{ac}$ and $V_{in}^*$ so that the controller does not need to
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

reverse the sign in the PI controller during the negative half cycle. A simple PI controller is used to regulate the output voltage. The PI controller output is directly feed into the phase delay modulator. This modulator produces an output delay with respect to the switch current fully controlled by the PI output $V_u$ (Figure 7.20). The output of this modulator generates 8 gate drive signals capable of directly driving the cyclo-converter.

![Controller Block Diagram](image)

**Figure 7.20. A Controller Block Diagram.**

### 7.1.4 Simulation Results

The system outlined in section 7.1.3 was simulated using the MATLAB inbuilt simulators Simulink and PLECS. The results for the output voltage $V_{ac}$ for a wide range of load conditions are shown in Figure 7.21. It can be seen that $V_{ac}$ is close to a perfect sinusoid at rated load condition. For lighter loads, the distortion in $V_{ac}$ is higher. One key reason is that the primary track current no longer decreases to zero in the nulls of the Barracuda waveform and this causes distortion in the $V_{ac}$ waveform. The PI controller tries to control the cyclo-converter so that less distortion is introduced. However, under light load, it is difficult to use a simple PI controller to correct for all the non-linearities present.
Figure 7.21. Simulated Waveforms of AC-AC IPT System under (a) 10\% Load, (b) 50\% Load and (c) Rated Load

One aspect to note is that the PI controller control output increases at lighter load conditions. This is true because the clamp time should be inversely proportional to the load changes. However, it should be noted that the clamp time only marginally changes over a wide load range. This is because the circuit has an operating $Q_{20} \approx 5.50$, and if the output voltage was to remain constant during regulation ($V_c/V_{oc} \approx 4.26$), then according to Figure 4.7, a constant normalised output voltage of 4.26, for different $Q_{20}$ values from 6 to 10, will only result in a change in $\omega t_c$ of less than 10$^\circ$. For even higher $Q_{20}$ values, the change is even less. This change in clamp angle is equivalent to a change in $V_{fb}$ of less than 0.056, which conforms to results shown in Figure 7.21.

Figure 7.22 shows the step load transient response from 10\% load to rated load conditions. It can be seen that the distortions is low when a 90\% change in load occurs at 32.5ms. The PI controller quickly adjusts to the new load condition by reducing $V_{fb}$. 
Chapter 7. Investigation of an AC Controlled Cyclo-converter.

7.1.5 Discussion

It has been outlined in section 7.1 that the key advantage of using an AC-AC IPT system is the elimination of any large DC capacitors used to generate DC first and then produce a 50Hz AC output. In addition, it also eliminates the need for one more conversion stage thereby improving efficiency. However, the lack of any significant energy storage has disadvantages in other applications. One typical problem with such systems is its inability to drive reactive loads (such as a motor). This is because reactive loads requires the secondary cyclo-converter to both source and absorb energy during operation. Absorbing energy without energy storage cannot be fundamentally achieved. An alternative method is to design such a system so that bi-directional power transfer is permitted and the power can be transferred back and forth in the IPT system to support the reactive load. This means that the IPT system will need to transfer both the real and reactive power to such loads, but, the cost of doing so is quite high and the efficiency of the system is dramatically reduced. In summary, the proposed system has
great difficulty driving reactive loads. The preferred option is to use a conventional IPT pickup controller to generate DC output at the secondary, and then produce 50Hz AC. This way, the reactive energy from any secondary load can be easily feed back to the DC side while the IPT system only needs to be rated to deliver the real power demands.

One other disadvantage is the component rating required to realise such a system. A conventional system as shown in Figure 7.1 has component ratings which are root two times lower than the AC-AC IPT system when rated for identical power delivery. This is because the AC-AC system investigated requires a modulated carrier on the track current that inherently requires higher peak currents to transfer the necessary power. Thus all switches, capacitors and inductors need to be rated for these higher current peaks. In addition, the magnetic ferrites within the system will also need to be rated for higher saturation peaks. Hence, careful considerations must be taken into account when designing such systems.
8 Conclusions and Future Work

8.1 General Conclusions

A comprehensive study of AC processing pickups for IPT systems was undertaken in this thesis. This novel pickup topology has the ability to produce either a controllable AC or DC output with very high efficiency.

A general introduction to IPT systems and its applications was outlined in chapter 1 of this thesis. An overview of traditional power flow controllers used in IPT systems was presented in chapter 2. A discussion of the advantages of each controller for its applications was systematically presented. As shown the earliest IPT pickup controllers (such as the simple voltage regulator or coupling controllers) that utilize magnetic shielding are largely obsolete due to their poor efficiencies and impractical mechanical arrangement. A more recent development in power flow controllers is the decoupling (or shorting) controller. It has been shown to be simple, robust, and widely used in high power industrial applications with multiple pickups. However, when this controller is used in applications with large variations in coupling such as EV charging or biomedical implants, an inefficient method of overrating circuit devices is required. This is particularly evident in systems where the coupling coefficient can change by over 100%. Under the maximum coupling conditions the short circuit current is 100% larger than required which means all the components have to be rated for 200% of rated current which reduces efficiency and increases cost. Alternative controller known as the tuning/detuning controller uses a saturable inductor in the AC compensation path to realize a variable reactance component. However the use of a variable inductor in the non-linear region of the B-H curve limits the efficiency of the overall system. In addition, the variable inductor is expensive to manufacture because it has to manage the high resonant current without fully saturating.

Primary current control can handle a wide load range, a varying coupling condition and
can obtain excellent efficiencies due to its ability to wind down the system. Frequency controllers which change the operating frequency of the primary converter, can operate using changes in tuning components on both the primary and secondary sides to finely adjust the operating frequency to bring the system back into a resonant condition. However, primary side control techniques suffer from their relatively slower transient responses as the output voltage cannot be regulated in sub-millisecond time intervals due to the limited bandwidth of the resonant networks in the IPT system and the wireless communication channel having delays. Moreover, with other wireless communication devices used in applications, the interference generated in many circumstances may lead to loss of data packets which results in output voltage variations that are not acceptable.

Presently, in order to power high frequency AC loads such as fluorescent lights or stage lights, the most common IPT secondary pickup controller rectifies the AC power, which is then regulated using a DC shorting (decoupling) switch and then an extra resonant converter or DC-AC PWM inverter is required to produce a controllable AC source. However, the addition of a second converter is not ideal because of the large number of components required which increase cost. In addition, the two stage conversion process has losses in each stage which reduce efficiency. Likewise, a DC-AC PWM inverter also has a high component count and even more switching losses than the resonant converter due to a hard switching operation at these high frequencies.

Following a general review of IPT controllers, the rest of this thesis focused on developing a new AC processing pickup controller for controlling the power flow in an IPT system, on the secondary side. This controller has been found to have many advantages over existing controllers and can be usefully applied to a wide range of other applications including Electric Vehicle (EV) inductive charging, wireless powering of biomedical implants and industrial applications with multiple secondary pickups on one long primary track.

### 8.1.1 The Parallel AC Processing Pickup

A simple parallel AC processing pickup controller that can provide controlled AC power
over a wide resistive load range is proposed in chapter 3. The AC processing pickup employs switches operating under Zero-Voltage-Switching (ZVS) conditions to clamp parts of the resonant voltage across a parallel tuned LC resonant tank to achieve a controllable AC current source. The fundamental properties and design equations of the AC processing pickup have been investigated. A 550W prototype circuit was designed to verify the validity of the theoretical analysis. A maximum efficiency of 96% was measured with a 500W parallel AC processing pickup.

Practical implementation aspects of the AC processing pickup were discussed in chapter 4. Firstly, a fourth order polynomial approximation was used to produce a look-up table so that the controller can be directly implemented in a microcontroller. Secondly, series and parallel connection of switching devices were outlined in order to achieve higher ratings for practical AC processing pickups for high power applications. Thirdly, a new PWM control strategy was outlined which can significantly decrease the losses in the AC switch of a parallel AC processing pickup, enabling reduced heatsink and pickup size. Lastly, a practical 1.2kW lighting system was built with feedback voltage control to regulate the output voltage to a light very precisely.

A new pickup suitable for providing a controllable DC output using an AC processing pickup with a rectifier was outlined in chapter 5. An equivalent loading model was developed for the rectifier stage and verified using simulation and experimental results. It was found that this AC controlled pickup demonstrates desirable current source properties for battery charging applications.

8.1.2 The Series AC Processing Pickup

Chapter 6 presented a series AC processing pickup using an AC switch (operating under Zero-Current-Switching (ZCS) conditions), in series with a resonant network to produce a controllable AC voltage source suitable for driving incandescent lights. By cascading a rectifier after the pickup, it can also output a controlled DC voltage. This modified topology eliminates the need for a buck converter output stage as is common in traditional series tuned
controllers, with the added advantages of controlled transient response, lower steady state losses, and reduced cost compared to traditional series tuned controllers. The ability of the circuit to control the resonant current and voltage directly allows either controlled AC or DC outputs and eliminates the need to overrate components at start up. A prototype system was constructed and shown to be able to control the output power over a wide load range for a 1.2kW system with a maximum measured efficiency of 93%.

8.1.3 Applications of AC Processing Controllers on IPT Systems

Several alternative application areas for the AC processing controller have been investigated. In addition, to AC lighting, these include battery charging and direct AC mains supply. In lighting systems, it was shown to be the most practical means of producing either a controllable AC current or voltage source for driving a light. Due to its direct AC-AC conversion topology, the system efficiency is greatly improved compared to the AC-DC-AC traditional topology. In addition, the simple circuit structure allows IPT lighting systems to be made at lower costs.

In battery charging applications, the most important advantage was its ability to control the output power over large variations in coupling conditions, while maintaining very high efficiencies compared to other IPT controllers to date. In addition, the soft switching characteristics of the AC processing controllers allow it to operate with a high efficiency of 97% and low EMI.

This controller also can be used to realize direct AC-AC IPT system providing the IPT supply and load are suitably chosen. In order to achieve this, a power supply with low energy bus was used to drive the system. This supply takes 50Hz mains input and converts it into high frequency AC with a modulated envelope. The AC controller inputs modulated high frequency waveform and converts it back to 50Hz mains output without requiring a DC link. Using this new method, efficiency is improved by removing the extra DC conversion stage. In addition, it has advantages of eliminating expensive DC capacitors that contribute to a significant portion of the overall IPT system cost. Such systems are safer under a fault caused
Chapter 8. Conclusions and Future Work.

by shorting the DC link capacitors, compared to traditional systems where it may cause catastrophic failures. The system was analyzed theoretically and simulation results were used to verify its operation.

8.2 Publications and Patents

Three journal and four conference papers have been published as a result of this thesis:


Zealand.

Two patents have also been filed as the outcome of this thesis:


8.3 Suggestions for Future Work

The research conducted in this thesis has investigated the use of AC processing controllers in IPT secondary pickups. To gain a thorough understanding of this new controller, theoretical analysis, simulation results and experimental measurements were undertaken. Although a detailed study has been presented, there are still various areas left for future work which the author believes will be useful research outputs. Some of the suggestions are relatively open ended while others could be regarded as an extension to this thesis.

8.3.1 A Direct AC-AC IPT System using AC Processing Controllers

Chapter 7 has outlined a direct AC-AC IPT system without significant energy storage. This preliminary system shows that this technique, using the AC controller, is achievable, however many aspects of the system need to be improved for a wider range of applications and should be considered in future. The basic operation of the matrix converter used allows power to be transferred from the supply to a load. However, by forcing the switching phase in the matrix converter, energy can be transferred the opposite way, from load to supply. Using this method, heavy reactive loads can be supported by such systems as the reactance power can be transferred back and forth between the load and source, while the real power is delivered to the load. In addition, with advanced control techniques an introduction of the Energy Injection (EI) mode can be used to improve the system performance as this gives the system an idle state where power is neither transferred nor decoupled.
8.3.2 Power Factor Correction on the Primary Side

One disadvantage illustrated with the AC processing pickup is that it reflects a reactive load back onto the IPT track. For both the series and parallel AC processing pickup, capacitive loading is presented on the track. This phenomenon is shown in Figure 8.1 with the equivalent capacitive loading. The reflected capacitive VAR load will be transformed by the LCL network to an inductive loading on the H-bridge, causing extra conduction and switching losses in the H-bridge. The losses are quite evident in the EV charging example of chapter 5 where system efficiency drops significantly at higher coupling conditions as the reactive VAR’s are much higher. To compensate for this inductive loading, an extra series capacitor and an AC switch could be added as shown in Figure 8.1. The AC switch will operate in a similar way to the AC processing pickup, however, this time it is used to change the equivalent capacitance to fully compensate for the inductive loading on the H-bridge. Using this method the overall efficiency of the system could be improved. Here, a simple series switched capacitor compensation technique is shown. However, there are many other possible methods of reactive compensation that could be used. A detailed investigation into what type of compensation technique is best for the application needs to be undertaken. In addition, future work needs be done to find the best place to add reactive components to the LCL network. Reactive compensation in other power supply topologies also needs to be considered.

![Figure 8.1. Primary LCL Power Supply with Series Switched Capacitor.](image)

8.3.3 Combined Primary and Secondary Power Flow Control

In this thesis, only the use of control on the secondary side using the AC processing controller has been investigated. This has presented significant advantages of efficiency
improvement on the secondary pickup. However, the reflected VAR load on the primary supply results in lower efficiencies. These effects are more evident when the coupling condition of the system is very high. One method of alleviating the problem is by combining primary track current control and AC processing control together. A primary track current controller will reduce the track current and the overall losses on the primary side. In addition, it is always good in terms of the system efficiency to generate the minimum amount of magnetic field by controlling the primary track current such that just enough power is delivered to meet the specifications. This combined control system would require a master control and a wireless communication channel to form the link.

### 8.3.4 Roadway Powering of Electric Vehicles using Multistage IPT System

Electric Vehicles (EV) are becoming more popular as developed nations try to reach their carbon dioxide emission levels. One common problem for current EV technology is the limited battery storage that only allows them to travel 100 – 300km. Although this capacity is sufficient for daily transportation to work, it is not enough to support long range transits from city to city on State Highways. Since battery recharging of EV’s takes much longer than refilling a gasoline car, there are pressing needs to power EV’s in motion as they travel along the highway. One possible method is to install charging pads along the road, enabling power to be delivered in motion. Each pad is individually driven by a small power supply. However, this system is costly and inefficient, since each pad requires its individual power supply and the cost of the system is increased. In addition, the traditional LCL power supply has an inherently slow transient response, incapable of controlling the primary current sufficiently quick enough. As such, the primary converters needs to power up in advance before the vehicle arrives, causing greater standing losses and lowering efficiency. A better option is to use a multi-stage IPT system as shown in Figure 8.2. In this system, one large primary power supply is used, but capable of supporting many in-motion EV’s. Next, a simple controller is closely coupled to the IPT track and used to drive each individual primary pad. Power is then transferred to the secondary pickup via the coupling between this primary pad and a
secondary pad. Although this system seems to be more complex, it has several advantages. Firstly, one large power supply is now used to support many moving vehicles and each intermediate controller is very simple, as such, the relative cost can be reduced. Secondly, synchronization of all the primary pads now becomes inherent as they are all driven by one primary supply and all current phases are referenced to the supply to reduce or eliminate power transfer between primary pads. Moreover, by using the AC processing controller on the intermediate stage, a fast transient response can be obtained which is ideal for powering moving vehicles.

A typical multi-stage IPT system is shown in Figure 8.3 where an AC controller as proposed in this thesis is used to power each primary pad ($L_{1s}$) at a suitable current upon demand. The intermediate stage composes of an AC processing controller in the front end driving a series tuned resonant tank. The series tuned tank has several advantages. The series resonant tank acts as a filter network to reduce the harmonics in the inductor current waveform in the pad. In addition, it reduces the peak switch voltage across the AC switches as the pad inductance is cancelled by the tuning capacitor. However, many aspects of the system need to be investigated before it can be adopted. One typical problem with any moving vehicle is that both the primary pad and pickup inductance can change due to the presence of ferrite. As such, the use of an AC processing controller in the intermediate stage requires further investigation to determine its ability to drive an off-tune reactive load. In addition, power transfer between the primary pads must be minimized in order to improve efficiency. Moreover, the transient response of the AC controller for this application must be improved given the response times required for powering vehicles travelling at >100km/h are stringent.
This type of multi-stage IPT system could be useful in many other applications beyond roadway systems. The ability to localize the magnetic field to reduce losses appear to have significant future potential but need much further study to understand and overcome practical limitations which could form the basis of separate Ph. D. studies in their own right.

### 8.3.5 Wireless Powering of Multiple Biomedical Implants

In this thesis, various medium-high power industrial applications using AC processing pickups have been demonstrated. However, the advantages that the AC pickup has on an IPT system are useful at any power level. Wireless powering of biomedical implants has predominately used primary side control because it has high efficiency. In addition, medical electronics require a high level of safety and secondary side control has largely been neglected because any failure would require surgery to fix it. However, if there is a need to have multiple biomedical implants inside a patient, primary side control cannot be used as the power required for each pickup may dynamically change – the use of an AC processing pickup in such an application is clearly advantageous given its high efficiency and capability to handle large variations in coupling.
Appendix A. Series AC Pickup Details

The series AC processing pickup outlined in this thesis is shown in Figure A.1. System specifications for the series AC processing pickup is shown in Table A.1.

![Series AC Processing Pickup](image)

**Figure A.1. Series AC Processing Pickup.**

<table>
<thead>
<tr>
<th>Spec. Parameters</th>
<th>Values</th>
<th>Spec. Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
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The schematic for the series AC processing pickup is shown in Figure A.2.
Figure. A.2. Series AC Processing Pickup Schematic.
Appendix B. Normalised Characteristics

The parallel AC processing normalized characteristics is captured in Table B.1.

Table B.1. Normalised Table for Parallel AC Processing Pickups.
Similarly, the series AC processing normalized characteristics is captured in Table B.2.

Table B.2. Normalised Table for Series AC Processing Pickups.

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References


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