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Improved Couplers for Charging Stationary and Moving Electric Vehicles

Mickel Bipin Budhia

For Mum and Dad

thank you for all the support.
**Abstract**

Inductive Power Transfer (IPT) uses a varying magnetic field to transfer energy without contact. The contactless nature of IPT makes it an ideal method in the context of Electric Vehicle (EV) charging. The driver no longer needs to remove a plug from an unsightly and vandal prone charging stand then “plug in” the vehicle – rather the charging is done invisibly and out of mind. Well designed magnetic couplers enable energy to be directed across an air gap to an EV with little loss. The focus of this thesis is to find the best topologies suitable for both stationary charging and dynamically powering an EV while on the move.

The performance of the couplers determines the overall feasibility and cost effectiveness of the complete charging system. Stationary charging is envisaged where the transmitting coupler is buried in the ground and a receiver is mounted underneath an EV, thus the driver simply needs to park over the transmitter. Couplers in a dynamic system need to transfer significantly higher power levels to a moving vehicle with adequate tolerance across the lane. Both applications are especially demanding because power levels of 2-60kW need to be transferred over 150-300mm air gaps with sufficient tolerance to horizontal misalignment, especially if automatic guidance is not desired. The magnetic field emissions also need to be controlled to ensure the systems comply with international guidelines and regulations.

The performance of circular shaped couplers is initially investigated as these are the most common in the literature and industry. Existing designs are optimised but still offer necessarily poor coupling due to the shape of their fundamental flux paths. Bar shaped couplers were investigated as an alternative due to theoretically better flux paths but these were found to produce a two sided magnetic field, one side of which must be shielded resulting in very inefficient systems.

A novel coupler topology called the Double D (DD) pad is proposed with a high single-sided flux path that enables it to outperform circular couplers. DD pads that are smaller and lower cost than circular couplers enable equivalent power transfer over an area five times larger when operating under the same conditions. Due to the prevalence of circular couplers, an interoperability investigation is done and shows that a DD receiver will work better with a circular transmitter than a circular receiver. Finally, DD pads are considered for dynamic powering and measurements taken on a laboratory scale system show this is a feasible option.
Acknowledgements

There are many people I would like to thank for their support and encouragement during this long and arduous journey to greater enlightenment.

Firstly I would like to acknowledge my supervisors Associate Professor Grant Covic and Professor John Boys. Grants’ suggestions and ability to endlessly proof read and improve draft papers have been a huge amount of help. John is a limitless source of ideas, wisdom and encouragement.

I would like to express my special thanks to David Huang, Michael Kissin, Frank Hao and Stefan Raabe for their assistance with laboratory experiments and/or for proof reading this thesis. In addition I would like to thank my colleagues from the Power Electronics group and friends: Craig Baguley, Jonathan Beaver, Zak Beh, Nick Keeling, Joshua Lee, Jerry Liang, Ganesh Nagendra, Jimmy Peng, Baljit Riar, Daniel Robertson, Edward van Boheemen, George Wang and Hunter Wu.

Finally, but by no means least, I would like to thank both of my parents, Bepin and Bharti for their support.

Mickel Budhia

26th February 2012
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Nomenclature

Definitions

Bipolar  Describes a track in which the receiver interacts with both the forward and return currents paths of any phase.

Unipolar  A track in which the receiver only interacts with the magnetic field from the forward current in any phase.

Acronyms

AGV  Automatic Guided Vehicle
ARPANSA  Australian Radiation Protection and Nuclear Safety Agency
BP  Bipolar pad
CNS  Central Nervous System
DD  Double D
DDQ  Double D pad with a spatial quadrature coil added
EMF  Electromagnetic Field
EV  Electric Vehicle
FEM  Finite Element Method
HEV  Hybrid Electric Vehicle
ICE  Internal Combustion Engine
ICNIRP  International Commission on Non-Ionising Radiation Protection
IPT  Inductive Power Transfer
MMF  Magnetomotive force
NI  Amp.turns
PHEV  Plug-in Hybrid Electric Vehicle
Rx.  Receiver
SAE  Society of Automotive Engineers
Tx.  Transmitter
(V)LF  (Very) Low Frequency
Symbols

\begin{align*}
C_1 & \quad \text{Primary (transmitter) tuning capacitor} \\
C_1 & \quad \text{Secondary (receiver) tuning capacitor} \\
C_d & \quad \text{Average coil diameter} \\
C_{DD/Q} & \quad \text{Tuning capacitors for the DD and quadrature coils of a DDQ receiver} \\
F_r & \quad \text{AC resistance factor} \ (F_r = R_{ac}/R_{dc}) \\
I_1 & \quad \text{Primary current} \\
I_{sc} & \quad \text{Short circuit current of the secondary coil} \\
k & \quad \text{Coupling coefficient} \\
L_1 & \quad \text{Inductance of the primary} \\
L_{1,so} & \quad \text{Inductance of the primary with the secondary open circuit} \\
L_{1,ss} & \quad \text{Inductance of the primary with the secondary short circuit} \\
L_2 & \quad \text{Inductance of the secondary} \\
M & \quad \text{Mutual inductance} \\
N_1 & \quad \text{Number of turns on the primary} \\
N_2 & \quad \text{Number of turns on the secondary} \\
\Phi & \quad \text{Magnetic flux} \\
p & \quad \text{Track pitch} \\
P_c & \quad \text{Core loss} \\
P_{out} & \quad \text{Output power of the IPT system} \\
P_p & \quad \text{Pad pitch} \ (\text{distance between DD transmitter centres in a roadway system}) \\
P_{su} & \quad \text{Uncompensated power output of the receiver coil} \ (\text{defined as } V_{oc} \cdot I_{sc}) \\
P_w & \quad \text{Winding loss}
\end{align*}
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Q_2$</td>
<td>Quality factor of the receiver tuned circuit</td>
</tr>
<tr>
<td>$Q_L$</td>
<td>Native quality factor of an inductor</td>
</tr>
<tr>
<td>$VA_1$</td>
<td>Volt-ampere of the primary</td>
</tr>
<tr>
<td>$V_{oc}$</td>
<td>Open circuit voltage of the secondary coil</td>
</tr>
<tr>
<td>$V_r$</td>
<td>Reflected voltage of the receiver onto the primary</td>
</tr>
<tr>
<td>$\delta_{Al}$</td>
<td>Distance between ends of ferrite cores and aluminium ring</td>
</tr>
<tr>
<td>$\delta_x$</td>
<td>Horizontal offset between the transmitter and receiver in the $x$-axis</td>
</tr>
<tr>
<td>$\delta_y$</td>
<td>Horizontal offset in the $y$-axis</td>
</tr>
<tr>
<td>$\delta_z$</td>
<td>Vertical separation between the transmitter and receiver</td>
</tr>
<tr>
<td>$\kappa$</td>
<td>Kappa (metric for comparing receivers on distributed systems)</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Operating frequency of the IPT system</td>
</tr>
</tbody>
</table>
Chapter 1. Introduction

1.1. Introduction

At present petroleum is the main source of energy for personal mobility however its continued use is not sustainable. An increasing global population is driving demand for personal transportation powered by the Internal Combustion Engine (ICE). In the early 1990s there were well over half a billion cars and trucks worldwide and it has been estimated that there will be more than 2.5 billion by 2050. The remaining petroleum reserves are not sufficient to support all of these vehicles. Peak oil is fast approaching and this occurs when the global rate of extraction reaches a maximum followed by an irretrievable decline. Analysts predict this will occur in the next 5-10 years and the total remaining supply of oil and gas will be exhausted in the next 40-60 years [1, 2]. These factors result in stress on the supply and demand balance and will put global stability at risk.

The current rate of petroleum consumption is adversely affecting the environment due to increased air pollution and greenhouse gasses produced as by products. Tailpipe emissions include hydrocarbons, nitrogen oxides, carbon oxides and particulate matter and these cause local, regional and global effects. Typical local effects are smog and respiratory diseases due to aerosol particles and nitrogen oxides. However, the greatest concern is global warming and this affects the entire planet [3]. In recognition of this, stringent regulations on emissions and fuel economy have been imposed on auto manufacturers.

However, due to the vast number of ICE vehicles in use, regulations will only achieve a slight reduction in greenhouse gas production. Pure Electric Vehicles (EV) are a practical long term option as they significantly reduce dependence on fossil fuels. Energy storage in an EV allows effective use of intermittent renewable energy sources such as wind and solar [4-6] and uptake of EVs has been increasing since the 1990s, however market penetration has been low because they are not as cost effective as conventional ICE vehicles. The battery is the last remaining technological bottleneck and it typically makes up 60-80% of the electric vehicle purchase price, which is around double that of an equivalent conventional vehicle.

If EVs are to gain widespread use, major improvements are required in battery life, cost, and grid connection. The latter option allows opportunistic charging after each trip rather than a long charge at the end of the day. This reduces battery wear significantly by minimising the
depth of discharge and the EV is more cost effective because a smaller battery is required. Studies have shown that opportunity charging allows a 47kWh battery to be reduced to 32kWh. This approach also reduces range anxiety, which is another impediment to EV penetration. At present, EVs are recharged with a plug and socket arrangement, as described in section 1.3, and this makes opportunistic charging impractical because user intervention is required. Successful opportunity charging requires numerous plug/unplug cycles during the day creating a significant source of driver anxiety. Enabling safe and convenient contactless EV charging without user intervention is the main motivation for this thesis however the ultimate goal is to achieve cost parity with conventional vehicles without any noticeable differences or loss of performance to the end user.

EVs can be almost as cost competitive as gasoline vehicles if they are both recharged and powered by the road and this is the second objective of this thesis. Battery capacity can be reduced even further and range anxiety is completely eliminated. The infrastructure for such a dynamic charging system is relatively small because highways typically make up only 1% of roadway miles but carry 22% of all vehicle miles travelled. An EV that has 50% of its driven miles connected to a dynamic charging system would be as cost effective as a conventional vehicle and does not incur additional gasoline costs. Here a significant proportion of EVs are always connected to the grid via a combination of stationary charging points and an electric road, to allow new ‘smart grid’ options to be explored resulting in significantly improved energy diversification creating truly sustainable personal transport [7].

1.2. A brief history of electric vehicles

EVs are not a new concept – they were invented in 1834, a time when horse drawn carriages were a prevalent form of transport. By the late 19th century technological development had reached a point where self-propelled road vehicles were possible. Inventors in Europe and North America were working on steam and internal combustion engines as well as EVs. Some notable achievements were the invention of the four-cycle combustion engine by Nicholas Otto in the late 1800s and the use of such engines by Gottlieb Daimler and Karl Benz in the first commercial automobiles in the late 1880s. Simultaneously, electric power generation and distribution were being developed and implemented by Thomas Edison and George Westinghouse. In the early 1900s, 800 of the 2,370 vehicles in the New York, Chicago and Boston areas were electric while 1,170 were powered by steam (noting these
borrowed mature technology from trains and ships) and the remaining 400 were gasoline powered [8-10].

Gasoline powered vehicles were noisy, odorous and unreliable and prone to mechanical failure however the worst drawback was that a strong arm was required for the dangerous procedure of hand cranking to start them. EVs were by comparison, silent, clean and simple to operate however they suffered from limited range and lower speeds. Lead acid batteries were in the early stages of development and required careful maintenance and charging. The slow rate of electrification meant finding places outside of cities for recharging was difficult and doctors and wealthy women were the target market for EVs due to their frequent short intra-city trips. The expansion of ac power distribution outside cities (in which Edison’s dc supply predominated) meant that an inefficient rotary converter with a current limiting rheostat was needed for recharging. Also there were no standardised plugs and sockets adding further inconvenience [8, 10].

Technological developments made by numerous manufacturers resulted in more reliable, powerful and fuel efficient gasoline engines leaving the last major problem – hand cranking. This was addressed by Charles Kettering and in 1911 he invented the electric starter motor. Conventional wisdom dictated that a large motor would be needed to turn an engine however Kettering realised that a small motor operating in severe overload conditions for a short time would be able to turn the engine. This invention resulted in the irreversible decline of the EV and the last models were produced in late 1920s [8, 10].

EVs were largely marginalised by car manufacturers for 50 years however they received renewed interest due to the energy crisis of the 1970s [2]. Between 1974 to 1977 a U.S. company called Sebring-Vanguard, Inc. produced 2300 small electric cars with power outputs of 2.5-6hp. In 1996 General Motors developed the EV1 however this was discontinued. Both these cars suffered from limited range and although battery technology has improved it still cannot compete with the extremely high energy density of gasoline (50 times that of a lithium-ion battery) [2, 9].

In consequence, Hybrid Electric Vehicles (HEV) were developed and these use a propulsion system comprising a conventional engine, an electric motor and a smaller lower priced battery. The vehicle derives its energy from the combustion engine which is used to directly or indirectly propel the vehicle and or recharge the battery depending on whether a series or parallel topology is used [11]. Regenerative braking is responsible for the main reduction in
fuel consumption; in this process kinetic energy is recovered and stored rather than being dissipated as heat as in a conventional vehicle [12]. However fuel economy is highly dependent on the driving cycle. If there is a considerable amount of ‘stop and go’ driving, the reduction in fuel consumption compared to an equivalent ICE vehicle can be as high as 50.6% however on a highway the reduction is only 5.2% [13]. As a result, Hybrid vehicles are not considered a sustainable long term option for personal mobility [14].

Plug-in Hybrid EVs (PHEV) such as the Chevrolet Volt have recently been introduced and these allow recharging from the grid. As such they can be operated as pure electric EVs for a limited range, however the overall vehicle efficiency is lower than that of a pure EV due to the extra weight of the engine [15, 16]. These vehicles are therefore also not considered to be a long term sustainable option.

Nowadays, the major driving force behind EVs is environmental awareness and several car manufacturers have introduced pure electric vehicles such as the Nissan Leaf, Mitsubishi iMIEV and Tesla Model S, but these still do not have a range equivalent to that of an ICE vehicle and therefore are more suited to urban commuting. These EVs rely on conventional contact based charging and standards are being set as described in the subsequent section.

1.3. Conventional EV charging

Using plugs with trailing cables from a charging stand is the most common method of charging EVs, however the power levels involved are significantly higher than available from commonly used mains plugs. This necessitates comprehensive plug and socket design and regulation to ensure public safety and interoperability between manufacturers. This has been addressed by the Society of Automotive Engineers (SAE) and the guidelines for conductive charging are encapsulated in the SAE J 1772 standard. The standard covers general electrical and physical requirements as well as communication protocols and performance requirements of the chargers. It is supported by Chrysler, Ford, General Motors, Honda, Nissan, Tesla Motors and Toyota [17].

The plug design specified by SAE J1772 is based on a design from a company called Yazaki and a compliant plug manufactured by Delphi inserted into an EV is shown in Figure 1-1(a). The 75mm square socket contains 5 pins as shown in Figure 1-1(b). The presence of the proximity detection pin is required in order to initiate the charging process – there is no ac
present at the power pins if the proximity pin has not made contact. The level of power flow is determined by communication via the data pin [17, 18].

Figure 1-1: (a) SAE J1772 compliant plug inserted into an EV and (b) pin layout in the socket.

The power levels considered in the SAE J1772 standard are presented in Table 1-1 [19]. The ac power levels have been finalised but the plug and socket need modification for high power dc fast charging. The levels of contactless power transfer considered in this thesis are chosen to fit within the finalised Level 1 and Level 2 ac standards.

<table>
<thead>
<tr>
<th>Level</th>
<th>Voltage</th>
<th>Power delivery</th>
<th>Rated current (A)</th>
<th>Rated power (kW)</th>
<th>Finalised</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC Level 1</td>
<td>120V AC</td>
<td>Single-phase</td>
<td>≤16</td>
<td>≤1.92</td>
<td>Yes</td>
</tr>
<tr>
<td>AC Level 2</td>
<td>240V AC</td>
<td>Single-phase</td>
<td>≤80</td>
<td>≤19.2</td>
<td>Yes</td>
</tr>
<tr>
<td>DC Level 1</td>
<td>200-450V DC</td>
<td>DC</td>
<td>≤80</td>
<td>≤19.2</td>
<td>No</td>
</tr>
<tr>
<td>DC Level 2</td>
<td>200-450V DC</td>
<td>DC</td>
<td>≤200</td>
<td>≤90</td>
<td>No</td>
</tr>
<tr>
<td>DC Level 3</td>
<td>200-600V DC</td>
<td>DC</td>
<td>≤400</td>
<td>≤240</td>
<td>No</td>
</tr>
</tbody>
</table>

Table 1-1 Charging Power Levels Specified by SAE J1772

The high power levels presented for Level 2 and Level 3 dc charging are a result of the present market assumption that a vehicle should receive replenishment of on-board energy storage from a stationary source in a similar manner to refuelling with gasoline. In response, industry has put effort into developing energy storage for fast charging systems. However, present technology means batteries are not capable of accepting such high charge rates without compromising either the energy density or the battery life (due to overheating). Therefore a lower charge rate for a longer time is preferred [20] and contactless opportunistic charging is an ideal method for charging while also substantially prolonging the life of the battery.

1 Images from http://delphi.com/manufacturers/auto/hevevproducts/charging-cordsets/charge-coupler-con-cable/
1.4. IPT background

Inductive Power Transfer (IPT) is a method that enables energy transfer between a galvanically isolated source and sink, and is based on phenomena described by Ampère’s and Faraday’s laws. A time-varying current in a wire creates a changing magnetic field and this is able to induce voltage in a loop of wire in the field. This is shown conceptually in Figure 1-2 where power is coupled across an air gap. Transformers work on the same principles however the coupling coefficient is greater than 0.95 due to the presence of a magnetic core linking the primary and secondary – in IPT systems the coupling coefficient may be as low as 0.01 [21].

![Figure 1-2: Principles of IPT.](image)

Ampère’s law given by (1.1) states that the line integral of $H \cdot dl$ around a closed path is equal to the current traversing the surface bound by that path. In this case the contour is the dashed line shown in Figure 1-2.

$$\oint_C H \cdot dl = I$$

(1.1)

Faraday’s law describes the electromagnetic induction that enables the transfer of energy to the sink. The induced voltage is referred to as the electromotive force ($V_{emf}$) and is given in (1.2) where $N$ is the number of turns of the coil and the flux ($\Phi$) is the integral of the normal component of the flux density over the surface area, $S$, of the coil as shown in Figure 1-2. The negative sign in (1.2) is described by Lenz’s law which states the current flow in the coil (when the sink is connected) will be such that it creates a magnetic field that opposes the field inducing it.

$$V_{emf} = -N \frac{d\Phi}{dt} = -N \frac{d}{dt} \int_S B \cdot ds$$

(1.2)

The physical phenomena described by Ampère and Faraday have been well understood for almost two centuries and indeed early attempts to evaluate practical contactless power systems were made. A patent for replacing the electrified third rail that supplies power to trains with an inductively coupled system was granted to Maurice Hutin and Maurice
LeBlanc in France in 1890 and America in 1894. Such systems are usually loosely coupled, so that only a small proportion of the flux created by the primary links with the secondary. This is illustrated in Figure 1-2 where only one of the six flux lines link the secondary coil. The inventors suggested several approaches to overcome the poor coupling:

- High frequency operation (2kHz). As shown by (1.2) a greater voltage is induced if the rate of change in flux is higher.
- Capacitive compensation on both the primary and secondary. These are used to tune the primary and secondary inductances allowing resonance that allows significantly greater power transfer as discussed in Chapter 2.4.
- The use of a permeable material such as iron to improve the flux linkage.

Power flow control methods were also presented where several switched coils were used on the secondary – more coils could be switched off if the power demand was high [22].

IPT reached the point of practical implementation in relatively recent times and this is a result of four key technological developments:

- Semiconductor switches that enable operation at 10-40kHz, and beyond.
- Modern low loss capacitors capable of handling high currents and enabling tuning.
- Ferrimagnetic materials that have low hysteresis and eddy current loss and can handle relatively high flux densities in the frequency range of interest.
- Litz wire consisting of individually insulated strands of wire (in the order of 0.1mm diameter) wound in bundles that are in turn bundled and twisted around each other. This results in effective copper utilisation as skin effect and eddy currents losses are reduced.

IPT systems offer several advantages and this has resulted in a wide range of applications [23]. The main advantages are electrical isolation and the elimination of exposed conductors that improves safety and enables operation in hazardous environments such as high humidity or the presence of inflammable gasses. Equally, operation in extremely clean environments such as IC manufacturing is preferable as the systems do not produce any residue [24, 25]. IPT is also suited to applications where physical connections are not viable such as biomedical implanted devices [26] or where such connections are not desired as in EV charging applications [27-30]. The solid state nature of IPT systems ensures extreme robustness – there are no moving parts. Conventional techniques for delivering power to
moving objects such as trains or bogeys on a monorail in materials handling equipment have relied on brush and bar or slip ring contacts, and festoon arrangements where only limited travel is required. Well designed IPT systems are inherently more durable than any of the conventional approaches resulting in far lower maintenance costs [29].

1.5. Industrial applications of IPT

IPT systems can be grouped into two general categories according to the power transferred:

- Those that allow power transfer along a long path, these are referred to as distributed based systems.
- Those that only allow power transfer in discrete locations, these are referred to as lumped systems.

Both types of systems are presented here in an industrial and consumer context.

1.5.1. Distributed systems

1.5.1.1. Monorail

One of the largest materials handling company in the world is Daifuku Co., Ltd of Japan and two of their IPT based products are shown in Figure 1-3(a) and (b). The system shown in Figure 1-3(a) is a prototype designed for transporting sheet materials. The primary is called a track and it is encapsulated in white conduit. The secondary comprises a coil which is wound on a ferrite core for improved power transfer and it is referred to as a pick-up. In these systems movement is allowed along the track but the secondary is tightly constrained and stays a fixed distance from the conductors.

![Figure 1-3: (a) Materials handling system and (b) vehicle assembly line.](image-url)
The systems shown in Figure 1-4(a) and (b) are also manufactured by Daifuku and are designed to transport stacks of 300mm diameter silicon wafers in a clean room environment. IPT is used to power bogeys in 80% of the clean rooms built in the last few decades due to its inert nature. The track in Figure 1-4(a) is in the order of 10s of meters long and is shown powering five bogeys. A close up view of a bogey is shown in Figure 1-4(b) and the Front Opening Universal Pod (FOUP) is lowered. The FOUP can carry up to 25 wafers and weighs around 9kg.

Conductix-Wampfler (formerly Wampfler AG) and Paul Vahle GmbH & Co. KG are other companies who have also released monorail based IPT systems. While the tracks are similar, the pick-ups vary and small U-shaped and large E-shaped types are shown in Figure 1-4(a) and (b) respectively. These are the magnetic components, an electronic power flow controller is needed in addition, as discussed in Chapter 2. The U-shaped pick up is designed to couple 300W and measures 150mmL by 73mmW by 95mmH and weighs 1.26kg. The large E pick-up is able to couple up to 10kW nominally and 20kW peak, and it measures 250mmL by 420mmW by 322mmH, and weighs 29kg. The systems operate at 20kHz and the largest power supply that Paul Vahle produces has a capacity of 45kW [31].
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1.5.1.2. Automatically Guided Vehicles (AGV)

Conductix-Wampfler and Paul Vahle have developed IPT systems for Automatic Guided Vehicles (AGV); these also use a track however the pick-up sits above the track rather than within it (as is the case with the types shown in Figure 1-5). Systems used in BMW and Audi assembly plants are shown in Figure 1-6(a) and (b) respectively. AGVs are loosely constrained compared to monorail systems therefore require some horizontal tolerance about the track as shown in Figure 1-6(b). The Paul Vahle system can transfer full power with horizontal offset either side of the track of 25mm [31]. In these systems IPT offers reduced maintenance costs and higher reliability when compared to traditional chain driven platforms.
1.5.2. Lumped systems

1.5.2.1. Battery charging – high power

In order to extend the range of an electric bus, Conductix-Wampfler developed a 30kW IPT system for installation at bus stops. The battery is fully charged overnight at the depot and there are several stops where the battery is partially charged as illustrated on the graph of Figure 1-7(a). The concept of the systems is shown in Figure 1-7(b) where the secondary coil in mounted in the chassis of the bus. The primary coil is shown in Figure 1-7(c) and is shown embedded in concrete in Figure 1-7(d). The system was first installed in Genoa, Italy in 2002 and due to its success another system was installed in Turin, Italy in 2003.

A single coil (shown in Figure 1-7(c)) can deliver 30kW however the bus required 60kW necessitating two primary and two secondary coils (the outline of two coils can be seen in Figure 1-7(d)). The nominal air gap between the primary and secondary is 50mm with a vertical tolerance of ±10mm. Therefore, a mechanical apparatus is used to raise and lower the secondary as required to ensure the ground clearance of the bus is not compromised.

Further work on bus charging has been done by Waseda University of Japan in 2010 [32]. This bus is smaller than the Italian one above, has an unladen weight of 3.3t, and is propelled
by a 50kW synchronous motor capable of providing 240Nm of torque. The 293V lithium-ion battery has a capacity of 12kWh and maximum charge rate of 5C (60kW). The IPT system developed can transfer 30kW over a 100mm air gap (but an actual mechanical clearance of 84mm due to packaging) with 92% total efficiency. The primary and secondary of the coupling apparatus weigh 35kg each and are 847mm square by 33mm thick, one of the prototypes is shown in Figure 1-8(b). The couplers were originally over 50kg each but the internal ferrite structure was optimised to reduce weight. Again, the objective of that work was to extend the range of the bus, which is used on a university campus, by enabling opportunity charging as illustrated by the graph of Figure 1-7(a).

The Korea Advanced Institute of Science and Technology (KAIST) has recently developed a dynamic charging system that enables power to be continuously coupled to a moving vehicle [33-35]. The primary consists of a long track wound across ferrite ‘poles’ as shown in Figure 1-9. The winding nature of the primary track means that alternating north and south poles are created along the length of the 100mm wide track. The secondary consists of two coils and the flux enters one and exits the other resulting in induction. While effective, the down side of this system is that it is necessarily expensive due to the intensive use of ferrite for flux path shaping in the primary (the ferrite in the secondary structure is not shown).
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Optimisation procedures undertaken by Waseda University cannot be practically applied to the primary designed by KAIST shown in Figure 1-9. The volume of ferrite must be maintained – a track current of 200A is used and if the ferrite is not thick enough it will saturate. The KAIST system can transfer 35kW over an air gap of 200mm with a total lateral displacement of 240mm. The maximum system efficiency is 74% and this occurs with 27kW of power transfer.

1.5.2.2. Battery charging – low power

IPT is also a practical method for recharging consumer devices such as phones and laptops and considerable research has gone into miniaturising and simplifying IPT systems to improve cost efficiency. The main advantages are convenience, reduced cable clutter and reduced eWaste. A typical system is shown in Figure 1-10(a) where a mobile phone is placed on a charging stand and held in place by permanent magnets.

![Figure 1-10: (a) Dedicated IPT charger, (b) charging mat and (c) Qi compliant mat allowing a large range of devices to be charged.](image)

However, the trend is to make a more versatile system based on a charging mat allowing several devices to be charged simultaneously as shown in Figure 1-10(b). The Qi standard was released by the Wireless Power Consortium in 2010. This group consists of several independent companies who are aiming to make a universal charging platform as shown in Figure 1-10(c) [36]. This means that only one platform is required for a large range of
devices and new devices only need to be purchased as the old ones become unfashionable – the charger is the same.

1.6. Motivation for this research

Plugs and sockets that are SAE J1772 standard compliant are considered very safe due to features such as proximity detection and shrouded pins. However, the plug and socket ultimately need physical connection and once this inconvenient process is done, the car is tethered to its station. This approach means that opportunistic charging every time the car is parked is a major inconvenience and cannot be implemented in places like taxi stands or traffic lights. There are several other safety and security issues with contact based charging of vehicles parked in public places. These include:

- Plugs, cables and charging stations being damaged by vandals.
- Plugs being removed by ill-natured people for fun [37].
- Ice build up in the plug in regions where snow is common such as Canada or Northern Europe.
- The plug requires maintenance due to wear and tear. The J1772 standard stipulates plugs must have a cycle life of 10,000 times. This corresponds to 27 years if charged once per day – if opportunity charging requires several plug/unplug cycles a day the plug will wear more quickly.
- Trailing cables present a trip hazard in car parks.
- Having numerous charging stations can have a negative visual impact.

1.7. Objectives of this thesis

This objective of this thesis is enable cost effective contactless power transfer to EVs. Charging can be done at home or in car parks in order to extend the range of EVs by enabling opportunity charging. However, the ultimate goal is realising an unlimited range and this can be achieved by dynamic charging. This thesis will focus on the design of the magnetic couplers that enable the contactless power transfer. The design of these is critical to ensuring the overall feasibility and cost effectiveness of the entire system. A universal secondary coupler is sought that allows both stationary and dynamic charging.

In the designs presented above, bus charging was achieved with a relatively small air gap and mechanical assistance or electronic guidance was required making the system less reliable
and more expensive. In this thesis couplers that can operate with an air gap of 125-300mm (300mm if the primary is buried) are required. The design is targeted at EVs but it also needs to be scalable for SUVs and buses. The horizontal tolerance needs to be sufficient to enable a driver to park without alignment assistance. In order to achieve these objectives, the work is split into several smaller objectives:

1. Analyse existing couplers in the literature and industry and determine what improvements can be made.
2. Propose new coupler topologies and determine their limits i.e. are they also suitable for dynamic charging.
3. Optimise the selected coupler topology to minimise resource use while maximising coupling.

1.8. Outline of this thesis

The following list briefly explains the contents of each chapter:

- Chapter 2 introduces the basic components and concepts of IPT systems. As this thesis only focuses on the magnetic design of couplers, an emphasis is made on the importance of having a high coupling coefficient. Magnetic field exposure guidelines are introduced and couplers must comply with these if they are to achieve mass deployment.

- Chapter 3 contains a review of the most common couplers used in lumped systems found in industry and the literature. Circular couplers, called power pads are then developed and optimised to reduce material requirements while maintain relatively high coupling coefficients. Some practical design considerations are presented when using the optimised pads in an IPT system followed by compliance with magnetic field guidelines.

- Chapter 4 looks at the fundamental flux paths produced by circular pads. These are found to be non-ideal and show that the pads need to become impractically large to achieve good coupling with ~200mm+ air gaps. Flux paths of bar type pads are investigated and found to enable good coupling. However, the bar pads produce flux out the front and back of the coupler therefore cannot be used in practice if high efficiency is desired. The Double D topology is presented and this produces a single sided flux while enabling relatively high coupling coefficients.
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- Chapter 5 looks at optimising DD couplers by investigating their coupling limits by making them very long or very wide. Subsequently, both length and width dimensions are varied resulting in a 2D design surface, and optimal pads are chosen by comparing their coupling coefficients as a function of their area.

- Chapter 6 shows that DD pads are polarised and therefore have better tolerance to lateral misalignment in one direction (despite this they are still far better than circular pads). A spatial quadrature coil is added and then optimised to improve the size of the power transfer zone. Power loss (copper and core) in DD pads is investigated via simulation and measurement. An interoperability investigation is undertaken where the DD pad is used with a circular transmitter. Finally, compliance of a 2kW DD pad system with magnetic field exposure guidelines is investigated and the stray fields are compared to that of a similar sized circular pad charging system.

- Chapter 7 presents two types of dynamic power transfer systems. The first is based on a unipolar transverse three-phase track and this is shown to have several advantages over conventional track designs. The second approach is a scoping exercise where DD primary pads are used to make a section of roadway and a power profile from a DD secondary is obtained by measurement. Subsequently, primary pad spacing is investigated via simulation.

- Chapter 8 summarises the main achievements of this thesis and suggests future work. A unique mode of operation of a pad topology derived from the DD topology has been recognised by the author and the pad is called a bipolar receiver. A few key results are presented followed by recommendations for further work.
Chapter 2. Inductive power transfer concepts

2.1. Introduction

In this chapter the basic components of IPT systems are explained. The modular nature of IPT systems permits independent investigation of each of the components. Firstly, a system overview is given followed by a closer look at the magnetic coupling apparatus that enables the contactless power transfer as these are the focus of this thesis. Coupling systems may be broken down into two general categories namely; lumped and distributed. The former allows power transfer at discrete points while the latter enables power transfer along a line. Both types of coupling systems are found in the literature and are presented in the context of charging either stationary or moving electric vehicles.

Subsequently, the approach used to measure the coupling in both lumped and distributed systems is presented. Designs that enable good coupling mean the demand on the electronics is not too great permitting cost effective solutions. Poor coupling will result in expensive solutions that can make widespread EV charging infeasible. Finally, magnetic field exposure guidelines are presented. Compliance with these guidelines is imperative if such IPT EV charging systems are to be implemented. The guidelines have been established to prevent adverse health effects caused by exposure to a time varying magnetic field.

2.2. IPT system overview

An IPT system consists of three main components; these are a primary power supply, a magnetic coupling apparatus and a secondary power flow controller as shown in the system layout of Figure 1-2. The source of input power varies depending on the specific application – it may be 12-24V dc automotive rails or a three-phase line for high power requirements. Similarly, the dc output voltage of the secondary power flow controller also varies with the application; charging small mobile electronic devices typically requires less than 5V while industrial applications may require over 500V. The main scope of this thesis is recharging EVs in a home environment therefore the most likely source is the single-phase 230V mains and this can provide 2kW from a standard socket. Larger capacity sockets, similar to those used for stoves, can be installed if shorter charge times are desired.
Regardless of power transfer capacity, all IPT systems are generally modular in nature and consist of similar main components. These can be further decomposed into sub-components for individual design and analysis as long as the intersection layers between these sub-components are maintained. This thesis investigates the magnetic coupling apparatus only and the intersection is considered to be the current ($I_1$) produced by the power supply as shown in the centre of Figure 1-2.

The operational frequency range of IPT systems is typically 10-150kHz however the specific frequency depends on the particular application. Systems for small electronic devices typically operate at higher frequencies (greater than 100kHz) to ensure the coupling apparatus is as small as possible for increased cost efficiency given the power levels are normally also low (<10W). Higher power industrial applications and EV charging is done in the VLF range (10-50kHz), this is mainly due to the limits of presently available semiconductor switches, voltage safety limits and magnetic field exposure guidelines, these are described in further detail in section 2.5 [24, 38-41].

In EV charging applications the power supply typically uses a voltage-sourced full-bridge inverter with four reverse conducting switches as shown in Figure 2-2(a). The inverter output voltage is stepped due to the alternate switching of the $S_{1+/−}$ and $S_{2+/−}$ switch pairs. Care must be taken to ensure the switches in a leg, formed by either $S_1$ or $S_2$, are not turned on at the same time or shoot-through will occur. This “care” is achieved by driving one switch in a leg with an inverted signal. The drive signals to each leg are square waves with 50% duty cycle at the desired operational frequency of the IPT system as illustrated in Figure 2-2(b). The phase shift ($θ$) between the drive signals to each leg determines the magnitude of the fundamental component of the output voltage $V_b$ [42].

Compensation networks are used on both the primary and secondary sides as shown in Figure 1-2 and these networks enable resonance allowing significantly greater power transfer as well as filtering the stepped output of the inverter reducing RFI. Compensation on the primary
side allows current $I_1$ in $L_1$ to resonate and the large reactive current creates a greater flux density around the inductor than if only the output of the supply inverter was used to drive the inductor directly. Power transferred is proportional to the square of the flux density. In this thesis the primary inductor is usually excited with an rms current of 23A but the structure of the inverter and tuning topology is such that the switches only have to conduct the load current to provide the real power (and some losses). Typical switch currents are in the order of hundreds of milliamps when the system is resonating with $I_1 = 23$A but not supplying power and the switch current rises to approximately 10 to 15A when the system is supplying a full 2kW to the load. A similar compensation technique is used on the secondary inductance to improve power transfer and is described in section 2.4.

![Diagram](image_url)

Figure 2-2: (a) Full-bridge power supply inverter and (b) phase shift control of inverter resulting in a square wave output voltage.

### 2.3. Magnetic coupling apparatus

Power is transferred from the primary side to the secondary side via the magnetic coupling apparatus and its design is critical to ensuring the feasibility and efficiency of an IPT system. This coupling needs to be made as large as possible to ensure efficient power transfer yet the
absolute minimum amount of materials should be used to ensure cost efficiency. These conflicting requirements make magnetic design challenging as a large number of solutions are yielded however the most optimal options are couplers that are also light, thin, robust and easy to construct.

As discussed earlier, the types of magnetic coupling apparatus can be grouped into two broad categories, these are: lumped and distributed. Both types consist of a primary inductance ($L_1$), a secondary inductance ($L_2$), and the mutual inductance ($M$) between these is responsible for the power transfer. A lumped system is here assumed to be based on similarly sized primary and secondary coils and power can only be transferred when these coils are closely aligned and have sufficient mutual inductance. Ferrite is typically used to enhance mutual coupling as it offers a relatively low reluctance path and enables field shaping.

In a distributed system the primary is an elongated coil, which is referred to as a track, and one or more secondary coils (typically wound on a ferrite core) can couple to a small portion of the track and provide power to loads. In the context of recharging stationary EVs the lumped topology is considered more suitable than the distributed type given vehicles are generally parked in known fixed locations, for example, parking lots, taxi ranks and garages. Evaluating and designing suitable magnetic couplers that can meet modern specifications is the focus of Chapters 3-6 in this thesis. Distributed systems are naturally suited to dynamic charging where a vehicle is powered as it travels along a length of road with an embedded track. This is discussed in Chapter 7.

### 2.3.1. Coupling coefficient

A large range of coupling apparatus have been designed for specific systems with varying levels of power transfer making comparisons between topologies difficult. The total power output is due to several electrical parameters such as the current in the primary inductance, operational frequency and the effect of the compensation networks on both sides.

The coupling coefficient ($k$) is commonly used in transformer and induction motor design, and is used here for lumped systems in order to decouple the electrical and magnetic characteristics. The coupling coefficient $k$ indicates the proportion of flux linkage between the primary and secondary to the total flux produced by the primary and here ranges from 0 to 1 as expressed by (2.1). This parameter is used throughout the thesis to compare coupling topologies and find optimal designs for a given topology. The key to ensuring system
feasibility and efficiency is to achieve a suitably high coupling coefficient with a given operational air gap between the magnetic couplers with minimal size and material cost. Couplers with \( k > 0.6 \) are approaching a transformer disallowing resonance, however achieving such high coupling coefficients with an air gap of 200mm will require couplers that measure over 1.3m square (details presented in Chapter 5). These are very large, resource intensive and expensive thus a balance between \( k \) and ultimately cost must be struck.

\[
k = \frac{M}{\sqrt{L_1 L_2}} \tag{2.1}
\]

The secondary coil in a distributed system is usually much smaller than the primary and therefore only couples with a small portion. This results in very small coupling coefficients (~0.01) and the value also depends on the design of the track making comparing different secondary couplers difficult. In recognition of this, a new term called kappa (\( \kappa \)) has been proposed as it allows a direct measure of the magnetic coupling efficiency of various secondary structures [43, 44]. In a distributed system with only one secondary that is able to couple with the entire length of the track, \( k \approx \kappa \), however this type of design is unlikely in practice. Kappa is calculated as shown in (2.2) where \( N_1 \) and \( N_2 \) are the number of turns on the primary and secondary respectively.

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

In a lumped system a \( k \) of more than 0.2 is preferred to ensure good system efficiency whereas in a distributed system coupling coefficients of 0.01-0.1 are not uncommon. However, even with such systems the magnetic coupling efficiency of the secondary structure can be high given typical values of \( \kappa \) range from 0.35 to 0.95 [44].

2.3.2. Lumped couplers

Lumped couplers may be further broken down into either closely coupled or loosely coupled types as explained in the following subsections.

2.3.2.1. Closely coupled

Closely coupled lumped systems operate with relatively small air gaps where the user typically has to “plug-in” the primary. This was the case with charge paddles used in an early electric vehicle as shown in Figure 2-3(a) [45, 46]. The structure of the magnetic components of the primary and secondary is shown in Figure 2-3(b). The primary comprises a coil wound
around a ferrite puck and the secondary consists of two coils wound around shortened rounded centre legs of two modified E cores glued together. The main mutual flux path between the couplers is shown in Figure 2-3(c). The charge paddle system has the advantage of isolation however the issues (listed in Chapter 1.6) are still similar to that of conventional plug-in charging systems therefore this topology is not considered suitable for further investigation in this thesis.

![Image](image1)

Figure 2-3: (a) Charge paddle, (b) Magnetic structure with primary partially inserted and (c) main flux paths in a closely coupled system.

### 2.3.2.2. Loosely coupled

Loosely coupled lumped systems operate with a large air gap and require no user intervention, and are the subject of investigation in this thesis. A complete envisaged domestic EV charging system is shown in Figure 2-4(a). The power supply and compensation network is mounted on a garage wall while the primary, hereafter referred to as a transmitter (Tx.), is buried in the ground. The secondary, which is referred to as a receiver (Rx.), is mounted upside down on the EV chassis so it is facing the transmitter as shown in Figure 2-4(b).
Traditional couplers used in loosely coupled lumped systems are intuitively based on common magnetic core topologies. The most common are pot cores [47-50] and U-shaped cores [51, 52] and these are shown in Figure 2-5(a) and (b) respectively. Such designs are not recommended for EV charging as they are necessarily thick and therefore compromise ground clearance or require extensive chassis modification. Also, the coupling coefficient is highly dependent on the area through which mutual flux passes as shown by (2.3) where the surface, $S$, is shown in Figure 2-5(b). Parking unguided EVs requires large horizontal tolerance (~300mm) making these coupler designs infeasible as they are smaller than the tolerance range. As shown in Chapter 3, small couplers usually operate with a small air gap to ensure the necessary coupling and this can result in highly constrained and sensitive systems.

$$M = \frac{N_2}{I_1} \int_S B_1 ds$$  \hspace{1cm} (2.3)

Circular coupler designs, which are derived from pot cores, have been developed in recognition of the limitations mentioned above. The ferrite structure, shown in Figure 2-5(a), has been flattened to form a disc and planar circular coils are used in order to increase the surface area through which mutual flux passes and to make the coupler thinner. Most circular

\footnote{Image from: www.arup.com/News/2011_03_March/HaloIPT_RollsRoyce.aspx}
coupler designs found in industry and the literature use ferrite discs, annuli or wedges [32, 49, 53-56] however, using such large pieces of ferrite necessarily makes the designs fragile and expensive. Large pieces of ferrite are prone to cracking with the shock and vibration experienced on a vehicle and careful mechanical design is therefore required increasing cost. Cracking over time causes changes in pad inductance and the compensation networks are no longer matched resulting in inefficient operation.

In this thesis, circular couplers called power pads are investigated and these use small ferrite bars arranged radially as shown in Figure 2-6. The cores are held in place by a shock absorbing coil former (not shown). The aluminium ring and backing plate make the pads very robust and also shield the surrounding area from stray magnetic field as described in Chapter 3.7. In larger power pad designs each of the ferrite strips are made up of several bars rather than a long single bar, to allow the pad to flex and twist without breaking any ferrite. This approach allows cost effective power pads that are more robust and lighter than designs using solid ferrite discs. A compromise between the volume of ferrite and coupling is presented in Chapter 3 where power pads are optimised.
Couplers based on coreless coils are shown in Figure 2-7 and several designs are presented in [38, 54, 57]. The advantage of such couplers is that the inductance of the transmitter is constant and independent of the position of the receiver, likewise the inductance of the receiver is also constant and both coils can be easily tuned to resonance. As discussed in Chapter 3.6, the inductance of power pads especially using ferrite varies as they are moved relative to each other. However, if the operational movement is constrained the effect on tuning is not too large. Coreless coils are generally not suitable for high power applications where ferromagnetic or conductive (i.e. aluminium) materials are in close proximity to the system due to eddy current loss. Field shaping with ferrite constrains flux to desired paths improving the coupling coefficient and consequently preventing excessive energy loss in surrounding materials due to leakage magnetic field, which also needs to be considered for safety reasons [58].

In order to compensate for the lower coupling coefficient prevalent with coreless coils and to maintain good efficiency, these systems are normally operated at frequencies in the range of hundreds of kHz to 10s of MHz [59, 60]. Efficient operation at such high frequencies is not possible with high power systems due to performance limitations of current semiconductor devices as mentioned in section 2.2. Systems that have relatively large air gaps in relation to the size of the transmitter as shown in [60] are generally only suited to low power applications where efficiency is not a major issue such as sensor networks or powering small swarm robots.

2.3.2.3. Coupling arrays

Transmitters comprising arrays of hexagonal coils with ferrite backing have been used to transfer low power levels to any location over a large area [40, 61]. Such a system is useful for charging several small electronic devices such as mobile phones and MP3 players when they are placed on a “charging mat”. The structure of the primary is shown in Figure 2-8(a),
where individual transmitter coils are activated to enable detection and localisation of the receiver.

With the structure shown (Figure 2-8(a)) the receiver can largely couple with one, two or three transmitter coils resulting in non-uniform power over the mat. As a result, a multilayer transmitter coil array was developed consisting of three layers of transmitter coils as shown in Figure 2-8(b). However, as shown in [62], each row or column of coils needs to be switched to prevent field cancellation in the centre of the multilayer transmitter. These methods are impractical with high power systems given each coil needs to be controlled independently adding significant switch loss due to the large currents used.

2.3.3. Distributed systems

Distributed system topologies vary widely and can be configured for systems that only permit a relatively small amount of movement across the track. With more advanced receiver structures, relatively free movement is possible. In the following subsections track topologies are presented followed by receiver topologies.

2.3.3.1. Single Phase tracks

The simplest track topology is the single phase loop type as shown in Figure 2-9(a) [63, 64]. The track is considered to be unipolar because the receiver is only exposed to a single track conductor, and the current in it, at any time. This topology is not commonly used in industrial systems because the inductance is large due to the large area encompassed requiring very large terminal voltages to obtain the desired track current. Also, the magnetic field about the track is not well constrained – there is a net current through the contour shown in Figure
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2-9(a) resulting in a stray magnetic field. This makes compliance with magnetic field exposure guidelines and regulations, discussed in section 2.5, difficult.

The parallel-cable or single phase bipolar track is a more popular topology and is shown in Figure 2-9(b). The track is considered to be bipolar because the receiver is exposed to both forward and reverse currents from the power supply. This track is able to power several receivers over a long linear distance with a small amount of horizontal tolerance at the designed operational air gap. Stray magnetic fields are less of a problem – there is no net current through the contour shown in Figure 2-9(b) therefore (in theory) there is no magnetic field [58]. The single phase bipolar track is typically used in materials handling systems that have multiple bogeys on a monorail and the track is in the order of tens of metres long [65-67].

![Figure 2-9: Single phase tracks (a) simple loop unipolar track and (b) bipolar track with three receivers.](image)

### 2.3.3.2. Distributed tracks

The single-phase bipolar track does allow some horizontal tolerance as illustrated by the shaded zone in Figure 2-9(b) however this is typically not sufficient for vehicles guided by human operators and necessitates high accuracy tracking for AGVs. Tracks that allow power transfer with more lateral tolerance (across the track) are sought. Two topologies found in the literature are the distributed-cable track and the meander-coil track [68-70] as shown in Figure 2-10(a) and (c) respectively. The power transfer zone with both of these track topologies is wider than possible with the bipolar track in Figure 2-9(b) however, with the
Chapter 2 – Inductive power transfer concepts

distributed-cable track the current in each conductor is $I_1/3$ resulting in significantly lower power transfer.

The directions of currents in the distributed-cable track taken on section line AA’ are shown in Figure 2-10(b). The receiver for a distributed-cable track is only exposed to unidirectional currents from the power supply and it uses a horizontal coil, which is named as such because it couples horizontal flux. There is a power null when the receiver is placed in the centre of the track as here the flux generated is vertical but the coil is designed for horizontal flux capture (flux directions are further explained in the following sub-section). A distributed-cable track is also difficult to create in practice because the differing inductances of the individual loops, which are connected in parallel, will lead to uneven current sharing among the conductors resulting in an inconsistent power profile.

The meander-coil track of Figure 2-10(c) uses two parallel conductors wound back and forth to cover a large area. The individual track conductors are effectively in series, and the current is uniformly distributed across the power transfer zone. As each conductor is carrying $I_1$, the flux density about the track is favourably high. However, the disadvantage of this topology is that the long track length results in a high inductance requiring large driving voltages.

The positioning of the conductors creates differing current directions, as shown in Figure 2-10(d). This results in flux vectors that continuously vary between horizontal and vertical laterally across the track (only changing in magnitude with time) and this results in many null points as the receiver is only sensitive to flux in one particular direction (discussed further in section 2.3.3.4).

Poly-phase tracks are presented in Chapter 7 of this thesis and these address the issues with the existing topologies described above. Poly-phase tracks also consist of several conductors forming individual loops in the power transfer zone. However, these loops are driven with currents that are sequential in phase and are not an integer multiple of 180°.
2.3.3. Tightly constrained receivers

There are numerous types of receiver topologies used on single phase tracks but due to the varied applications they can be grouped into two broad categories: tightly and loosely constrained. Tightly constrained receivers are used in applications where there is essentially no movement across the track such as monorail materials handling systems [58]. This allows the receiver to be placed within the track enabling a higher kappa. A few common topologies are shown in Figure 2-11(a)-(d). The first three couplers in Figure 2-11 referred to in order are: U-core, H-core and E-core [71-76]. The S-core of Figure 2-11(d) is an improved non-standard core shape that has recently been proposed. The S-core has been shown to provide a significantly higher $\kappa$ than the other topologies shown in Figure 2-11 [77]. These tightly constrained couplers are however, not suitable for RPEV applications and are not considered further in this thesis.
2.3.3.4. Loosely constrained receivers

The receiver topologies shown in Figure 2-12(a) and (b) are commonly used in AGV applications where the track is buried in the ground and horizontal and vertical movement is required along the direction of travel [78-80]. The receiver of Figure 2-12(a) is referred to as a flat bar while Figure 2-12(b) is referred to as a flat E-core receiver.

As mentioned earlier, single phase tracks are the simplest topology however they offer limited horizontal tolerance and this necessitates more complex receiver topologies than the two shown above in Figure 2-12. A simulation showing a magnetic flux density contour and flux lines in a cross-section of a single phase track with a pitch of 200mm is shown in Figure 2-13(a). The lightly shaded square shows the operating zone of the receiver and the magnetic field direction within the zone, and this varies with position. The magnetic field direction is vertical midway between the conductors however it is horizontal above either of the conductors (and opposite in direction).

The contour plot of Figure 2-13(b) shows the $x$ component of the magnetic flux density with reference to the axes to the right of the figure. The green colour shows a flux density of 0mT while red shows a high positive flux component ($+x$) and purple shows a high component in the $-x$ direction. A schematic of a flat bar receiver is added to Figure 2-13(b) and in the position shown, the horizontal flux capture is at a maximum therefore the coil is referred to as a horizontal coil. The $z$ component of the magnetic flux density is shown in Figure 2-13(c)
and the coil has been rotated by $90^\circ$ to enable vertical flux capture. The coil is therefore referred to as a vertical coil.

![Figure 2-13: (a) Magnetic flux density contours and flux lines in a cross-section of a single phase track, (b) $x$ component of the flux density with a horizontal flux receiver and (c) $z$ component of the flux density with a vertical flux receiver ($I_1 = 200\, \text{A}$).]

If the horizontal coil is added to the flat E-core as shown in Figure 2-12(b), it forms a quadrature receiver as shown in Figure 2-14 [81]. The output of the orthogonally wound coils is summed by a quadrature power flow controller enabling significantly greater horizontal tolerance as coupling is possible with either magnetic field component.
The vertical flux produced by both conductors adds in the centre resulting in a vertical component shown Figure 2-13(c) that is stronger than the maximum horizontal component shown in Figure 2-13(b). This generally results in a slight peak in the power profile formed when the receiver is moved across the track. Note the flux patterns shown in Figure 2-13(b)-(c) change when an actual ferrite receiver is added (due to its relatively low reluctance) therefore total output power is reasonably smooth during the transition between the output from the vertical and horizontal coils. In practice additional ferrite can be placed on the ends of the quadrature receiver or in the centre to further balance the power profile [80, 82].

![Figure 2-14: Quadrature receiver comprising horizontal and vertical coils.](image)

### 2.4. IPT receiver

Power transfer in an IPT system is achieved through mutual coupling in a similar manner to a transformer. Due to the large operational air gap however, the leakage inductance is high resulting in coupling coefficients that are typically in the range of 0.1-0.4 (in lumped systems) whereas $k$ is over 0.95 in a conventional transformer [21]. Idealised fundamental flux linkages in a lumped U-core coupling system are shown in Figure 2-15(a). The leakage fluxes on the primary and secondary sides are shown by $l_1$ and $l_2$ respectively, and in a good coupler design these are minimised. Excessive leakage results in inefficient systems and can make compliance with magnetic field exposure regulations difficult.

![Figure 2-15: Coupling in an IPT system with two U-cores.](image)
the copper and core loss are not considered. The combined effect of the inverter and compensation network typically results in the primary inductance being driven with a controllable current source therefore it is more convenient to use the mutual inductance model shown in Figure 2-16(b) [86]. This allows the primary and secondary to be largely designed independently.

A sinusoidal current in the primary inductance induces, via the mutual inductance, a voltage in the secondary inductance and when there is no load the voltage is referred to as the open circuit voltage or $V_{oc}$ and is given by [21, 24, 87]:

$$V_{oc} = j\omega M I_1$$

(2.4)

If a load resistor is connected across the secondary a current will flow and if the resistor is made to be $0\Omega$, the current is the short circuit current, $I_{sc}$ and is given by:

$$I_{sc} = \frac{j\omega M I_1}{j\omega L_2} = \frac{M}{L_2} I_1$$

(2.5)

The uncompensated VA, $P_{su}$, of the receiver is its apparent volt-ampere capacity, which is the product of $V_{oc}$ and $I_{sc}$ as given in (2.6). This is a useful measure and is used throughout this thesis along with the coupling coefficient to compare different receiver designs operating under the same conditions. Clearly having the highest coupling coefficient possible with a given air gap is preferable bearing in mind cost constraints. However, in practical systems (run with a constant $I_1$) a compromise may need to be made as the peaks in $P_{su}$ and $k$ do not necessarily occur at the same design point if a given variable is swept between its limits. A constant current is used in practice to ensure the full capacity of the wire is used for cost efficiency. Further discussion on optimisation can be found in Chapter 5 where optimal solutions are selected accounting for the variations in both $P_{su}$ and $k$. 

![Figure 2-16: (a) T circuit and (b) mutual inductance models.](image_url)
Chapter 2 – Inductive power transfer concepts

\[ P_{su} = |V_{oc}I_{sc}| = \omega \frac{M^2}{L_2} I_1^2 \]  
(2.6)

The power that can be delivered to a resistive load is reduced due to the voltage drop across the inductance of the receiver and this limits the maximum power that can be delivered. This maximum power is:

\[ P_{max} = \frac{P_{su}}{2} = \frac{1}{2} \frac{\omega M^2}{L_2} I_1^2 \]  
(2.7)

In consequence, tuning is used to improve power flow and this achieved by compensating for the inductive reactance of the receiver with a capacitor. The value of the capacitor is determined by (2.8) so that its impedance matches that of the receiver forming a resonant tank at the system operational frequency.

\[ C_2 = \frac{1}{\sqrt{\omega^2 L_2}} \]  
(2.8)

The capacitor can be placed either in series or parallel with the secondary inductance as shown in Figure 2-17(a) and (b) respectively, noting the load is replaced with an equivalent ac resistance, \( R_{Eq} \) [21, 88-90] [91]. The quality factor (\( Q \)) is a standard measure of a resonant network and it describes the ratio of the reactive to real power [92, 93].

![Figure 2-17: (a) Series and (b) parallel connected compensation capacitors.](image)

In the case of a tuned receiver, the formula for the quality factor of the secondary network, \( Q_2 \), varies depending on whether series or parallel tuning is used and is given in [91] and below:

\[ Q_2 = \begin{cases} \frac{\omega C_2 R_{Eq \parallel}}{\omega L_2} & \text{parallel} \\ \frac{\omega L_2}{R_{Eq \parallel}} & \text{series} \end{cases} \]  
(2.9)

Compensation boosts both the voltage and current in the resonant tank, however only one parameter can be applied to the load. Parallel compensation allows a voltage boost while
series tuning allows a current boost. Both tuning topologies enable increased power flow to the load by a factor of $Q_2$:

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

(2.10)

There are however limits on the maximum values of $Q_2$ that can be realised in practical systems. The bandwidth of a resonant circuit is shown by (2.11) and this range is centred on the operational frequency. The limits of the range are defined by the upper and lower frequency at which the output power drops by half. If the tuning is not accurate (caused by capacitor manufacturing tolerances) the maximum $Q_2$ that can be achieved is limited. In addition the VA of the receiver, which dictates practical receiver size due to heating, is $P_{out} Q_2$ for series tuning and $P_{out} \sqrt{Q_2^2 + 1}$ for parallel tuning. In consequence, there is an effective limit of 10 on the maximum quality factor and the work in this thesis assumes a conservative limit of 6 [24].

$$BW = \frac{\omega}{Q_2}$$

(2.11)

### 2.4.1. Considerations for lumped systems

As this thesis is primarily about lumped systems further elaboration on $k$ is needed. The mutual inductance is easily calculated from inductance measurements as shown 2.12 where $L_{1,o}$ refers to the measurement of the primary with the secondary open while $L_{2,s}$ refers to measuring the secondary with the primary shorted.

$$M = \sqrt{L_{2,o}(L_{1,o} - L_{1,s})}$$

(2.12)

The coupling coefficient can be determined from:

$$k = \frac{\sqrt{(L_{1,o} - L_{1,s})}}{\sqrt{L_{1,o}}}$$

(2.13)

The expression shown in (2.10) does not enable an intuitive appreciation of the importance of having a high coupling coefficient – coupling is indicated by the $M^2/L_2$ ratio. Using the definition of $k$ in (2.1), (2.10) can be written as:

$$P_{out} = \omega \frac{M^2}{L_1 L_2} L_1 I_1^2 Q_2 = k^2 V_1 I_1 Q_2 = VA_1 k^2 Q_2$$

(2.14)
Thus the power delivered to the load depends on the input VA of the primary pad, the square of the coupling factor and the loaded quality factor of the secondary resonant network. Power transfer can be improved by increasing the VA on the primary and allowing relatively large quality factors (<10) however doing this will increase the standing losses in the couplers.

The ideal approach is to increase the coupling coefficient by improving the magnetic circuit and this is the focus of this thesis. The power transferred is proportional to $k^2$ therefore even small improvements are extremely beneficial. As mentioned earlier, $Q_2$ is ideally limited to ~6 and the resulting boost in $P_{out}$ can be applied to any receiver design therefore $Q_2$ is not critical in comparing the performance of sets of couplers (other than to avoid saturating the ferrite as discussed in Chapter 5).

### 2.4.2. Native quality factor

An important objective in any coupler design is to ensure that the native quality factor, $Q_L$, of both the transmitter and receiver pads is high. The native quality factor of an inductor is given in (2.15) where $R_{ac}$ is the ESR of the coil and is a result of copper loss due to skin depth and proximity affects as well as core loss due to hysteresis loss and eddy currents.

$$Q_L = \frac{\omega L}{R_{ac}}$$

(2.15)

Having a high $Q_L$ is essential to ensure the IPT system is efficient since the power dissipated in a pad is given by:

$$P_{loss} = \frac{VA}{Q_L}$$

(2.16)

In the case of a transmitter pad the VA is $VA_1$ and this further illustrates why increasing this parameter to improve power flow (in (2.14)) is not recommended. Numerous coupler designs are presented in this thesis however they typically have native quality factors that are greater than 250. A 2kW IPT system is presented in Chapter 3 has a primary VA of 36kVA resulting in a loss of ~140W. Realising high $Q_L$ values is achieved by careful magnetic design as discussed in Chapter 3 and 6 – ferrite strips are used as flux guides and are placed purposefully to ensure minimise eddy current loss in shielding materials.

The losses in the receiver pad are usually much lower than the transmitter coupler given the native $Q_L$ is much higher than the largest operational circuit $Q_2$ (~6). Thus a 2kW pad has its
highest VA$_2$ of $Q_2P_{out} = 12$kVA and pad losses under these worst case conditions is 48W or 2.4% of $P_{out}$.

2.5. Magnetic field exposure guidelines

In order to achieve widespread application of IPT EV charging systems, it is essential that the level of the leakage magnetic field around power pads is within appropriate guidelines and regulations. The International Commission on Non-Ionising Radiation Protection (ICNIRP) has done extensive research on health effects due to exposure to EMFs and has produced a set of guidelines. Exposure level limits are suggested and earlier versions ICNIRP guidelines have been used as a base for regulations created by the Australian Radiation Protection and Nuclear Safety Agency (ARPANSA). The following sections briefly describe the reasons for the limits in the ICNIRP guidelines and what values they are. Power pad compliance is investigated in Chapters 3 and 6.

2.5.1. ICNIRP Guidelines

ICNIRP has established guidelines for protecting people exposed to EMFs from adverse health effects [94]. The restrictions in the guidelines are based on evidence regarding acute effects. Currently available knowledge indicates that following the restrictions will protect people from adverse health effects due to exposure to low frequency EMF. Acute effects are determined by the effect that EMFs have on the retina. This is part of the central nervous system (CNS) and serves as a general conservative model for such neuronal circuitry, this conducts nerve impulses. Risks posed by EMFs are mainly due to transient nervous system responses. These are peripheral nervous system (PNS) and CNS stimulation, the induction of retinal phosphenes and effects on some aspects of brain function. The PNS extends to bodily parts such as arms and legs and connects to the CNS. A retinal phosphene manifests itself as a visible ring or spot of light and is the result of stimulation caused by some effect other than light, the most common example is pressure on the eyeball.

The permeability of tissue is similar to that of air resulting in little magnetic field perturbation however it is somewhat conductive allowing a varying magnetic field to induce currents. When currents are induced in the retina it is stimulated and the result is similar to looking at a light source – this causes phosphenes. The levels at which these phosphenes are formed depends on the individual and ICNIRP has used volunteers to determine the EMF levels at which they occur. The inherent uncertainty with this approach means reduction factors are
applied when establishing the exposure guidelines. The guidelines may not necessarily prevent interference with medical devices such as pacemakers, metallic prostheses and implanted defibrillators.

The ICNIRP has set separate reference levels for general public and occupational exposure to time varying EMFs in the frequency range of 1Hz to 10MHz. Additional Radio Frequency (RF) specific restrictions need to be considered at frequencies greater than 100kHz. IPT systems investigated in this thesis operate at either 20kHz or 38.4kHz so the additional restrictions are not considered. The reference levels are determined from base restrictions by mathematical modelling of a 2mm x 2mm x 2mm cube of tissue and are calculated for maximum field coupling; this results in maximum protection. Base levels are determined by CNS and PNS stimulation and the formation of retinal phosphenes.

Occupational exposure levels are defined for adults in workplace environments where the field levels are generally known. ICNIRP recommends administrative controls for workplaces such as limiting access, the use of audible and visual warnings and as engineering controls in necessary i.e. detecting people in the vicinity and shutting off equipment. The general public, however, are usually unaware of exposure therefore the reference levels are far more stringent. Also, individuals of all ages and health status are exposed and some may be more susceptible to EMF. Both occupational and general public exposure reference levels are frequency dependent.

Magnetic field strength and flux density levels are relevant to IPT systems – there are no practically reachable points where a user can be exposed to strong electric fields. Given the ease of measurement with a flux probe, flux density is used to determine system compliance in this thesis. The device used is a Narda ELT-400 exposure level tester that has a frequency range of 1Hz-400kHz (-3dB) with two selectable high pass filters starting at 10 and 30Hz. These filters prevent the probe picking up variations in the earth’s magnetic field as it is moved. The probe contains three 100cm$^2$ coils that enable isotropic flux density measurements with an uncertainty of ±4% (50Hz - 120kHz) [95].

The RMS electric and magnetic field strength levels for occupational exposure are shown in Table 2-1 and for general public exposure in Table 2-2. The occupational exposure limit (3kHz - 10MHz) is 100μT and the general public limit is 27μT.
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<table>
<thead>
<tr>
<th>Frequency range</th>
<th>E-field strength (E (kV m⁻¹))</th>
<th>Magnetic field strength (H (A m⁻¹))</th>
<th>Magnetic flux density (B (T))</th>
</tr>
</thead>
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<tr>
<td>1 Hz-8 Hz</td>
<td>20</td>
<td>1.63 x 10⁵/f²</td>
<td>0.2/f²</td>
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<tr>
<td>8 Hz-25 Hz</td>
<td>20</td>
<td>2 x 10⁴/f</td>
<td>2.5 x 10⁻²/f</td>
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<tr>
<td>25 Hz-300 Hz</td>
<td>5 x 10²/f</td>
<td>8 x 10²</td>
<td>1 x 10⁻³</td>
</tr>
<tr>
<td>300 Hz-3 kHz</td>
<td>5 x 10²/f</td>
<td>2.4 x 10⁵/f</td>
<td>0.3/f</td>
</tr>
<tr>
<td>3 kHz-10 MHz</td>
<td>1.7 x 10¹</td>
<td>80</td>
<td>1 x 10⁻⁴</td>
</tr>
</tbody>
</table>

Note f is in Hz

<table>
<thead>
<tr>
<th>Frequency range</th>
<th>E-field strength (E (kV m⁻¹))</th>
<th>Magnetic field strength (H (A m⁻¹))</th>
<th>Magnetic flux density (B (T))</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 Hz-8 Hz</td>
<td>5</td>
<td>3.2 x 10⁴/f²</td>
<td>4 x 10⁻⁵/f²</td>
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<tr>
<td>8 Hz-25 Hz</td>
<td>5</td>
<td>4 x 10⁴/f</td>
<td>5 x 10⁻³/f</td>
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<td>50 Hz-400Hz</td>
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<td>6.4 x 10⁴/f</td>
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<tr>
<td>3kHz-10MHz</td>
<td>8.3 x 10²</td>
<td>21</td>
<td>2.7 x 10⁻⁵</td>
</tr>
</tbody>
</table>

Note f is in Hz

**Table 2-1: Reference levels for occupational exposure (RMS values)**

**Table 2-2: Reference levels for general public exposure (RMS values)**

The reference levels assume the electric and/or magnetic field is uniform with respect to the size of the human body and there is little variation. However, in many practical applications the source of the field is relatively close to the body resulting in non-uniform exposure. A very conservative measure would be to limit the source to ensure the exposed body part is within reference field levels however this could hinder the performance of the system. INCIRP suggests spatial averaging is used instead; systems are considered safe if the body average is within exposure limits. Local measurements may exceed the reference level however the base restrictions should not be exceeded. The values of the local limits have not been determined by ICNIRP, they suggest standardising groups determine these dosimetrically. ICNIRP also does not specify techniques or instrumentation for accurately determining electric or magnetic fields. These insufficiencies make design development and validation difficult.

### 2.5.2. ARPANSA radiation protection standard

ARPANSA has produced an exposure standard based on ICNIRP guidelines that covers frequencies from 3kHz to 300GHz [96]. It is noted, however, that the ARPANSA standard is
based on the INCIRP guidelines that were produced in 1998. These were revised in 2010 and the exposure limits were raised and the frequency ranges have been modified. For example, in the 0.8-150kHz range the 1998 general public exposure level was 6.25μT, it is now 27μT between 3kHz-10MHz.

The ARPANSA standard is also frequency dependent and it addresses issues that ICNIRP intentionally left out for standardising bodies. These include spot limits, body average and temporal averaging. Spot limits may be up to a factor of $\sqrt{20}$ greater than the exposure level at a given frequency. Temporal averaging allows exposure measurements to be taken over a period of time – in the of 10-100kHz frequency range this corresponds to 1-10 cycles respectively. Spatial averaging may be used for frequencies lower than 100MHz and this involves taking the average exposure level at four points on the human body, the head, chest, groin and knees. As long as the average value is at or below the exposure level and no spots exceed the applicable spot maximum, the system is considered to be in compliance.

The exposure levels for the ARPANSA standard are shown in Table 2-3. The body average reference level for general public exposure to magnetic fields is 6.1μT between 3-100kHz and no spots are allowed to be greater than 27.3μT. The occupational exposure level is 31.4μT between 3-65kHz with a spot limit of 140μT.

ARPANSA has not explicitly mentioned locations for the head, chest, groin and knees making spatial averaging difficult to apply. In this thesis, power pads are assumed to be floor mounted so the body average increases with decreasing height. This means the worst case adult exposure scenario occurs with a 1500mm tall person, this height roughly corresponds to a female 2 standard deviations below the mean [97]. To determine the centre of the head, 100mm is subtracted from the height and the remaining three parts are apportioned over the remaining height. This process results in the following heights: head-1400mm, chest-1050mm, groin-700mm and knees-350mm. APRANSA has not updated their standard to bring it in to alignment with the 2010 ICNIRP guidelines, therefore the maximum spot limit is still assumed to be a factor of $\sqrt{20}$ greater than the body average (27μT). A system is considered to be in compliance with the ARPANSA standard if the body average is less than or equal to 27μT and no spots are above 120.7μT.
2.6. Conclusion

There are two main types of coupling apparatus used in IPT systems; these are the distributed and lumped topologies. Distributed systems allow power transfer at any position along an elongated track however the lateral tolerance is very limited. A more complicated receiver topology, called a quadrature receiver, has been proposed and this can couple with either horizontal or vertical magnetic field directions resulting in improved lateral tolerance.

Conventional lumped couplers found in the literature are based on common core topologies and several have been presented and have been shown to offer limited performance in practical EV charging applications. Power pads used in this thesis are thinner, lighter and more robust than conventional couplers.

Having high coupling coefficients with the absolute minimum of material is one of the most critical requirements to ensuring the overall feasibility and efficiency of an entire IPT system. The power coupled is proportional to the square of the coupling coefficient so even small improvements in $k$ result in a considerable increase in power transfer.

The basis of magnetic field exposure guidelines and regulations has been presented and the aim of these is to prevent adverse health effects. The ICNIRP has not explicitly provided measurement techniques however this has been addressed by APRANSA. A body average approach is recommended and the system is considered to be in compliance if the body average (over the head, torso, groin and ankles) is lower than 27 $\mu$T and no points are greater than 120.7 $\mu$T.
Chapter 3. Design and optimisation of circular power pads

3.1. Introduction

In this chapter the design of a circular pad structure suitable for EV battery charging is investigated. The complexity of the problem arises as the magnetic field needs to be shaped with ferrite and aluminium, and this means that analytic or precomputed solutions are impractical and have limited flexibility [98]. Those solutions are generally more suited to situations where there are no magnetic or conductive materials near the coil as in lower power applications [54, 99].

Consequently, 3D Finite Element Method (FEM) simulations and experimental data are used in this thesis. The software used is called JMAG and it is produced by the Japan Research Institute, Ltd. Optimisation requires hundreds of pad models to be considered and building a significant proportion of them is expensive and time consuming. The approach used here is to confirm simulated and measured results match prior to running several iterations where a pad is modified slightly. Approximately 5% of simulated pads have been built and simulation results agree, with errors typically being less than 10%. Using this approach is both cost and time efficient and magnetic flux vector plots provide good insight into the operation of power pads [100] [101].

EV charging systems in this thesis are assumed to require power levels of 2-7kW with an operational air gap of 100-250mm. Typical small cars have ground clearances in the range of 100-150mm however a protective layer of asphalt covering a transmitter pad may be up to 75mm thick. This increases the air gaps to 175-325mm. The system also needs to transfer full power with sufficient horizontal tolerance to enable a driver to park easily – a radius of 100-200mm should be sufficient, however with automatic guidance ~50mm may be acceptable.

The optimisation process begins by analysing the layout of already existing 420mm pads. To ensure cost efficiency and minimum weight, any added ferrite must be used in the most effective manner – this is determined by volumetric comparisons where $P_{su}$ with a given air gap is divided by the volume of ferrite. The 420mm pads have limited coupling with practical EV charging air gaps due to their small size and therefore models of scaled up pads are investigated subsequently. Pads 600mm in diameter are chosen and allow 2kW power transfer after an optimisation process.
Finally, this chapter concludes with the design of a 2kW pad system and its compliance with human magnetic field exposure guidelines. These are set by the International Commission on Non-Ionising Radiation Protection (ICNIRP) however no explicit measurement techniques are provided. This has been addressed by the Australian Radiation and Nuclear Safety Authority (ARPANSA) and using this approach pad compliance can be easily determined.

### 3.2. Structure and model of circular pads

The ground based pad is referred to as the Transmitter (Tx.) and the vehicle mounted pad as the Receiver (Rx.). Both pads are made identical to simplify the optimisation process however in practice the number of turns on the receiver can be adjusted to provide the voltage required by the vehicle battery. Each power pad has six main components as shown in the exploded view of Figure 3-1. The aluminium ring and backing plate make pads rigid as well as shield the chassis of the EV and surrounding area from stray magnetic fields, this is discussed in section 3.7.

![Exploded view of a power pad](image)

Figure 3-1: Exploded view of a power pad

The dimensions of the pad and therefore the model are shown in Figure 3-2, along with the excitation conditions of the transmitter. The power supply intended for such an EV charging system operates from single-phase with near unity power-factor and has unique features making it low cost [41, 102]. This supply uses an inductor-capacitor-inductor (LCL) impedance converting network that converts the voltage sourced inverter into a current source suitable for driving the primary pad inductance. It also filters the square wave output of the bridge minimising RF interference.
The power supply achieves cost efficiency through the removal of the large DC bus capacitance and by using the leakage inductance of the output isolating transformer to form the first inductor of this LCL network. The second inductor is formed by the inductance of the transmitter pad and connecting wires. In practice this is often larger than desired so a series capacitance is added to achieve the desired value. The reduced component count in the power supply makes it extremely light and compact. Using a low bus capacitance increases safety, which is especially important in a domestic setting, however this results in a 100Hz modulation on the DC bus. This modulation combined with the modulated output of the inverter bridge results in peak currents that are twice as high as the RMS current. Consequently care must be taken in the design to ensure the ferrite in the primary pad does not saturate. As the focus of this thesis is magnetic design and optimisation, the operation of the power supply will not be discussed further. It is assumed to produce a 23A rms sinusoidal current in the transmitter at either 20kHz, 38.4kHz or 50kHz with a peak current of 46A.

Measured and simulated $V_{oc}$, $I_{sc}$ and $P_{su}$ profiles are shown in Figure 3-3(a)-(c). In these results the “horizontal offset” describes the distance between the pad centres while the “vertical offset” or “$\delta_z$” describes the separation between the plastic covers that form the face of each pad, and each is 5mm thick. The EV battery requires an input power of 2kW and this is assumed to be the available maximum power rating of a typical household mains socket.

As shown in Figure 3-3, there is a small error between the measured and simulated results. The results of the simulation are slightly conservative and the error may be due to the following reasons: A compromise between the size of the surrounding air region and the number of elements in the model has been made to achieve the highest accuracy to simulation time ratio. Symmetrical boundaries are used and the JMAG solver forces flux to be parallel to the bounding air region. To minimise the effect this has on results, the air region is made 4 to 5 times larger than the pad diameter. The manufacturing tolerances (±3mm) result in
positions for the coil and ferrite that are not as precise as in the simulation. The peak flux density is less than 150mT. Therefore non-linearity in the ferrite has been ignored and it is considered to be an isotropic material with a relative permeability of 2300. However, overall the results are in excellent agreement and enable further pad designs to be explored with confidence.

Figure 3-3: Measured and simulated results for the 420mm diameter circular pads (a) $V_{oc}$, (b) $I_{sc}$ and (c) $P_{su}$ against horizontal offset at specified vertical offsets. The operational $Q$ of a 2kW system operating with a 40mm air gap is shown based on measured results ($I_1=23A$ at 50kHz).

3.3. Power null and development of oval pads

A null occurs in all of the profiles shown in Figure 3-3(a)-(c) when the horizontal offset is 160mm regardless of vertical separation. This is consistent between measured and simulated
Chapter 3 – Design and optimisation of circular power pads

results and occurs at the point where mutual coupling between the coils reduces to zero. This null is a result of zero net flux through the receiver coil and is due to the shape of the magnetic field created by the transmitter. This is shown with 3D FEM simulation results where magnetic flux density vectors on a cut plane through the centres of horizontally misaligned pads (in the null position) are shown in Figure 3-4 where the receiver pad is open circuited.

Superimposed idealised flux paths are drawn showing the flux around coil parts (a) and (b). Flux (Φ<sub>a</sub>) from coil part (a) does not link the receiver coil but it increases the magnetic flux density in the centre of the transmitter pad. The majority of the flux produced by coil part (b) flows along the path indicated by Φ<sub>b1</sub> and passes in to and out of the centre of the coil resulting in no net linkage. Flux path Φ<sub>b2</sub> is pushed outward due to magnetic pressure from Φ<sub>b1</sub> and the point of inflexion is a result of repulsion flux created by eddy currents in the aluminium shielding ring. The flux path Φ<sub>b2</sub> cannot return through the middle of the receiver pad due to pressure from Φ<sub>b1</sub>. It is forced toward the backing plate and travels through the ferrite and returns via the middle of the transmitter pad without linking the receiver coil. If Φ<sub>b2</sub> returned though the middle of the receiver there would be flux linkage with the receiver resulting in an induced voltage. The flux paths remain exactly the same when the receiver is shorted – further illustrating there is no mutual coupling.

Although power transfer at the null is extremely small and practical operation in its vicinity is not possible, its existence is responsible for the fundamental horizontal tolerance limit of circular pads. As shown in Figure 3-3, regardless of separation and hence the power output when the pads are centred, all profiles must pass through this null when displaced horizontally and therefore it defines the gradient of the horizontal power profiles. The null occurs after an offset of approximately 38% of the pad diameter with the 420mm pads.

Figure 3-4: Simulated magnetic flux density vector plot with idealised flux paths superimposed on 420mm diameter pads in the null position.
Consequently, the pad diameter needs to be increased substantially in order to ensure acceptable horizontal tolerance for EV applications.

The pick-up operational $Q$ for a system using 420mm diameter pads with a separation of 40mm is shown in Figure 3-3(c) with a desired power output of 2kW. The system will still allow full power transfer with a horizontal offset of 90mm assuming a $Q$ of 6 is allowed. Although power transfer is sufficient, the relatively small air gap means practical implementation is infeasible. EV charging systems are designed for a given operational separation with some tolerance to changes in height. If the operational air gap is less than 40mm with 420mm pads, small changes in vertical travel result in large variations in coupled power as shown by the $P_{su}$ at the origin of the profiles of Figure 3-3(c). Excessive mutual inductance means that the coupled voltages and currents are much larger than expected and the control switches have to shunt much higher currents. To avoid this, the secondary electronics need to be significantly overrated to avoid damage, increasing system cost.

Furthermore, the transmitter pad forms the last inductor of the LCL network and it is critical that its inductance remains within the range that can be driven safely from the power supply [41]. The inductance of the transmitter pad rises significantly due the presence of ferrite in the receiver pad if the separation between the power pads becomes much lower than originally designed for. This will detune the primary network forcing the power supply inverter bridge to supply additional reactive current and this ultimately damages the switches. These reasons provide a strong incentive to design and optimise power pads to enable operation at greater separations for practical implementation.

### 3.3.1. Oval pads

The position of a null in a power profile of a circular pad is related to its diameter therefore oval shaped pads were investigated to determine if horizontal tolerance could be improved. The layout is similar to that of a circular pad with equivalently spaced ferrites however a rectangular section is added in the middle as shown in Figure 3-5(a). Like the circular pad, the coil also consists of 12 turns of 4mm diameter Litz wire. The oval pads are larger and thus have better coupling than the original 420mm pads. In order to make a fair comparison a 472mm diameter circular pad, which has the same area and volume of ferrite, was used. Oval pads are directional therefore a complete horizontal profile requires simulation in one quadrant. However, it has been observed that as profiles that have components in both the $x$ and $y$ axes fit between the profiles in either axis only two profiles are needed for complete
Chapter 3 – Design and optimisation of circular power pads

characterisation. For reference, the longest axis of an oval pad is the $x$-axis and the shorter the $y$-axis as shown in Figure 3-5(b).

The results of the comparison are shown in the $P_{su}$ profile of Figure 3-5(c), with both pads simulated under the same conditions with vertical separations of 60mm. The 472mm circular pad outperforms the oval pad up to an offset of 50mm. After this point the circular pad offers similar performance to oval pads on a 45º offset (the middle of the shaded region of Figure 3-5(c)). The oval pads perform significantly better with a large offset in the $x$-axis but at this point the coupling is too low for the pad to be usable. Oval pads are extremely sensitive to misalignment in the $y$-axis and this polarised nature is a result of the pad geometry.

![Figure 3-5: (a) Plan view of an oval pad (b) axes of an oval pad system (c) comparison of power profiles of circular and oval pads. $I_1 = 23A$ at 50kHz.](image)

Both oval and circular pads use planar coils on a ferrite substrate and these produce a reasonably uniform flux density above the pad. This is illustrated by flux density contours on $xy$ planes 40mm above circular and oval pads in Figure 3-6(a) and (b) respectively. This uniformity means mutual flux is mainly proportional to the common area between pads. The
change in common area, between initially aligned oval pads, as a receiver pad is moved in the x-axis is significantly smaller than if the offset were in the y-axis resulting in highly polarised pads.

The flux density above the 472mm pad in Figure 3-6(a) is higher than that above the oval pad of Figure 3-6(b) resulting in better power transfer as shown in Figure 3-5(c). Given the material requirements for the oval pads is similar to that of the 472mm circular pads and the tolerance is comparatively poorer, oval pads have been omitted as a design option.

![Figure 3-6: Flux density on a plane 40mm above (a) circular and (b) oval transmitter pads ($I_1 = 23A$ and each pad has 12 turn coils).](image)

### 3.4. Optimisation of circular pads

#### 3.4.1. Investigating coupling effects due to pad layout

The layout of the 420mm pads to date has been determined by the availability of ferrite bars, current rating of the Litz wire, and experience (assuming both pads are identical). Assuming the transmitter pad is driven with a constant current source, increasing the number of turns increases the uncompensated power of the receiver linearly due to the increased magnetomotive force (mmf). However, doing this increases the pad inductance resulting in higher terminal voltages. There are practical limits based on insulation, safety and standards so that terminal voltages should be below 1kV and ideally below 500V. Wide coils with many turns also increase cost and cause high $P^2R$ loss that lowers system efficiency. The chosen 12 turn coil is a good compromise between copper loss, coupled power, flux density
in the ferrite and pad inductance. The inductance of the transmitter pad is 60uH and at 23A at 50kHz this corresponds to a terminal voltage of 430V.

The amp-turns ($NI$) determine the flux density in the transmitter pad and hence coupled power to the receiver pad. The flux density must be constrained to avoid saturation of the ferrite (which usually occurs at 200-350mT depending on the type) and to ensure hysteresis loss is not excessive. The peak flux in the transmitter pad is around 100mT due to double modulation, which is a result of the operation of the low cost power supply discussed in section 3.2. Note the RMS flux density is shown in Figure 3-4. The current in the receiver pad is significantly lower than the transmitter so that even under maximum resonance ($Q = 6$) the ferrite in the receiver does not saturate. If the required power cannot be transferred within these constraints a larger pad is required.

The optimisation approach employed here is to use the original pad, as shown in Figure 3-2, as a base and by simulating possible changes, make a recommendation on the preferable variable to maximise. Four possible changes without adding extra ferrite are to: increase the pitch of the coil, change the coil centre diameter (indicated by E in Figure 3-2), or to maintain the same centre diameter while moving the ferrites radially inward or outward. The coil is simulated with a rectangular cross-section that is divided into 12 sections rather than individual turns to reduce the number of elements and hence simulation time. Increasing the pitch involves widening the cross-section. This approach has been verified with measured and simulated results on bar shaped receivers and is accurate if the gap between individual turns is less than a few wire diameters [103].

### 3.4.1.1. Coil pitch

Pad parameters against coil width normalised to the original pad are shown on the graphs of Figure 3-7(a) and (b). Simulations where the normalised coil width is less than 1 have been undertaken for completeness. Practical implementation would require multi layer coils or wire rectangular in cross section but this may undesirably add to the thickness of the pad.

Based on $P_{su}$, the coil width of the original pad is close to optimal. Ideally the coil width for pads with a separation distance of 40mm is 0.9 of the original pad (43mm); this translates to approximately 40% of the ferrite length being covered by the coil. The coupling coefficient reaches a maximum when the coil width is 2.2 times the original.
3.4.1.2. Ferrite and coil positions

The layout of the original 420mm pad is shown in Figure 3-8(a). The extremes of the three layout changes that can be done without adding extra ferrite are shown in Figure 3-8(b) to (d): (b) radial ferrite movement, (c) coil and ferrite and (d) coil diameter range. The results are presented in Figure 3-9 and these show that considering all parameter variation permutations is not necessary to determine optimal pad layout. Performance is determined by the relative positions of the coil and ferrites. It may appear that more power can be transferred if the coil radius is increased by 4mm and the ferrites moved in by 4mm since the maxima of both curves are symmetric about the y-axis. However moving the coil out by 4mm means the ferrite is effectively moved in by 4mm relative to the coil.

The coil central diameter (shown in Figure 3-2 by \( C_d \)) should be approximately 57% of the pad diameter (240mm) as this provides a good compromise between \( k \) and \( P_{su} \). The highest \( k \) occurs when the ferrite is centred on a coil with a \( C_d \) of 208mm however there is almost a 100VA drop in \( P_{su} \). High coupling coefficients are desirable in IPT systems however in this case a higher \( P_{su} \) is preferred (as the transmitter is powered by a constant current source).

The position of the null in the horizontal \( P_{su} \) profiles (shown in Figure 3-3) is determined by the coil diameter. Using a larger coil to shift the null is not recommended as \( P_{su} \) drops significantly as shown in Figure 3-9(c) resulting in worse overall performance.
Figure 3-8: (a)-(d) Various pad designs where: (a) original, (b) varying ferrite, (c) varying coil and ferrite, and (d) varying coil position.

Figure 3-9: (a) $V_{oc}$, (b) $I_{sc}$, (c) $P_{uw}$ and (d) $k$ against ferrite, coil and ferrite and coil position variation from that of the base pad ($\delta_z = 40$mm).
3.4.2. Improving coupling by adding ferrite

Coupling will improve if more ferrite is added and this may be done by adding more bars, making the bars longer, wider or thicker or by adding feet that extend the portion of ferrite uncovered by the coil. To determine which variable should be maximised, various simulations were undertaken while keeping the pads at a fixed separation and vertically aligned for ideal power transfer. An efficiency comparison based on ferrite utilisation was then made where $V_{oc}$, $I_{sc}$, $P_{su}$ and $k$ are plotted against changes in the volume of ferrite (here this change is represented in variations in length, width or shape). Adding ferrite that is not utilised efficiently will make the pad unnecessarily heavy and expensive.

The volumetric comparison focussed on changes to the original design of Figure 3-2 and each variable was varied from its minimum to maximum possible value given geometrical constraints. The notation in the legend shows the limits of the variable and the increment used is preceded by a comma. The results are shown in Figure 3-10(a)-(d) below and the four curves intersect at a point equal to the volume of the initial pad. The curve for the pad with feet has a small offset at this point because additional ferrite is required to make the feet.

The gradient of the curves relating to increasing ferrite length is highest, indicating that this gives the best ferrite utilisation. This is expected since longer blocks permit higher flux paths in air above each bar as shown by the idealised flux lines in Figure 3-4 (flux paths are further studied in Chapter 4). If the maximum $P_{su}$ is considered alone for these pad design options, then it appears that the addition of feet is the best option. However, this only occurs because the feet effectively shorten the air gap between the pads resulting in higher power transfer. This becomes apparent when $k$ in Figure 3-10(d) is considered. All $k$ curves except for the variation in ferrite length are clustered and have relatively low gradients indicating that the flux generated by the transmitter is not being coupled effectively to the receiver. The curve for pads with added feet is below that of increased ferrite length indicating poor ferrite utilisation.

The performance of the pad is influenced least by the thickness of the ferrite bars. If pad weight is a critical design parameter, thinner blocks may be used, however doing so makes saturation more likely and increases iron loss. The peak flux density in a 6mm thick bar was found under the conditions simulated here, to be 176mT, whereas it is 102mT in a 10mm thick bar. It is possible to operate with thinner blocks.
Figure 3-10: Volumetric comparison of changes in pad parameters at a separation of 80mm. (a) $V_{oc}$, (b) $I_{sc}$, (c) $P_{su}$ and (d) $k$ against volume of ferrite as the length, number, thickness and width are changed. $I_1 = 23A$ at 50kHz

3.4.3. Relating coupled power to pad size

The 420mm pads do not achieve sufficient power transfer for practical EV charging even with bars that are made as long as possible. Larger pads are required for greater coupling and improved horizontal tolerance. The position of a null in a power profile for a given pair of identical pads occurs when the horizontal offset is ~40% of the pad diameter and this null can only be shifted further from the origin by using larger pads. This also has the desirable effect of smoothing the power and coupling profiles for a given offset.

Circular pads by their nature are scalable therefore the conclusions reached from the investigation in sections 3.4.1 and 3.4.2 have been applied to a variable size pad model (although the assumption that both the primary and secondary pads remain identical remains).
Coils that cover ~40% of the ferrite strips provide a good balance between coupling and pad inductance as shown in section 3.4.1.1. Optimal coverage can be achieved by winding coils with a larger pitch or by increasing the number of turns. In practical designs the number of turns should increase with pad size to ensure suitable power transfer and effective material usage. Using large pads to transfer relatively low power levels means the VA/cm³ of ferrite is low. The features of the variable size pad model are as follows:

- The number of turns (N) of 4mm diameter wire is adjusted according to pad size.
- The eight 30mm wide ferrite bars were made as long as practically possible and were centred on the coil.
- The coil central diameter (Cd in Figure 3-2) has been set to 57% of the pad diameter.

These parameters are shown for pads ranging from 300mm in diameter to 800mm (in 100mm increments) in Table 3-1. The layout of a 600mm pad is shown in Figure 3-11. Results are presented in Figure 3-12 with a curve for each of the six pads on each subfigure. The first two plots, (a) and (b), are referred to as $V_{oc}$ and $I_{sc}$ vertical profiles respectively. These voltage and current profiles are formed as the separation is increased between horizontally aligned pads. The simulation frequency has been changed to 38.4kHz as this optimisation work is being done in parallel with power supply development and presently available switches are more suited to a lower frequency, $I_1$ remains at 23A. Performance at 50kHz can be easily determined by linear scaling. The graphs of Figure 3-12(c) and (d) show $k$ and $P_{su}$ profiles respectively.

<table>
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<tr>
<th>Pad Dia. (mm)</th>
<th>Fe. L (mm)</th>
<th>Coil W (mm)</th>
<th>N (turns)</th>
<th>Coil Centre (Cd) (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>300</td>
<td>83</td>
<td>33</td>
<td>8</td>
<td>172.4</td>
</tr>
<tr>
<td>400</td>
<td>120</td>
<td>51</td>
<td>13</td>
<td>230</td>
</tr>
<tr>
<td>500</td>
<td>176</td>
<td>70</td>
<td>18</td>
<td>288</td>
</tr>
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<td>402</td>
</tr>
<tr>
<td>800</td>
<td>314</td>
<td>126</td>
<td>31</td>
<td>460</td>
</tr>
</tbody>
</table>

TABLE 3-1 PAD PARAMETERS (IN MM)

The power profile of Figure 3-12(d) shows coupled power increases substantially with small changes in vertical offset when the pads are relatively close together. A 500mm pad has a $P_{su}$ of 1kVA at 100mm and it is 3kVA for a 600mm pad, the power coupled triples while the pad diameter only increases by 20%. This large variation makes matching a particular pad for an application challenging as the highest power pad is not necessarily suitable when tolerance to misalignment is considered. As discussed in section 3.3, using a 420mm pad to couple 2kW
resulted in designs that were extremely sensitive to changes in both the horizontal and vertical position relative to normal operating position of 40mm (similar behaviour is shown for a 400mm pad in Figure 3-12(d)).

Extreme sensitivity to changes in vertical offset is common to all pads and is clearly illustrated by the initial steepness of all the curves. This steepness results in a high $P_{su}$ for pads with relatively close vertical spacing and this combined with the fixed position of the null results in designs that are necessarily intolerant of horizontal offset. For these reasons designs that operate with coupling factors of 0.1 to 0.4 (in the shaded region of Figure 3-12(c)) are preferable and considered more practical in an EV context where insensitivity to
positioning error is of major importance. The corresponding shaded region illustrating practical $P_{su}$ ranges is on the profile of Figure 3-12(d).

The relationship between pad size and coupled power becomes more apparent as the pads become larger. The useful vertical operation zone appears to increase steadily and is roughly shown to be linear by the gradient of the shaded area. Clearly larger pads are less sensitive to absolute vertical separation and this is indicated by the divergence of the shaded area with larger pads.

The power transfer limit of circular pads with the layout shown in Figure 3-11 is 2kVA at 220mm and an 800mm diameter pad is needed to achieve this. Larger pads of this type can be used however space under an EV typically limits pads to less than 1m² in area. Large pads are also not cost effective and are heavy so they can affect the overall vehicle efficiency. As noted earlier, the operational $Q$ of the receiver pad controller is typically limited to 6, therefore 800mm pads with a coupling coefficient of 0.15 can be expected to couple 2-5kW. A null in the power profile for this 800mm pad will occur when the horizontal offset is 304mm however, as shown with 420mm circular pads in section 3.3, the practical operational horizontal tolerance is approximately half of the distance to the null (150mm). This fundamental power transfer limit means that large circular pads are required to reduce sensitivity to positioning error.

A pad design guide where the diameter of both pads is determined by the required operational separation and a desired $k$ is shown in Figure 3-13. The increase in pad size is directly proportional to the air gap between pads as long as $k$ is fixed. This is because both self and mutual inductance are proportional to area and although these increase with the square of the pad diameter, the flux density above the transmitter pad reduces according to $1/\delta_z^2$ where $\delta_z$ is the air gap.

The key points from the investigations in the previous sections are:

- The average coil diameter ($C_d$ in Figure 3-2) should be 57% of the pad diameter.
- The ferrites should be centred on the coil.
- The ferrite bars should be made as long as possible.
3.4.4. Optimisation of a 600mm pad with added ferrite

In this section the performance is further improved by adding ferrite however a cost measure is introduced to ensure this increased volume of ferrite is compared against the improvement in performance. For this optimisation process a 600mm diameter pad was chosen as there is sufficient space between the radial strips, shown in Figure 3-11, to allow the addition of ferrite. The pad is able to couple 1kVA with an air gap of 150mm and a significant performance increase can be expected as more ferrite is added.

3.4.4.1. Width of ferrite strips

The increase in coupling for relatively small changes in the width of the bars appears to be reasonably linear as shown in the graph of Figure 3-10(d). Simulations were undertaken using ferrite sectors to investigate this further. The layout of the 600mm diameter pad model with 18 ferrite wedges is shown in Figure 3-14(a) where 18 turns of 4mm diameter wire are used. Physical implementation of a sector type pad is challenging due to the small points at which the ferrites must meet in the centre of the circle. Cutting to such fine tolerances is impractical because ferrite is brittle and pressure moulding dies that are used on green ferrite cannot be made to such small dimensions [46, 56]. Consequently, a modified design is investigated where the sector tips are removed enabling a practical ferrite structure with a hole in the centre of the pad as shown in Figure 3-14(b).
Chapter 3 – Design and optimisation of circular power pads

Figure 3-14: (a) Layout of wedge pad model with a hole diameter of ~0mm (b) hole diameter increased to 160mm (c) wedge pad with 18 of 2° sectors and (d) sector angle increased to 20°.

The diameter of the hole was varied to determine the effect of using shorter ferrite wedges. The graph of Figure 3-15(a) shows $P_{su}$ and ferrite utilisation (measured in VA coupled per cm$^3$ of ferrite) against the hole diameter. The power coupled is relatively insensitive to hole diameters less than 80mm. A hole diameter 12% of the pad diameter (72mm) is a good compromise since there is extremely little gain in power transfer if the hole is made significantly smaller, in addition the ferrite utilisation is also at its maximum.

The graph of Figure 3-15(b) shows $k$ and $k$/mm$^3$ of ferrite. The coupling coefficient reduces in a similar manner to $P_{su}$ as the hole is made larger however the coupling for a given volume of ferrite reaches a maximum with a diameter of 100mm. The coupling coefficient is related to the length of the ferrite strip but the reduction in the volume of ferrite is proportional to the square of the hole diameter. At the peak $k$/mm$^3$, $P_{su}$ is reduced to 1.35kVA, in consequence, a diameter of 72mm is chosen to ensure a higher coupled VA with a track current of 23A.

Figure 3-15: Varying hole diameters in a wedge pads where (a) $P_{su}$ and ferrite utilisation, (b) $k$ and $k$/cm$^3$ are plotted against hole diameter. $I_{t} = 23A$ at 38.4kHz

The effect of ferrite width on coupling was determined by adjusting the angle of ferrite sectors in a 600mm pad with a 72mm diameter hole. The range of variation is shown in Figure 3-14 where (c) shows 2° wedges and (d) 20° wedges (an annulus is formed because there are 18 wedges). The results are shown in Figure 3-16(a) and (b). As with the original
pad, the increase in coupling is relatively linear for angles ranging from 2° to 6°. The rate of increase in coupling reduces significantly beyond 6° indicating ferrite is not being used efficiently.

Figure 3-16: (a) $P_{su}$ and VA/cm³ (b) $k$ and k/mm³ against volume of ferrite at 150 and 200mm separations. $I_1 = 23A$ at 38.4kHz

An advantage of having more ferrite is that saturation is less likely, although with the 2kW power rating of these pads this is not a concern. There appears to be little advantage in using pads with wedge shaped ferrite bars and considering the difficulty with forming sectors in practice, further development of this option was ceased.
3.4.4.2. Adding ferrite pieces

In many applications shorter charge times are desired, this can be achieved if custom wiring is installed in a household system permitting a higher current rating. It was decided that a pad with a diameter of 600mm should provide sufficient coupling to allow 5kW of power transfer with adequate tolerance to horizontal misalignment at 150mm separation. A model of the larger pad based on the previously found optimal layout was investigated via simulation assuming the ferrite bars could be cut to the desired size. The coil width was made up of 18 turns of 4mm diameter Litz wire, the bars were made as long as possible and the coil was made 57% of the pad diameter. Since the layout is similar to the 420mm circular pad, the overall thickness of the pad is still 25mm.

Initial simulations however showed that the pad was unable to meet the desired performance requirements mentioned above with a horizontal tolerance of 100mm (operating with a transmitter pad current of 23A at 38.4 kHz). The ferrite is already being utilised relatively efficiently (VA/cm³) but more needs to be added to improve coupling. There are numerous possibilities, therefore designs that require simple ferrite cutting operations are preferred. The most optimal of the different pad designs are shown Figure 3-17(a)-(e) and ranked in order of power. In all cases it is assumed that the ferrite I120 bars have the following dimensions: 120mmLx30mmWx10mmT. The performance and material use of each pad is described by the following data:

- $P_{su}$ at separations of 100, 150 and 200mm.
- The number of I120 bars required to make the pad.
- The ferrite utilisation efficiency (calculated at a $\delta_z$ of 100mm).

Horizontal profiles of all the various pad topologies have been compared and it has been noted that regardless of the maximum $P_{su}$ when the receiver pad is centred on the transmitter, all pads experience a null when offset horizontally by approximately 40% of the pad diameter. The slope of power profile is determined by the position of the null and the $P_{su}$ when the pads are centred. At a spacing of 100mm, the $P_{su}$ for pads of Figure 3-17(e) is 2.5kVA and it is 3.8kVA for pad type (a), both pads have a null in their profile at approximately 240mm thus the slope for type (a) is greater. As shown in section 3.4.1.2, it is not advisable to increase the diameter of the coil with the aim of shifting the null as $P_{su}$ will drop substantially further increasing sensitivity to misalignment.
Designs with the best ferrite utilisation have long or narrow ferrite blocks shown in Figure 3-17(d) and (e). The former provides favourable magnetic paths that increase flux density above the pad while the latter enables flux around the coil to be guided more effectively (concept described in Chapter 5.3.2).

![Figure 3-17: (a) – (e) Comparison of different ferrite arrangements at separations ranging from 100-200mm. \( I_1 = 23A \) at 38.4kHz](image)

The desirable features of both pads have been combined to create a model of an optimal pad as shown in Figure 3-18(a). This pad weighs 15.6kg and is capable of a transferring 5kW across an air gap of 150mm with a horizontal tolerance of 90mm allowing a full power charging zone 180mm in diameter (using a maximum operational \( Q \) of 6).

The coupling coefficients for this pad against horizontal offset at 100-200mm separations are shown in Figure 3-18(b), the corresponding \( P_{su} \) profile is shown in Figure 3-18(c). Also included on the graph of Figure 3-18(c) is the required operational \( Q \) for 5kW power output at separations of 100 and 150mm. The optimised pad requires 31.5 ‘I’ cores and has a ferrite utilisation value of 3.80VA/cm\(^3\), which is within a few percent of the most optimal utilisation shown by the design of Figure 3-17(e), which is 3.89VA/cm\(^3\) of ferrite.

Assuming the EV is constrained in the forward direction as would typically be case when parking in a garage, a 180mm wide charging zone is possible with an air gap of 150mm. This is considered within the ability of a driver if appropriate markings are made in the parking space. A ground clearance of 200mm can be achieved if the transmitter is elevated by 50mm yielding a completely practical solution. The installation cost within a garage of such a system is minimal for a user since the floor needs little modification. An appropriately marked transmitter pad is unlikely to become a tripping hazard if a gently tapered circular ramp is placed around it.
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The above optimisation process was done assuming that the desired ferrite sizes are available or larger bars can be cut to size. In cases where ideal ferrite is not available, preferable compromises are to make the ferrite blocks as long as possible or as narrow as possible. The layout shown in Figure 3-17(c) uses readily available ferrite bars and although the ferrite utilisation is low, good performance is achieved since a large number of bars are used. Conversely the pad in Figure 3-17(e) also uses unmodified bars and has the highest ferrite utilisation, although it has only 18 bars resulting in a lower $P_{su}$.

![Figure 3-18: (a) Ferrite structure of optimised pad, (b) $k$, and (c) $P_{su}$ and $Q$ against horizontal offset at 100, 150 and 200mm separations. $I_1 = 23\,\text{A}$ at 38.4kHz](image)

3.5. Effect of aluminium ring on performance

In the work of this chapter, an aluminium ring has been assumed to be important but may not always be necessary and its inclusion adds to the pad diameter (shown by $\delta_{Al}$ in Figure 3-19). The ring was added to reduce magnetic field leakage and sensitivity to surrounding metallic materials as well as provide structural rigidity in the designs presented in section 3.2. The extended radius, $\delta_{Al}$, was made to be a few mm to allow for variations in ferrite manufacturing tolerances and to ensure the pad was as small as possible. The effect the ring has on coupling is investigated with a model comprising identical 700mm diameter pads with the layout and dimensions shown in Figure 3-19. This investigation begins with removing the ring and varying the coil diameter ($C_d$). Once an optimal diameter is found the ring is replaced and then moved outward (by increasing $\delta_{Al}$). A recommendation on preferable
designs is made based on a trade-off between pad size and coupling. Magnetic field leakage is investigated with regard to ring position in section 3.7.1.

3.5.1. Varying coil diameter ($C_d$) with and without a ring

As shown in section 3.4.1.2 a coil diameter ($C_d$) that is 57% of the pad diameter provides a good compromise between the power transferred and the coupling coefficient for a 420mm diameter circular pad with an aluminium ring. To determine the effect the ring has on performance, a simulation was done where the coil diameter was varied from 256mm to 576mm on pads with aluminium rings (R) and without (NR) at a vertical separation of 125mm. The uncompensated power and coupling coefficient are shown in Figure 3-20(a) and the mutual and self inductances in Figure 3-20(b).

![Figure 3-19: Model of 700mm diameter pad used for aluminium ring investigation by varying $\delta_{Al}$ with or without a ring.](image)

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>FeL</td>
<td>Ferrite width</td>
<td>28</td>
</tr>
<tr>
<td>FeL</td>
<td>Ferrite length</td>
<td>279</td>
</tr>
<tr>
<td>$\delta_{Al}$</td>
<td>Extended Rad.</td>
<td>--</td>
</tr>
<tr>
<td>$C_d$</td>
<td>Coil Dia.</td>
<td>398</td>
</tr>
<tr>
<td>$P_d$</td>
<td>Pad Diameter</td>
<td>700</td>
</tr>
<tr>
<td>$C_w$</td>
<td>Coil width</td>
<td>100</td>
</tr>
<tr>
<td>$A_l$</td>
<td>Ring Thickness</td>
<td>10</td>
</tr>
<tr>
<td>25 turn coil (Ø4mm Litz, 810x0.1mm). 10mm thick ring</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$I_1$</td>
<td>= 23A at 20kHz</td>
<td></td>
</tr>
</tbody>
</table>
Figure 3-20: Parameters against coil diameter normalized to a 700mm pad (a) $P_{su}$ and $k$, and (b) $M$ and $L_2$

The performance of the pad with a ring is in agreement with those presented in 3.4.1.2 and there is a significant increase in power if the ring is removed. For a pad without a ring, a coil diameter of 416mm (~60% of the pad diameter) is a good compromise between the power transferred and the coupling coefficient. Figure 3-20(b) shows the self inductance drops sharply as $C_d$ increases whereas it drops relatively slowly if the ring is removed. Inductance is proportional to the flux around the coil ($L = N\Phi/I$) and therefore reduces when induced eddy currents in the ring cancel a portion of the flux created by the coil. For a coil positioned relatively far away from the ring ($C_d < 350$) the effect is small resulting in the curves for $L_2$ R and $L_2$ NR being close together as shown in Figure 3-20(b). However, as the coil gets closer...
to the ring larger eddy currents are induced that cause more flux cancellation resulting in a rapid drop in inductance \((L_2 R)\).

The inductance of a pad with a ring is greatest when the coil is roughly centred on the ferrite strips \(415 < C_d < 500\) because flux is able to enter and exit from the ends as shown in Figure 3-21(a). However, for a pad without a ring, the maximum inductance occurs when \(C_d = 511\)mm but it appears that the flux on the outside edge of the coil only has a small area through which it can enter or exit the ferrite strip – the ferrite distribution is not even about the coil.

![Magnetic flux density vectors in a cross-section of circular pads](image)

Figure 3-21: Magnetic flux density vectors in a cross-section of circular pads. (a) With rings and (b) without rings \((\delta_{Al} = 40\text{mm}, \delta_z = 125\text{mm and } I_1 = 23\text{A})\)

This is explained in terms of inductance and this determines how much energy a pad can store in its magnetic field. Note there is no flux out the back of the pads due their construction – a coil sits on ferrite that sits on a layer of aluminium shielding. Flux lines tend to arrange themselves to minimise energy storage by traversing paths of least reluctance. If \(C_d\) is less than its ideal value the flux lines repel each other more on the inside of the coil.
because there is less volume (the number of flux lines linking the coil is relatively constant) reducing the inductance. When the coil diameter is optimised, the overall field repulsion is minimised due to the increased volume on the inside of the coil thus the stored energy and hence inductance for this particular arrangement is maximised.

3.5.2. Coupling dependency on ring position

The results shown in Figure 3-20(b) indicate that pad inductance drops when a ring is added due to the flux cancelling effect. Reducing flux leakage at the expense of power transfer is often necessary to ensure high power systems are able to meet magnetic field leakage standards as discussed in section 3.7.1. In order to separately determine the effect of the ring and backing plate on power transfer and coupling, simulations were undertaken where $\delta_{\text{Al}}$ was varied from 0 to 160mm with and without a ring. The uncompensated power and coupling coefficient are plotted against the extended radius are shown in Figure 3-22, the label “R-P” implies both the backing plate and ring are present while the label “NR-P” indicates only the extended backing plate is present.

![Figure 3-22: Profiles of $P_{\text{su}}$ and $k$ as $\delta_{\text{Al}}$ is increased for pads with and without rings at a 125mm air gap ($I_t = 23A$ at 20kHz)](image)

Placing the ring close to the ends of the ferrite significantly reduces performance given $P_{\text{su}}$ increases by $\sim 27\%$ when it is removed. Pads with and without rings reach a $P_{\text{su}}$ of 3.9kVA however by this point the diameter is increased by 170mm. The $P_{\text{su}}$ profiles diverge slightly when $\delta_{\text{Al}}$ is greater than 100mm showing the ring has a slight “flux catching” effect although this is not practically useful for power transfer.
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The magnetic flux density vectors of a pad cross-section in Figure 3-21(a) show the ring creates a higher reluctance path by causing the magnetic field to bend and this reduces leakage flux as described in section 3.7.1. Magnetic flux density vectors in a cross-section of a pad with a 40mm plate are shown in Figure 3-21(b). Flux is easily able to enter the ferrite strips through their ends resulting in increased power as shown in Figure 3-22. A $\delta_{AI}$ of 40mm is suggested for an optimised pad because there is little increase in performance if larger extended radii are used and smaller pads are preferable.

3.6. Implementation of a 2kW system

The design and testing of a 2kW charging system operating off a standard single phase power socket with an air gap of approximately 200mm is described in this section. The selection of the pad diameter was based on two conflicting criteria: the lack of ferrite cutting and annealing equipment, and making the pad as small as possible. As shown with a 600mm diameter pad in section 3.4.4, smaller strips of ferrite can be added in between the main strips to increase power however cutting ferrite is problematic [56].

Allowing an increase in diameter of 100mm boosts power transfer significantly as shown on the pad design curves presented in section 3.4.3. An eight strip 700mm pad will have a coupling factor ranging from 0.1 to 0.2 with an air gap of 220 to 160mm. Increasing the number of strips improves coupling while ensuring efficient use of ferrite (shown in section 3.4.2) as such a twelve strip pad was implemented with each of the strips comprising three standard ferrite N87 type I93 cores made by EPCOS. A 26 turn coil using 4mm diameter Litz wire was chosen as a compromise between pad inductance and ferrite strip coverage.

The built pad is shown in Figure 3-23(a) along with the measured and simulated results in Figure 3-23(b). As before, a power null occurs when the secondary is offset from the primary at approximately 40% of the pad diameter further illustrating the fundamental tolerance limit of circular pads. The operational frequency was reduced to 20kHz as the switches used in the inverter were best suited to a slightly lower frequency. Assuming an operational $Q$ of 6 is allowed, this pad allows a charging zone with a full power misalignment diameter of 260mm and an air gap of 210mm. This is better than the initial specification of a minimum charge radius of 200mm.
Although there are large sections of aluminium that do not appear to be shielded from the coil by ferrite, the loss in the aluminium backing plate is low. This is reflected by the high $Q_L$ value of the pad that was measured to be 250 with a calibrated Agilent E4980A precision LCR meter. This low loss is partly attributed to the low resistivity of aluminium (minimising $I^2R$ loss), the physical distance between it and the coil and because it only has to shield leakage flux. The ferrite strips guide a significant proportion of the flux generated by the transmitter coil away from the backing plate and upward to the receiver. The pads were operated at 2kW for several hours without thermal issues.

3.6.1. Bifilar winding

A large pad inductance of approximately 540μH means that 1.6kV is required across the terminals of the transmitter to get a current of 23A. This presents a potential safety hazard and requires careful terminal insulation. Compensation capacitors can be placed in series with the winding internal to the pad structure to effectively reduce the inductance and hence the voltage at the external terminals. However, this approach inherently reduces pad reliability due to additional internal connections and ideally should be avoided.

Alternatively, the pad can be bifilar wound, to lower the inductance and hence the driving voltage while keeping the $NI$ product and therefore the generated flux constant. The second approach was chosen noting a bifilar wound pad needs to be driven with 46A. The inductance of each 12 turn coil was 130μH and 131μH for the inner and outer winding respectively resulting in a total inductance of 130μH when connected in parallel. The terminal voltage is
780V and is below the general upper safety limit of 1kV. By constraining the voltage with a bifilar winding, suitable components can be selected with VA ratings that ensure operation without fault.

3.6.2. Practical design considerations – changes in pad inductances with movement

Because the pads operate with a relatively large air gap, the transmitter and receiver pad inductance variations are minimal therefore there is little effect on tuning in both the power supply and secondary power flow controller during operation as a result of misalignment. If smaller air gaps are chosen, the coupling is improved, but the coupled power will be more sensitive to any misalignment. Consequently such variations will need to be handled in the design of the system as discussed in [41, 104]. The graphs in Figure 3-24(a) and (b) show the measured variation of the transmitter self and mutual inductances against vertical separation and horizontal offset respectively.

![Figure 3-24: L₁ and M against (a) vertical offset and (b) horizontal offset (δz = 200mm, 26 turns)](image)

3.7. Circular pad compliance with the ARPANSA standard

Simulations and measurements have been undertaken to investigate leakage fields around pads in a system transferring 2kW across an air gap of 200mm. The highest leakage flux density occurs on a horizontal plane that passes through the centre of the air gap. Leakage directly above and below the pads is lower due to shielding provided by the aluminium case that is further investigated in section 3.7.1. The leakage flux pattern about vertically aligned pads is symmetric therefore leakage measurements only need to be taken on one line. For
Chapter 3 – Design and optimisation of circular power pads

cases with misaligned pads, the measurement line is opposite to the direction of movement of the receiver as illustrated in Figure 3-25. Leakage increases as the receiver is moved away from the transmitter – the ferrite within the receiver constrains flux. An offset of 130mm is used as this is the maximum horizontal tolerance possible while maintaining 2kW power transfer.

![Figure 3-25: Measurement line for worst case field leakage (δ_z = 2 00mm)](image)

Measured and simulated results are plotted in Figure 3-26(a) along with the ARPANSA spot limit of 27.3μT. This is reached at a distance of 500mm from the centre of the pad or 150mm from the pad edge. Leakage is marginally worse with misaligned pads as the 27.3μT limit is reached 550mm from centre. Assuming ARPANSA updates the spot limit based on the 2010 ICNIRP guidelines [105], a revised spot limit of 120.7μT would be reached at a distance of 400mm from the centre of the pad or 70mm from its edge. Given the pads are placed under a vehicle, there is little chance of any body part being exposed to high flux densities.

![Figure 3-26: (a) Measured and simulated flux density and (b) measured flux density at spatial averaging points on a person standing 170mm from the edge of a power pad (I_1 = 23A)](image)
Spatial averaging using measured flux density on a person standing 170mm from the pad edge shows the system easily complies with the current ARPANSA regulations. The results are shown in Figure 3-26(b) above.

Leakage flux simulations hitherto have used an aluminium shielding ring that was positioned close to the ends of the ferrite strips to minimise the size of the pad. As shown in section 3.5, performance improves significantly if the ring is moved further out or is removed. In the next section, leakage flux is investigated with ring position and a pad design that offers a good compromise between leakage and coupling is recommended.

3.7.1. Aluminium ring position and leakage magnetic flux

The approach used here is similar to section 3.5 where the rings on 700mm pads, shown in Figure 3-19, are moved away from the ends of the ferrite strips by increasing $\delta_{Al}$. Leakage flux along the measuring line of Figure 3-25 is plotted for designs with extended radii ranging from 0 to 160mm in Figure 3-27. The first curve is labelled “No ring” and shows the output without a ring and having $\delta_{Al}=0$. A larger backing plate does attenuate leakage slightly but a ring is required for the greatest reduction. The curve labelled “40mm Plate” is for a pad with the ring removed but with the backing plate still in place, when the ring is added the flux density reduces from 12$\mu$T to 4.6$\mu$T. Removing the ring increases leakage substantially as does placing it very close to the ferrite ends.

![Figure 3-27: Flux density along a contour with the origin in the middle of the air gap ($\delta_z = 125$mm and $I_1 = 23$A)](image)

The flux density up to a distance of ~550mm from the pad centre is not strongly affected by $\delta_{Al}$ however at greater distances the leakage reduces as $\delta_{Al}$ increases. The upper limit of $\delta_{Al}$
is determined by the maximum space available on an EV chassis and the extra cost for the additional aluminium. A typical small car is 1.6m wide and the flux density will be below 10μT by the door sill for all pad designs making compliance with the ARPANSA standard highly likely. The flux density behind the pads is relatively low due to shielding from the aluminium backing plate. Cross-sections of energised pads with and without rings (δ_Al=40mm) are shown in Figure 3-28(a) and (b) respectively. Adding a ring relatively close to the ends of the ferrite results in lower power transfer therefore a plate without a ring provides a compromise between leakage and coupling.

The above results show the ring should not be removed or else the pads will be adversely affected by metallic objects close to their edges. The chassis of an EV is typically made out of steel and excessive leakage flux bulging about the air gap as shown in Figure 3-28(a) will reduce the $Q_L$ of a receiver pad. The extended backing plates with rings contain the flux as shown in Figure 3-28(b) allowing installation on a steel chassis.

Larger leakage results in more energy lost in the surrounding EV chassis – this loss is proportional to $B^2$, thus slight reductions in flux density can be very effective. The ring reduces leakage by decreasing the area through which flux can escape as shown in Figure 3-21(a). Eddy currents in the ring cause the flux that does escape to curve inwards to the opposite pad, this flux path is less likely to run parallel to the surrounding chassis further.
reducing loss. Conversely, when the ring is removed with the backing plate still present as shown in Figure 3-21(b) the shaping of the flux path is reduced. Flux is allowed to travel unimpeded toward the chassis resulting in larger leakage and therefore loss.

A receiver pad might need to be recessed into the chassis to maintain the ground clearance of an EV. If this is done without a ring the flux paths out of the ends of the ferrite strips will be perpendicular to the chassis resulting is large loss, to prevent this, the recess would need to be significantly larger than the pad. This is impractical as space under a vehicle is generally limited. An aluminium ring (with a $\delta_{\text{Al}}$ of 40mm) allows the pad to be installed with an almost similar sized recess. An 80mm increase in pad diameter significantly reduces field leakage at the edges of the pads and the $P_{\text{su}}$ is within 5% of that available if the ring is removed.

3.8. Summary and Conclusion

IPT is an extremely safe and convenient method for recharging EVs, but in order to ensure the system is as efficient, cost effective and as light as possible, it is critical that the desired coupling between the power pads is achieved with a minimum amount of ferrite. Existing circular pads 420mm in diameter have been modelled with 3D FEM software and the difference between simulated and measured results is 10% at most. This ensures pads optimised using the simulator will perform as expected in practice.

Initial results showed the fundamental horizontal tolerance limit for circular charging pads is approximately 40% of the pad diameter. Consequently circular pads need to be scaled up significantly to provide sufficient power transfer and tolerance to misalignment. Circular pads ranging from 300-800mm have been simulated and profiled. A 600mm pad with added ferrite was deemed sufficient to couple the required 5kW with a 150mm air gap. Different ferrite configurations were investigated and compared using a VA/cm$^3$ of ferrite metric. This showed that long and narrow evenly spaced ferrite bars give the most effective performance to weight result. Simulation results have been used to design an optimised pad that is able transfer 5kW across an air gap of 150mm with a horizontal tolerance of 180mm. The optimal pad design can be scaled to match different power requirements or separations.

A 2kW IPT system was also built and tested using 700mm diameter power pads. Existing ferrite bars were used for convenience and offer a better practical solution since a radial strip made up of individual ferrite pieces is less likely to fracture than a single long solid bar.
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Leakage fields have been investigated via simulation and measurement and show that the pads meet ICNIRP guidelines according to expected ARPANSA regulations. The quantitative results presented in this chapter form a basis for the proper design of circular power pads.
Chapter 4. Flux paths and large air gap power pads

4.1. Introduction

In this chapter the shape of the flux paths above a circular power pad is investigated and used to propose a number of new topologies with improved (more ideal) flux paths. The height of the flux path above a pad determines its coupling ability with large air gaps [106-108]. In the context of this thesis operational air gaps of 125mm to 300mm are considered practical and in this chapter 200mm is desired. Smaller air gaps are likely with retro-fit systems where a receiver is added to an EV chassis without any recess and the transmitter is placed on the ground whereas air gaps of 300mm are likely if the transmitter pad is buried. Higher flux paths are preferred as they necessarily improve coupling however the challenge is to produce these with minimal inductance and pad size. These conflicting requirements have resulted in a pad design evolution where the understanding gained from previous steps is applied to new topologies resulting in a novel coupler topology described in this chapter and referred to as a Double D.

This chapter begins with an investigation into the flux paths produced by circular pads and it is shown that the topology results in poor coupling if the air gap between a matched set of pads is greater than ¼ of the pad diameter. In contrast shaped bar pads are an alternative to circular pads [109] and as shown in this chapter they offer improved coupling. Herein a structure referred to as a flux pipe is developed and it consists of two coils electrically in parallel to lower the impedance seen by the power supply but magnetically in series to maximise the flux path height. Presented measurements show the shaped bar is very sensitive to aluminium shielding therefore a backing plate and ring cannot be used in the same manner as with circular pads. The presented Double D is much less sensitive to aluminium shielding which ensures it has a high native quality factor, this is essential to ensure efficient systems with reduced magnetic field leakage can be developed.

A second approach that is implicitly based on flux paths about pads is used to show that the DD topology fundamentally provides better coupling than the circular topology. All power pads are inductors therefore store energy in the magnetic field around the coil, note the field behind DD and circular pads is low due to the aluminium shielding. Pads that have the same inductance clearly store the same amount of energy, assuming the currents are the same but
the distribution of the energy is the key to achieving good coupling. Pads that store most of their energy close to the coupling face of the pad will necessarily have poor coupling whereas pads that have favourable fundamental flux paths that enable energy storage further away will have improved coupling. The effect is twofold as systems with identical pads are magnetically symmetric; a receiver that has a high fundamental flux path will be able to extract energy from further away resulting in significantly higher power transfer. The energy stored depends on the current though the pad and its inductance and in this chapter the inductances are matched to ensure the same energy storage at the same current.

4.2. Flux paths

The fundamental shape of the magnetic flux paths above a pad is due to the geometrical layout of the coil and ferrite therefore the shape is largely independent of any design optimisation and can be used to distinguish pad topologies. The height of the flux path determines the coupling ability of a pad topology with larger air gaps (>100mm). As stated above, higher paths are favourable as more flux available further away from the pad enabling higher coupling coefficients. Higher flux paths are longer resulting in a reduction in the magnetic field strength; however there is an optimal operating condition where there is a gain in coupling due to higher but weaker flux paths and this is ultimately reflected in the coupling coefficient. The flux path height depends on the size of the coupler therefore when making comparisons between various proposed designs in this chapter, the size of all of the couplers is kept the same.

4.2.1. Flux paths of circular and shaped bar couplers

In order to illustrate the concept of flux path heights of various topologies, 2D FEM simulations can be used. The height of the flux paths above a circular pad are in the order of one quarter of its diameter ($P_d$), as shown by the dashed line on a plane through the centre of a generic transmitter pad in Figure 3-1(a). The flux (for argument sake) passes through the ferrite strip, exits at the outer end and traverses a “semi-circular” path toward the inner end of the strip. The radius of the flux path is related to the length of the ferrite strips and this depends on the pad diameter. Assuming the pad size is fixed, only small changes in the height of the flux path can be achieved by increasing length of the strips – eventually all of them meet in the middle of the pad – various designs are shown in Figure 3-17 in Chapter 3.4.4.2. Using fewer longer strips is not practical as the volume of ferrite reduces thereby reducing
the coupling. The larger distances between the flux lines of Figure 3-1(a) indicates the flux density drops significantly at the desired operational height of 200mm (shown by the dashed line) resulting in a fundamentally low coupling coefficient.

The fundamental flux path height necessarily makes circular pads large if desirable coupling values with relatively large air gaps are required. The 700mm pad developed in Chapter 3 had a \( k \) of approximately 0.17 with an air gap 175mm, which is equal to \( P_d/4 \). If the same coupling coefficient is required with a 200mm air gap, the diameter must increase by an amount proportional to 4 times the increase in vertical height, which is 100mm. Such large increases in pad diameter make IPT charging systems based on circular pads potentially impractical and expensive for large air gaps and high power levels. This is especially important as shorter charge times are highly desired.

![Figure 4-1: Flux paths of (a) a circular pad and (b) a bar pad. Flux density above (c) a circular pad and (d) a bar pad.](image)

The investigation into an alternative inductive coupler topology began by examining modified bar shapes that have been used successfully in distributed IPT systems. These systems consist of an elongated track, which is typically embedded in the ground, and a rectangular ferrite bar with a coil wound along its length forming a pick-up [44]. The key difference between a circular pad and a flat bar is that the latter’s fundamental flux path height is in the order of half of its length (\( P_l/2 \)) as shown in Figure 3-1(b). Noting both the circular shaped pad and bar pad in Figure 3-1(a) and (b) are energised with identical currents, the flux lines 200mm above the bar pad are much closer together indicating better coupling.
(N.B. as shown here there are two coils wound on the bar and the reasons for using this arrangement are explained in section 4.3).

The magnetic flux density on the dashed line 200mm above the circular and bar pads is plotted in Figure 3-1(a) and (b) respectively. Both pads are the same length however the highest and lowest flux densities above the bar pad are $650\mu T$ and $450\mu T$ respectively and the corresponding highest and lowest values for the circular pad are $400\mu T$ and $200\mu T$. Although the fundamental path height of the shaped bar is twice that of a circular pad, the flux density is not twice as large because the path length is twice as long resulting in a reduced flux density. The reduction in the flux density above the bar pad is less than 50% resulting in a net gain in coupling illustrating the importance of the fundamental differences in the flux path heights of the two topologies. The flux density is more uniform above the bar type pad indicating less sensitivity to horizontal misalignment in the $x$-axis than the circular pad, while the greater magnitude ensures increased power transfer.

4.2.2. Shaped bar couplers

A simple bar as depicted in Figure 3-1(b) is not overly tolerant to horizontal misalignment, however this can be improved by adding wings to the ends (ideally with a minimal amount of extra ferrite) resulting in a shaped bar pad as shown in Figure 4-2(a). The exact shape was investigated experimentally in [109] and then evaluated here using magnetic simulation. The dimensions of the shaped bar pads were based on the coupling performance of circular pads – a coupling factor of 0.2 with an air gap of 200mm is desired enabling a 20kHz system to transfer 2kW. As shown in Chapter 3.4.3, a circular pad would need to be 900mm in diameter in order to achieve this. Based on the understanding of the flux paths discussed earlier, a pad measuring 550mm long was constructed with the size being partly determined by the available I120 cores. A photo showing the transmitter and receiver pads with the aluminium backing plate removed from the receiver is given in Figure 4-2(a).

Inductance measurements were taken with an LCR meter and $k$ was found to be 0.07 with a 200mm air gap, this less than half of that of the 700mm circular pad with the same air gap. The poor coupling is due to the low reluctance short flux path around the coil shown by the dashed line in Figure 4-2(b). The coil only covers a small portion of the midsection of the pad making the ferrite wings largely redundant. Flux is not forced to the ends of the pads where it could be launched to create the desired flux path, also shown in Figure 4-2(b).
Chapter 4 – Flux paths and large air gap power pads

Figure 4-2: (a) Bar pads with back plate removed from pick up and (b) flux paths produced by a bar pad.

The shaped bar pad may be thought of as one of the radial spokes of ferrite in a circular pad therefore the optimal coverage would be around 40% as discussed in Chapter 3.4.1. Achieving this relatively high proportion of coverage in a 550mm long pad with a tightly wound coil (55 turns of 4mm diameter wire) would result in an overly large inductance requiring an impractically high voltage to obtain the desired transmitter pad current. Adding internal compensation capacitors to reduce the inductance seen by the supply is possible, but can also make the pad difficult to drive, and to construct — fewer connections are preferred.

Power supplies are typically designed to drive a pad inductance with up to 15% variation, and any excessive deviation from the desired inductance causes the inverter bridge to source large reactive currents that increase losses and may damage the switches [109, 110]. A transmitter pad inductance typically varies by 10% as the receiver pad is brought within the targeted operating zone and moved horizontally. The series compensation is fixed and as such the relatively small variation in pad inductance becomes a large percentage change in impedance seen by the power supply making the system impractical as it is extremely sensitive to receiver movement.

An additional problem is that the length of wire required to cover the whole midsection would naturally increase cost and copper loss and make the design less cost effective. A compromise is to wind a large pitch coil however it has been experimentally determined that the effective coverage of the midsection decreases if the pitch is greater than a few wire diameters due to excessive flux leakage through the gaps between the turns. These small flux loops around spread conductors do not increase coupling and add significantly to the pad leakage inductance and consequently reduce the coupling coefficient. These conflicting requirements have resulted in the development of a flux pipe as described in the next section, which provides a natural solution to the above problems.
4.3. The flux pipe concept and implementation

A practical solution to achieving a desirable flux path with a shaped bar pad while
minimising copper use is to use two coils positioned at the ends of the midsection as shown
in the schematic of Figure 4-3(a), noting that both the transmitter and receiver pads are
identical. The ends of the pads are made up of wings that are referred to as the pole faces and
these surfaces transmit and receive flux between pads. The coils are electrically connected in
parallel to lower the impedance seen by the power supply and are considered magnetically
connected in series since part of the flux from one coil passes through the other. The mutual
flux between the two coils on a transmitter (or receiver) pad is referred to as intra-pad flux
and is measured by the intra-pad coupling coefficient \(k_{ip}\). A proportion of the intra-pad flux
forms the inter-pad mutual flux, which links with the receiver to couple power.

The intra-pad coupling coefficient can be increased by moving the coils closer together
however if they are too close the result is a lower effective pad length and in the extreme case
the flux path is undesirable as shown in Figure 4-2(b). Conversely, if the coils are moved too
far apart \(k_{ip}\) drops, ultimately reducing the coupling to the receiver.

A prototype flux pipe was constructed with standard rectangular ferrites as shown in Figure
4-3(b). The flux pipe is four bars wide to prevent saturation and each coil comprises 15 turns
of 4mm diameter Litz wire with an area of 6.36mm\(^2\), allowing a transmitter current of 23A.
The intra-coil coupling factor for a built pad is 0.60. The position of the coils (and hence \(k_{ip}\))
has been physically adjusted to maximise the coupling coefficient between the transmitter
and receiver. The remaining 40% of the flux that does not link the two coils on a pad is
leakage flux as shown in Figure 4-3(a). Note the ferrite cores have been staggered so that
there is no continuous (air) gap in the direction of the flux in the midsection (along the length
of the pad). This helps to reduce fringing flux that would otherwise cause loss in the
aluminium backing plate.
4.3.1. Simulation model of a flux pipe

The layout and dimensions of a 3D FEM model of a flux pipe is shown in Figure 4-4(a) and this was created based on readily available I93 cores. The position and structure of the flux launching wings has been optimised to increase the horizontal tolerance to misalignment between the primary and secondary pads in the y-axis.

Magnetic flux density vectors in an xy plane midway through a transmitter pad are shown in Figure 4-4(b). The flux density is reasonably uniform in the flux pipe and the relatively high flux in the wing shows it is operating effectively. There are (four) small areas where the wings join the flux pipe, creating a sharp corner and this results in an increased flux density, however magnitude of the flux density is still below the saturation point of the ferrite and is unlikely to create hot spots as the volume of ferrite involved is very small [111].
4.3.2. Performance of flux pipe pads compared to circular pads

Vertical and horizontal profiles were taken with the simulation model of the flux pipe pads and are shown in Figure 4-5(a) and (b) respectively, with results for a circular pad included for comparison. As shown, with air gaps of less than 125mm the circular pads have a greater coupling coefficient because they produce a relatively uniform magnetic flux above the surface (as expected from Chapter 3.3.1) and they have 1.7 times the area of the flux pipe pads. However, the flux density above the circular pad reduces rapidly due to the low fundamental flux path heights and is reflected in the steep gradient of the vertical coupling profile. As expected, the high flux paths created by the flux pipe enable higher coupling factors with air gaps greater than 150mm; at 200mm $k$ is $\sim 0.2$ whereas it is only 0.15 with the circular pads. In practice this means the power transferred is $\sim 1.6$ times higher than that possible with the 700mm diameter circular pads.

The flux pipe pads are naturally polarised because of their topology, therefore two horizontal profiles are needed to characterise the performance of a pair of matched pads. For this work, the length of the flux pipe pad is assumed to be along the $x$-axis as shown in Figure 4-15(a). Offsets between matched pads that have both $x$ and $y$ components lie in between the bounding orthogonal $x$ and $y$ profiles as shown in Figure 4-5(b). The flux pipe has extremely good tolerance to misalignment in the $y$-axis however the rate of reduction in its coupling in the $x$-axis is similar to that of circular pads, although this should not present a problem if the pad is oriented correctly on the vehicle as drivers naturally have better control in one direction depending on the type of parking space.

The flux pipe pads allow a charging power of 2kW within an elliptical zone that has a minor axis of $\sim 200$mm and a major axis of at least 250mm. This tolerance is sufficient to enable parking, without assistance for an EV. In consequence, the flux pipe pads offer far better coupling with a lighter and more cost effective design compared to the 700mm diameter circular pads. Copper losses are significantly reduced since only 42% of the length of wire used in a circular pad is needed. The ferrite is utilised far more effectively in the flux pipe and the power transfer density is significantly higher (VA coupled per cm$^2$ of pad area). However because the coil must be wound around the ferrite core there are significant losses in the backing plate.
4.3.3. Adding aluminium shielding

Aluminium shielding is necessary for power pads for either or both of the following reasons: to reduce sensitivity to their surroundings such as a steel chassis on an EV, and to meet magnetic field emission guidelines as described in Chapter 3.8. However, the magnetic field present within a pad induces eddy currents in this aluminium resulting in some power loss and a slight reduction in inductance.

The native quality factor ($Q_L$) of a transmitter pad (in this case) can also be interpreted as the ratio of the energy stored in the pad to the energy lost per radian at the operating frequency. The power loss can be calculated from (4-1) where $VA_{in}$ is the input VA at the primary pad terminals and $R_{ac}$ is the ac resistance of the pad:

$$P_{loss} = I_1^2 R_{ac} = \frac{\omega L_1 I_1 I_3}{Q_L} = \frac{VA_{in}}{Q_L}$$

(4-1)

Aluminium backing plates and shielding rings have been used on the 700mm diameter circular pads presented in Chapter 3 and their addition had little impact on the inductance and $Q_L$. The measured $Q_L$ of a transmitter pad was 291 at 20kHz and the inductance was 542μH. With an $I_1$ of 23A the loss is 124W and therefore this has a small effect on the efficiency of a 2kW system.

The flux pipe however, was experimentally found to be very sensitive to any shielding and this presents a problem as the pads cannot be used in practical applications where there are metallic materials around, without such shielding being present [109].
4.3.3.1. Effect of aluminium on the native quality factor

A physical flux pipe pad 558mm long was built to test the effect of aluminium shielding on $Q_L$. The pad used five strips made up of either five or six 193 cores and two 21 turn coils spaced 155mm apart as shown on the plan view in Figure 4-6. An aluminium backing plate marginally wider and longer than the pad was placed underneath with distances ranging from 0mm to 30mm indicated by $\delta_{Al}$ in the front elevation of Figure 4-6. The measured inductance and $Q_L$ values are presented in Table 3-1, the measurement ‘30 with end plate’ refers to a backing plate positioned 30mm away and an aluminium end plate placed at one end of the flux pipe as shown in the front elevation of Figure 4-6 – this is similar to placing a ring around a circular pad.

The inductance measurements are for the two coils connected in parallel. These are energised at 23A each so $I_1$ needs to be doubled to 46A when calculating the power loss. The inductance and $Q_L$ for a pad in air are 205μH and 260 respectively resulting in a loss of 210W. The coupling coefficient of the flux pipe is higher than the circular pad (as shown in Figure 4-5(a)), however there is only a gain in system efficiency with air gaps greater than 200mm. With an air gap of 200mm the flux pipe couples ~1.6 times more power than the circular pads for the same $VA_{in}$ but the power loss is ~1.7 times greater. Flux pipes created with coils wound about ferrite bars become more efficient compared to circular pads with larger air gaps but the power transferred becomes too small to be practically useful.

<table>
<thead>
<tr>
<th>$\delta_{Al}$ (mm)</th>
<th>$L$ (μH)</th>
<th>$Q_L$</th>
<th>$P_{loss}$ (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>No Al.</td>
<td>205</td>
<td>260</td>
<td>210</td>
</tr>
<tr>
<td>0</td>
<td>126</td>
<td>110</td>
<td>305</td>
</tr>
<tr>
<td>3</td>
<td>133</td>
<td>118</td>
<td>300</td>
</tr>
<tr>
<td>10</td>
<td>138</td>
<td>130</td>
<td>282</td>
</tr>
<tr>
<td>20</td>
<td>148</td>
<td>140</td>
<td>281</td>
</tr>
<tr>
<td>30</td>
<td>157</td>
<td>156</td>
<td>268</td>
</tr>
<tr>
<td>30 with end plate</td>
<td>153</td>
<td>86</td>
<td>473</td>
</tr>
</tbody>
</table>

Both the inductance and $Q_L$ drop significantly when an aluminium sheet is placed directly under the pad as shown in Table 3-1. The power loss decreases as the sheet is moved further away however even with a 30mm space, high system efficiency cannot be ensured. Assuming
the plastic lid of the pad and backing plate are both 4mm thick the pad would be 62mm thick making mounting on an EV chassis challenging. Adding an end plate has the greatest negative impact on both $L_1$ and $Q_L$, and here the $P_{\text{loss}}$ is the highest at 473W. This result is not unexpected given an intrinsic feature of the flux pipe is that it is essentially a flattened solenoid. As such a significant proportion of the flux enters and exits from the ends of the flux pipe and this is perpendicular to the end plate resulting in large eddy currents that reduce the inductance and cause excessive loss.

4.3.3.2. Two sided flux path of Flux pipe pads

Circular pads have a relatively low fundamental flux path height compared to the flux pipe pads described but despite this they have a very desirable single sided magnetic field that enters and exits from the front of the pad. This is as shown by the flux density contours on a cross-section of a circular pad with the aluminium backing plate removed in Figure 4-7(a). The magnetic field is single sided because the coil sits on ferrite and this creates a low reluctance path that channels most of the flux away from the backing plate and out the front of the pad. Thus the plate only stops leakage flux out the back rather than the main flux, which couples to a receiver. It is noted that radial ferrite spokes are used in the practical circular pad designs presented in Chapter 3 therefore there is a significant proportion of the backing plate that is not shielded by ferrite however the $Q_L$ value of 291 indicates the loss in the backing plate is low as the ferrite strips are purposefully spaced to capture the majority of the flux.

The solenoidial nature of the flux pipe means it produces a two sided flux path when the aluminium backing plate is removed. When a receiver is not present the flux out the front is the same as the flux out the back as shown in the cross-section of Figure 4-7(b) – the mmf driving these front and back fluxes is the same. The rearward flux path must be impeded with aluminium to avoid unwanted coupling to other objects and this results in extremely high eddy current loss and a decrease in inductance as shown in Table 3-1.

The built flux pipe was able to achieve far higher coupling coefficients than the circular pad with large air gaps even though the mutual flux, which is a proportion of the flux out the front, in less than half of the leakage flux out the back. This further illustrates the importance of the height of the fundamental flux paths.
Chapter 4 – Flux paths and large air gap power pads

A 3D FEM model of a flux pipe pad system was created to show where the greatest eddy current losses occur. An appropriately shaped base plate was added along with a shielding ring around the perimeter. A current density contour is shown in Figure 4-7(c) and illustrates the greatest loss occurs in the shielding ring. There is also significant loss directly under the coils – in practice there should always be a gap between any coils and the shielding material.

The two-sided flux paths mean that this flux pipe topology cannot be used in practice because of its sensitivity to shielding or surrounding materials. In addition, the flux out the end of the pad traverses a long path as it needs to turn around in order to enter the opposite end, and this will make compliance with magnetic field exposure guidelines difficult. The flux pipe topology in the form shown is unsuitable for charging EVs and as such has not been developed further, however it has shown two key points essential for developing pads:

- High fundamental flux paths are necessary for coupling with large air gaps
- Aluminium shielding cannot be used to restrict main flux paths or the losses will be unacceptably high.

![Diagram of flux paths and current density](image)

Figure 4-7: (a) Magnetic flux density contour on a cross-section of a circular pad, (b) magnetic flux density contour on a cross-section of an unshielded flux pipe pad and (c) current density in an aluminium case of a flux pipe transmitter pad ($I_1 = 23$A/coil at 20kHz).

4.3.4. Improving the front to back flux ratio

The ratio of the flux out the front of a pad to the flux out of its back is unity for the flux pipe topology shown in Figure 4-7(b). A topology with more flux out the front than the back and a flux path height proportional to half of the coupler length is sought. In order to change the
front to back ratio, auxiliary coils were added to a flat bar pad as shown in Figure 4-8. A prototype with a 29 turn main coil and two 8 turn auxiliary coils was built with an aluminium shielding plate underneath. The inductance of all coils in series was measured to be $579\mu\text{H}$ and $R_{ac}$ was $270\text{m}\Omega$ at 20kHz corresponding to a $Q_L$ of 270. Adding a shielding ring around the perimeter of the pad caused $Q_L$ to drop to $\sim150$ – although this is higher than the original flux pipe it is still low compared to a circular pad indicating poor efficiency. While the ratio of the front to back flux is improved (as indicated by a higher $Q_L$ value) the main coil is still a solenoid and produces a significant amount of undesirable back flux. Notably the improvement was only due to the auxiliary coils.

![Figure 4-8: Auxiliary coils added to a flat bar pad.](image)

**4.3.5. Equipotential surface**

A practical method for minimising flux out the back of a coupler is to avoid producing it in the first place – this can be achieved by removing the main coil of Figure 4-8 leaving a coupler topology that still has a fundamental flux path height proportional to half of the pad length. The remaining auxiliary coils were placed on a ferrite sheet with raised blocks at the centre of each coil as shown in Figure 4-9. This topology is referred to as an equipotential surface due to the relatively uniform flux distribution above the pad. The raised blocks are added to improve coupling in a similar manner to the ferrite feet added to circular pads in Chapter 3.4.2. Coplanar coils on a ferrite sheet do not produce any mmf that can drive flux out the back of the pad allowing aluminium shielding to be placed underneath (like a circular pad) creating a truly single sided flux coupler. The measured $Q_L$ of a prototype pad was 350 at 20kHz. Ideally shielding would not be required on the back of the pad but fringing flux between ferrite pieces that make up the sheet will cause loss in higher resistivity ferromagnetic materials such as steel.
Although the equipotential surface is truly single sided, the fundamental flux paths are not ideal and will necessarily result in a low coupling coefficient. The inductance ($L$) of the equipotential topology is made up of the inductance of each coil, referred to as coil ‘a’ and coil ‘b’, and the mutual inductance between them because they are connected in parallel and aiding (current flows in the winding direction shown in Figure 4-9). Inductance is due to flux linkage in a coil (of a certain number of turns) for a given current and therefore it can be described by considering the following fluxes: $\Phi_a$ the flux about coil a, $\Phi_b$ the flux about coil b, $\Phi_{ab}$ the flux in coil a due to coil b and $\Phi_{ba}$ the flux in coil b due to coil a. These fluxes are shown in Figure 4-9.

The mutual flux between the two coils is referred to as the intra-pad flux ($\Phi_{ip}$) and is illustrated by the large arc in Figure 4-9. A portion of $\Phi_{ip}$ couples to a receiver pad thereby transferring power and therefore this flux should be relatively large. An effective measure for quantifying coupling between the coils is the intra-pad coupling coefficient ($k_{ip}$) and it is the ratio of the mutual inductance between the coils to the total flux produced by the coils. In an equipotential coupler, $\Phi_{ip}$ is necessarily small due to the physical separation between the coils and because the section of coil that contributes to $\Phi_{ip}$ is small (as shown by the shaded region in Figure 4-9), this will result in a very low coupling coefficient between a transmitter and receiver. These reasons indicate that although this pad topology exhibits useful features, it is not suitable for IPT charging systems due to the lower overall coupling than desired.

**Figure 4-9:** Equipotential surface made up of co-planar coils on a ferrite sheet with main fluxes illustrated.

### 4.3.6. Simulation model with shaped coils

The intra-pad coupling coefficient of an equipotential surface can be increased by making the coils square and moving them closer, forming a distinct midsection. This is referred to as a
modified equipotential surface and a simulation model comprising an identical transmitter and receiver was made to investigate coupling performance. Ferrite sheets measuring 758mmL x 411mm x 16mmT were used as shown in Figure 4-10. Each coil consists of 20 turns of 4mm diameter wire, made square to simplify the model. The ferrite sheet extends past the coils by 10mm in the x and y-axes. The sides of each coil that are in the centre of the pad share a common boundary and therefore contribute to the intra-pad coupling coefficient. The remaining three sides of the coils are considered the return portions as they do not contribute to $k_{ip}$ – two sides are perpendicular to the midsection and the remaining one is far away and the current direction is opposing. The inner areas of both coils are referred to as the pole faces as the inter-pad flux (between Tx. and Rx.) passes through these surfaces.

![Simulation model of improved equipotential surface Tx. and Rx. pads.](image)

The results of a simulation with an air gap of 200mm are summarised in Table 4-2 where the left column contains data that relates to the inter-pad coupling (between primary and secondary) and the right contains data for the intra-pad coupling. Both the transmitter and receiver are identical and therefore have the same self inductance of 789μH ($L_1=L_2= L_{pad}$). Note in the simulation the coils are wound with one wire but in practice a bifilar winding would be used to lower the inductance and in order to maintain the same $NI$ in the transmitter, $I_1$ needs to be doubled. The coupling coefficient of 0.215 enables an uncompensated power of 2.4kVA, which is far higher than that possible with similar sized circular and flux pipe pads.

The intra-pad coupling coefficient of the modified equipotential surface is 0.154 and this is lower than the $k$ between pads indicating that the majority of the coupling is not due to flux on the fundamental flux path (shown in as $\Phi_{ip}$ in Figure 4-9). A significant proportion of the
coupling coefficient between pads is due to the return portions of the coil making the modified equipotential surface act like two circular pads placed close to each other.

| \( V_{oc} \) | 489 V | \( L \) (coil) | 341 \( \mu \)H |
| \( I_{sc} \) | 4.84 A | \( M_{ip} \) | 52.6 \( \mu \)H |
| \( P_{su} \) | 2418 VA | \( k_{ip} \) | 0.154 |
| \( M \) | 169 \( \mu \)H |  |
| \( L \) (pad) | 789 \( \mu \)H |  |
| \( k \) | 0.215 | \( I_1 = 23\)A at 20kHz |

**TABLE 4-2 MODIFIED EQUIPOTENTIAL PAD PARAMETERS AND INTRA-PAD COUPLING**

Magnetic flux density vectors in the \( yz \) and \( zx \) planes are used to show the magnetic field shape between pads in Figure 4-11(a) and (b) noting the origin of the axes is at the centre of the coupling surface of the transmitter. The \( yz \) plane intersects the \( x \)-axis at 200mm thus it is positioned close to the centre of the pole face. Note the flux density scale has been adjusted to a maximum of 5mT to better show the flux in the air gap. As shown in Figure 4-11(a) there is a large amount of flux around the return portion of the coil and although some flux does couple to the corresponding section of coil in the receiver (mutual) flux, a large proportion is leakage flux and this contributes to the pad leakage inductance.

Note there is some flux cancellation close to the ends of the ferrite sheet shown in Figure 4-11(b); this is because the current direction in the return portion of the coil is opposite to the current direction in the centre. The position of the cancellation is biased to the end of the pad due to the increased \( NI \) in the centre. There is no flux directly out of the back of the pads however there are flux paths that travel behind the pad in order to link the return portion of the coil winding, but these are long and therefore the flux density is low.

The leakage inductance of a modified equipotential coupler can be reduced by removing the ferrite from under the return portions of all the coils, and this will favourably increase \( k \) as the inductance of the coil will reduce but \( M \) remains essentially unaffected. The results for pads with ferrite removed are shown in Table 4-3, where the first column contains parameters for inter-pad coupling and the second for intra-pad coupling. There is almost no change in \( M \) and there is a reduction in \( L \) from 789\( \mu \)H to 706\( \mu \)H, which increases \( k \) to 0.24.

The coupling mechanism between the pads is dramatically changed and this results in a new pad topology that is very different to the equipotential surfaces. The return windings on the transmitter and receiver do not couple directly but rather increase the magnetic potential
around the pad forcing the mutual flux to the receiver in both the \(xz\) and \(yz\) planes in a similar manner as shown in Figure 4-11(a) and (b), but this is done with minimal added inductance.

![Diagram](image)

**Figure 4-11:** (a) Magnetic flux density vectors in \(yz\) plane \((x=200)\) and (b) flux density vectors in \(xz\) plane \((y=0)\) \((I_1 = 23A/\text{coil})\).

As shown in Table 4-3 the intra-pad coupling coefficient has increased to 0.47 indicating that the coupling between pads is mainly due to flux on the desired fundamental path illustrated in Figure 4-9. The large increase in \(k_{ip}\) is due to a more than double increase in the intra-pad mutual inductance \((M_{ip})\). This is far greater than expected given the coils are close together forming the midsection for both types of pad pairs (with a solid ferrite sheet and with ferrite removed from under the return portions). The increase in \(M_{ip}\) is due to favourable magnetisation of the ferrite shown by magnetic flux density vectors in the \(xy\) plane through the centre of the ferrite sheet for both pads in Figure 4-12(a) and (b). Note for both simulations the receiver is present with a vertical offset of 200mm, and is open circuited.
Chapter 4 – Flux paths and large air gap power pads

The pad of Figure 4-12(a) is a modified equipotential surface transmitter pad with an active left coil and an open circuited right coil. The flux density scale has been set to a maximum of 20mT. The greatest flux density occurs in the midsection of the pad as shown by the red vectors. This is due to the presence of ferrite on both sides of the coil enabling greater flux linkage than on the return portions.

The large volume of ferrite enables flux from the energised coil to spread out across the entire slab causing a significant amount of common mode flux in the adjacent coil. The density of the common mode flux is relatively uniform across the area of the coil on the right as shown by the green to light blue vectors. In consequence, the intra-pad coupling is due to a small change in the flux density hence $k_{ip}$ is only 0.15.

There is a change in the direction of the flux density vectors toward the left of the excited coil as labelled in Figure 4-12(a). This is due to opposing flux caused by an effective change in current direction (labelled as $I$) due to the coil structure. This creates a flux reversal and it is positioned to the left of the pad because of the increased flux density in the midsection due to the relatively low reluctance paths around coil (it is surrounded by ferrite).

Magnetic flux density vectors in a modified equipotential surface pad with ferrite removed from under the return portions of the coils is shown in Figure 4-12(b). The magnetisation of the ferrite is preferable as it channels a significant amount of flux from the excited coil through the neighbouring coil increasing $k_{ip}$ from 0.15 to 0.47. Also, the flux about the return portion of the excited coil is reduced thereby reducing the leakage inductance significantly.

Almost all of the magnetic flux shown in the central plane of the ferrite sheet is in the $x$-axis direction. There is only a slight divergence at the ends of the sheet thus the flux density is essentially described by $B = B_x$ and a similar flux pattern would be produced with an anisotropic material with the magnetisation direction in the $x$-axis. This is very different to

<table>
<thead>
<tr>
<th>$V_{ac}$</th>
<th>489 V</th>
<th>$I_1$ (coil)</th>
<th>240 µH</th>
<th>$M_{ip}$</th>
<th>114 µH</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_{ac}$</td>
<td>5.52 A</td>
<td>$M_{ip}$</td>
<td>114 µH</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$P_{sn}$</td>
<td>2703 VA</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$M$</td>
<td>169 µH</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$I_2$ (pad)</td>
<td>706 µH</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$k$</td>
<td>0.24</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

$\text{Table 4-3 Modified Equipotential Pad Parameters and Intra-Pad Coupling for Pads with Ferrite Removed from Return Portions of the Coils}$

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the modified equipotential surface of Figure 4-12(a) where the flux diverges from the coil. Note the flux density vectors are only shown on a plane in the $xy$ axis that passes through the centre of the ferrite sheet; clearly magnetisation is required in the $z$-axis in order for flux to leave or return to the surface of the ferrite sheet.

The pads of Figure 4-12 both have ferrite sheets that are 16mm thick but the modified equipotential surface of Figure 4-12(a) has a significantly greater ferrite cross-sectional area than the pad of Figure 4-12(b) resulting in a lower maximum flux density of 20mT.

![Magnetic flux density vectors in xy plane (z=16) mid way through ferrite sheet with the left coil energised and the right open circuit (a) full sized sheet and (b) ferrite removed from under the coils ($I_1 = 23A$/coil).]

However, as described above, the additional ferrite area only contributes to leakage inductance and ultimately lowers the coupling factor between two pads (0.24 vs. 0.22). The pad of Figure 4-12(b) has less ferrite, resulting in a maximum flux density of 60mT. The ferrite utilisation is measured in $VA$ coupled per cm$^3$ of ferrite (at an air gap of 200mm in this
case) and for the pad of Figure 4-12(b) it is 0.6VA/cm³. It is 0.24VA/cm³ for pad of the Figure 4-12(a).

Although the ferrite utilisation is higher in a pad with ferrite removed from under the end windings, the actual value is similar to the optimised circular designs of Chapter 3.442 where a value of 0.4VA/cm³ was possible (with a 200mm air gap). That is, even with highly desirable flux paths, the modified equipotential surface uses ferrite only slightly more effectively than circular pads. The pads can be made much lighter and more cost effective by removing ferrite appropriately as discussed in the next section.

4.3.7. Development of Double D type couplers

The maximum flux density of 60mT, shown in Figure 4-12(b), is very low for any pad design in this thesis given the ferrite material used is N87 which has a saturation flux density of 490mT at 25°C [111]. Removing ferrite and operating with higher flux densities will result in a higher ferrite utilisation value – this was done in the optimisation of circular pads in Chapter 3. The main magnetisation in the ferrite slab of Figure 4-12(b) occurs in the x-axis therefore ferrite strips can be used as shown on the schematic of Figure 4-13 while still maintaining good coupling.

The coils in Figure 4-13 are labelled ‘a’ and ‘b’ and the return portions create leakage flux, which is referred to as $\Phi_{bl}$ and $\Phi_{al}$, and intra-pad flux ($\Phi_{ip}$), a part of which couples to the receiver. In order to reduce the length of wire to minimise cost, leakage inductance and copper loss, the return portions of the coils should be made roughly semi-circular and since two such coils are placed back to back, this new pad topology is called the “Double D” or DD. The shaded mid-section of the pad is referred to as a flux pipe (of length $F_1$) because the coils are magnetically in series but connected electrically in parallel to lower the impedance seen by the power supply. The fundamental flux path height, $h_z$, is proportional to half of the length of the pad ($P_x/2$) but it also depends on the length of the flux pipe ($F_1$) therefore this is investigated separately in Chapter 5.
Figure 4-13: Simplified model of a DD pad with main flux paths $\Phi_{al}$, $\Phi_{bl}$ and $\Phi_{ip}$, due to coil a, b and mutual coupling between coils respectively.

A 3D FEM model was created to investigate flux paths produced in a DD pad system comprising an identical transmitter and receiver (vertically aligned) with an air gap of 200mm. The pads measure 0.77m by 0.41m and are ~35mm thick and use four strips spaced 33mm apart comprising six I93 cores; these are referred to as 4x6 pads. Double D pads of this size achieve a coupling factor 0.23 with a 200mm air gap; a more thorough investigation into matching pad size to air gaps is presented in Chapter 5 where pads up to 1.3m by 1.3m are considered. As long as the pads are correctly matched to the operational air gap the flux paths will be similar.

A cross-section of the simulation model of the 4x6 pads is shown in Figure 4-14 with idealised flux paths superimposed. The 240mm long flux pipe channels flux to the pole faces where it is launched on one of four general paths on the $xz$ plane shown. The majority of the flux goes to the opposite pole face (shortest path) forming the air gap leakage flux and this is reflected by the highest flux density on the $\Phi_{la}$ contour. This increased magnetic flux density above the flux pipe supports the mutual flux ($\Phi_M$), which is also in part attracted to the ferrite in the receiver pad.

The remaining two flux paths are largely nuisances as they do not directly or indirectly contribute to coupling but rather to magnetic field leakage, which is discussed in Chapter 6. The flux behind the transmitter pad ($\Phi_{lb}$) is due to linkage of the return portions of the windings, while the flux above the receiver pad ($\Phi_{lt}$) is due to linkage of the flux pipe. The nuisance flux paths are relatively long and as such the flux density in these areas is very low as shown in Figure 4-14.
4.4. Energy distribution about Double D and circular pads

Power pads act as a loosely coupled transformer and transfer power in the following manner: current in the transmitter creates a magnetic field about the pad and energy is extracted by the receiver by the opposite process. The transmitter pad stores energy in its magnetic field and the amount that is available for extraction depends on the inductance of the pad and the current flowing in it. Clearly the larger the amount of energy stored in the field the more energy can be transferred however inductance and current values have practical limits. Excessively large inductances require very high driving voltages to get the desired transmitter pad current but the mains supply is limited and in addition, high terminal voltages present safety hazards. In the preceding sections the fundamental flux path heights of circular and Double D couplers were used to compare topologies. In this section the energy distribution about circular and Double D pads is investigated and used to distinguish between the two topologies.

4.4.1. Importance of pad inductance

The uncompensated power ($P_{su}$) of a receiver pad is the product of the open circuit voltage and short circuit current as shown in (4.2).

$$P_{out} = \frac{\omega M^2 I_1^2}{L_2} Q$$

(4.2)

The square of the transformer coupling coefficient (shown in (4.3)) can be substituted into (4.2) (with an $L_1$ term) resulting in (4.4).
\begin{equation}
k^2 = \frac{M^2}{L_1 L_2} \tag{4.3}
\end{equation}

\begin{equation}
P_{\text{out}} = \omega k^2 L_1 I_1^2 Q \tag{4.4}
\end{equation}

The $L_1 I_1^2$ term in (4.4) is analogous to the equation for the energy stored in an inductor, (first term in (4.5)), indicating that a transmitter pad must have adequate inductance and current flow to enable the desired power transfer. Similarly the receiver pad must also have adequate inductance or it will not be able to extract the available energy however the receiver is ignored for now. The energy stored in the magnetic field in volume $V$ is shown in (4.5) and is equal to the energy stored in an inductor [112].

\begin{equation}
W = \frac{1}{2} L_1 I_1^2 = \frac{B^2}{2 \mu} V \tag{4.5}
\end{equation}

Determining the energy storage within particular elements in a particular volume of a material in a 3D FEM model can be done as described in Chapter 6.6.2 where simulation data is post processed to create a loss model for a Double D transmitter pad. The energy stored in a particular volume of material is found by applying (4.5) to every element within the material and summing the result of all the elements.

4.4.2. Energy storage zone of Double D and circular pads

The power extracted by a receiver pad depends on the operational angular frequency, coupling coefficient, inductance, current and the receiver $Q$ as shown by (4.4). However, considering the distribution of energy about a pad illustrates the fundamental performance difference between circular and Double D topologies. A topology will necessarily be better if more energy is available further away from the pad and given the receiver is assumed to be identical to the transmitter it will also extract more energy since the pads are magnetically symmetrical. As such the energy storage investigation is done without a receiver present.

In order to make a fair comparison of the energy distribution about the DD and circular topologies the pads need to have the same surface area and have the same stored energy. Both pads are of similar thickness as they use the same layer structure of ferrite cores and Litz wire. Practical measurements taken in this thesis (in Chapter 5) use built 4x6 DD pads that are 0.32m$^2$ yet the 700mm circular pads developed in Chapter 3 have an area of 0.38m$^2$. As such the diameter of the pads in the simulation model was reduced to 640mm with a ratiometric reduction in the lengths of the ferrite strips.
Chapter 4 – Flux paths and large air gap power pads

To ensure the total stored energy is the same the pads need to have the same inductance or be operated with adjusted currents as shown by (4.5). The former approach is used here and the pads are operated with a current of 23A. The circular pad is therefore modified slightly as its inductance was slightly lower than that of the DD. The number of turns was increased to 28 and the coil was moved outward by 13.5mm resulting in a coil diameter ($C_d$) of 195.5mm (dimension shown in Chapter 3.2).

Depending on the output power required, these pads can operate with air gaps of 100-200mm therefore energy storage is considered in a zone that begins 5mm above the pad surface and extends 240mm high with the same cross-sectional area as the pad. The zone is broken up into 12 sections creating sheets that are 20mm high for both topologies shown in Figure 4-15. Here the first slice is the one closest to the pad. Note half of the zone is hidden leaving a wireframe to show the position and structure of the transmitter pads. Clearly the energy about the pad varies in accordance with the current but to simplify the comparison the current is assumed to be a steady 23A, thus the difference in the distributed energy is the key to differentiating between topologies.

The inductances of the 640mm diameter circular and 4x6 Double D pads are almost exactly the same as shown in Table 4-4. The total energy stored in each of the pad inductances is shown by $W_{tot}$ while the energy stored in all 12 sheets of the air zone as shown in Figure 4-15 is described by $W_{zone}$ in Table 4-4. The proportion of the total energy stored to the energy stored in the air zone is given as $\%W_{zone}$. For both topologies the energy stored in the air zone is small compared to the total energy, indicating most of the energy storage occurs within a very small volume around the coil noting there is no flux out the back of the pads. The energy density within the complete zone, which has a volume of $0.077m^3$, is indicated by $W/V_{zone}$ in Table 4-4.
Chapter 4 – Flux paths and large air gap power pads

There is slightly more energy stored in the zone above a circular pad indicating that the Double D topology has long flux paths around the pad, which are illustrated as $\Phi_{lt}$ and $\Phi_{lb}$ in Figure 4-14. Although the difference in energy is small, the flux density responsible can cause difficulty with compliance with magnetic field exposure standards as stray fields need to be in the $\mu$T range. This is discussed further in Chapter 6.

<table>
<thead>
<tr>
<th></th>
<th>Circular</th>
<th>Double D</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1$</td>
<td>569.50</td>
<td>569.45</td>
</tr>
<tr>
<td>$W_{tot}$</td>
<td>150.63</td>
<td>150.62</td>
</tr>
<tr>
<td>$W_{zone}$</td>
<td>4.953</td>
<td>4.918</td>
</tr>
<tr>
<td>% $W_{zone}$</td>
<td>3.29</td>
<td>3.26</td>
</tr>
<tr>
<td>$W/V_{zone}$</td>
<td>64.49</td>
<td>64.03</td>
</tr>
</tbody>
</table>

**Table 4-4 Parameters of the Circular and Double D pads with a current of 23A**

The total amount of energy stored in the zone shown in Figure 4-15 is fairly similar for circular and Double D pads however the distribution of the energy within the zone is critical for good coupling. The zones comprise 12 sheets and the one closest to the pad is referred to as sheet 1. The energy stored in each sheet is plotted in Figure 4-16(a). The circular pad stores slightly more energy very close to the pad (in sheet 1) however the rate of decrease in energy storage is greater than that of the Double D. The graph of Figure 4-16(b) shows the ratio of the energy stored in the Double D to the circular pad. The trendline shows the rate of improvement in energy storage above the DD is initially decreasing however a point of inflection is reached at sheet 7, which is situated 140mm above the pad. The DD offers significantly improved energy storage from sheet 8 upward indicating the coupling coefficient will be significantly higher than that of the circular pad with air gaps greater than 160mm.
The energy plots illustrate the circular topology has a low fundamental flux path height and therefore will offer limited coupling with large air gaps. Circular pads may be more suitable for extremely small systems that operate with a very small air gap but these are not considered practical for EV charging. Vertical profiles comparing the coupling coefficient against air gap for DD and circular pads are presented in Chapter 6.8.1.

The slices of Figure 4-15 have the same volume for both pad topologies, therefore the graph of Figure 4-16(a) effectively shows the energy density above the pads. Sheet 11 is positioned 200mm above the pads and the energy density above a Double D is 4.8mJ/m³ and for the circular pad it is 3.5mJ/m³. The transmitter pad must be sufficiently large to produce these large energy densities high above the pad and the receiver needs to be of sufficient volume to be able to extract the available energy. Thus there are fundamental size requirements for pads that operate with large air gaps and this is investigated in Chapter 5.3 where DD pads are matched to operational air gaps ranging from 100-300mm. Small pads cannot achieve high energy densities with large air gaps and will necessarily result in inefficient systems.

### 4.4.3. Performance variables

Performance characteristics of both a DD on DD and a circular on circular (700mm) pads are compared in Table 4-5 for an operational air gap of 200mm.
Chapter 4 – Flux paths and large air gap power pads

<table>
<thead>
<tr>
<th></th>
<th>Circle</th>
<th>DD-DD</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{sw} (\delta_z = 200 \text{mm})$</td>
<td>905</td>
<td>1921</td>
<td>VA</td>
</tr>
<tr>
<td>$k$ (vertically aligned)</td>
<td>0.158</td>
<td>0.231</td>
<td></td>
</tr>
<tr>
<td>$M$</td>
<td>86</td>
<td>128</td>
<td>$\mu$H</td>
</tr>
<tr>
<td>$L$ (Transmitter)</td>
<td>542</td>
<td>589</td>
<td>$\mu$H</td>
</tr>
<tr>
<td>$Q_L$</td>
<td>291</td>
<td>392</td>
<td></td>
</tr>
<tr>
<td>193 cores (Tx. + Rx.)</td>
<td>72</td>
<td>48</td>
<td>pieces</td>
</tr>
<tr>
<td>Litz wire (Tx. + Rx.)</td>
<td>66</td>
<td>86.6</td>
<td>m</td>
</tr>
<tr>
<td>Size</td>
<td>0.70 Dia</td>
<td>0.77x0.41</td>
<td>m</td>
</tr>
<tr>
<td>Surface area</td>
<td>0.38</td>
<td>0.32</td>
<td>$m^2$</td>
</tr>
<tr>
<td>$P_{sw}/\text{Area@200mm}$</td>
<td>2383</td>
<td>6085</td>
<td>VA/$m^2$</td>
</tr>
<tr>
<td>$P_{sw}/\text{Vol. Fe @200mm}$</td>
<td>12.6</td>
<td>40.0</td>
<td>VA/pc</td>
</tr>
<tr>
<td>$P_{sw}/\text{Wire @200mm}$</td>
<td>13.7</td>
<td>22.2</td>
<td>VA/m</td>
</tr>
</tbody>
</table>

TABLE 4-5 PARAMETERS OF THE CIRCULAR AND DOUBLE D PAD SYSTEMS ($I_1 = 23A$ AT 20kHz)

The Double D topology offers a 146% improvement in the coupling coefficient over the larger circular pads enabling an uncompensated power of close to 2kVA. The native quality factor ($Q_L$) of the DD is significantly higher allowing increased system efficiency and the DD pads use all materials more effectively allowing a higher performance to cost ratio.

4.5. Conclusion

The height of the fundamental flux paths above a topology determine coupling performance with relatively large air gaps (100-300mm). Circular pads have a fundamental flux path height that is proportional to one quarter of the pad diameter therefore 900mm diameter pads are required to get a coupling factor of 0.2 at 200mm. Shaped bar pads have been introduced and these have a fundamental flux path height that is proportional to half of the pad length. However, realising high coupling factors is not practical with a simple winding due to the large pad inductance and length of wire required. In consequence, a flux pipe has been developed and it has the advantage of an ideal bar topology where flux enters and exits at the ends permitting the greatest flux path heights with minimal inductance and material requirements. The flux pipe uses two coils connected electrically in parallel and magnetically in series to lower the impedance seen by the power supply.

A 550mm long shaped bar pad with a flux pipe achieved a coupling factor of 0.2 with a 200mm air gap however the topology is essentially a flattened solenoid and produces magnetic flux out the front and back of the pad. Adding aluminium shielding to mitigate the rearward flux results in excessive losses – the native quality factor of an unshielded pad is 260 and it drops to 86 with shielding. Auxiliary coils were added to the ends of the shaped bar to improve the front to back flux ratio with some success however the most practical
method is to avoid producing any mmf than can drive flux out the back of the pad. Coplanar coils on a ferrite sheet produce a single sided flux like circular pads but have the advantage of a fundamental flux path height that is proportional to half of the pad length. The topology was referred to as an equipotential surface due to the uniform flux distribution above it.

The flux linkage of the two coils in a single pad is quantified by the intra-pad coupling coefficient and this value is necessarily low for an equipotential surface. The coils are positioned apart and circular therefore the common edge is small resulting in little flux linkage. A simulation model, referred to as a modified equipotential surface, was built and the coils were made square to increase the common edge. It has been shown that although the coupling factor was 0.22 with a 200mm air gap, the majority of the coupling flux between a transmitter and receiver is not a proportion of the intra-pad flux but rather due to flux created by the return portion of the coils. The intra-pad coupling is low because the ferrite sheet allows the flux to spread creating a significant amount of common mode flux in the neighbouring coil. The intra-pad coupling coefficient was increased to 0.47 by removing ferrite from under the return portions of the coils and identical pads had a coupling coefficient of 0.24. The flux within the remaining ferrite slab is largely in the direction of the length of the pad and given the volume of ferrite is large the maximum flux density is well below the saturation value of the N87 material used. Higher flux densities are preferred as the ferrite utilisation factor, measured by the VA coupled per cm$^3$, is then generally higher. The magnetisation direction means that ferrite strips can be used along the length of the pad with little reduction in the coupling coefficient but this enables lighter and more cost effective pads. This results in a new pad topology that is called the Double D and it offers single sided flux paths and a fundamental flux path height proportional to half of the pad length.
Chapter 5. Optimisation of Double D power pads

5.1. Introduction

The relationships between the important design variables of the Double D pad introduced in Chapter 4 are investigated in this chapter and subsequently pads are designed for a 2kW charging system operating with a 200mm air gap. The design objective is the same as that for the circular pads of Chapter 3 – maximising coupling while minimising the use of Litz wire and ferrite to increase cost efficiency. Double D pads are naturally polarised and as such the evaluation of how to scale up the power transfer from a base design must take into account expansions in either the length or the width or a combination of the two. This makes pad selection challenging because designs exist on a 2D plane with pad length and width as the axes. Conversely, circular pads are not polarised thus scaling a base design ratiometrically only involves altering one dimension (the diameter) because designs exist on a 1D line.

As shown in Chapter 3, there were numerous ferrite and coil variables that needed to be considered (such as the length, width, and volume of ferrite or the position of the coil) and all of these have an effect on material usage, size and coupling. Correlating all the variables is computationally intensive and provides little insight into the operation of the pads. The approach used in Chapter 3 where the coil and ferrite related variables were separated and considered independently is used here. The process is iterative and the points learned from the previous analysis are used to create a base pad for further investigation.

A similar ratiometric scaling approach is used here as was used in Chapter 3 to help understand and optimise the Double D topology. However, considering all the ferrite variables such as width and thickness along with 2D size changes is beyond the scope of this thesis. As such, a heuristic approach is used based on experience gained in the circular pad optimisation process in order to limit the design options. Readily available ferrite cores and Litz wire are used and only identical pad sets are investigated. It is intuitive that transmitter and receiver pads of different sizes can be used to lower power transfer and improve tolerance to horizontal misalignment and this may be desirable in some applications that have very small air gaps. However, such asymmetrical systems require a comprehensive investigation and thus are not considered here.
To limit the range of ferrite size options, a standard I93 core is used, similarly the diameter of the Litz wire is fixed to 4mm. These are practical restrictions given the cores are readily available as part of EI sets and are an appropriate size for pads transferring 2-7kW over a 100-300mm air gap. Each ferrite strip within the DD pads is made up of an integer number of cores to avoid any difficulties and costs associated with cutting ferrite. Using smaller cores becomes impractical due to the large number of mating surfaces required to build up a sufficient pad size, in addition the mating surfaces also need to be ground to minimise air gaps, resulting in increased cost.

Excessively large pieces of ferrite in a receiver are prone to cracking with the shock and vibration experienced on a vehicle and they are also expensive to manufacture. These cracks can introduce small air gaps that lower the inductance of the pad affecting tuning and power transfer. Making a pad properly shockproof adds to the cost substantially. The I93 cores measure 93mmL x 28mmW x 16mmT and offer a good compromise between the small and large core size extremes. The joints within a ferrite strip comprising several I93 cores allow the pad to flex and contract and expand (due to vibration and varying climatic conditions) without placing mechanical stress on the ferrite. The Litz wire consists of 810 strands of 0.1mm diameter wire providing a cross-sectional area of 6.36mm². This enables a transmitter pad current of 23A without excessive resistive loss and allows a reasonably large number of amp turns to ensure adequate power transfer.

The first Double D pad was designed to transfer 2kW with a 200mm air gap and the dimensions were partly based on the flux pipe pads presented in Chapter 4 and [106], and scaling up an early prototype that was used to test the DD concept via simulation. The early results are not explicitly presented but are encompassed in section 5.3.4. The width of first DD pads was increased to 410mm to ensure that the area enclosed by each coil was sufficient to capture the desired flux over the intended gap. The base of the pad was constructed using 4 ferrite strips comprising 6 x I93 cores with 33mm spaces between the strips. This spacing was chosen because the ferrite cores are 16mm thick, thus flux entering each gap is able to enter into the side of a ferrite strip through an aperture that is marginally wider than the ferrite thickness (+0.5mm). The coils were each made up of 20 turns.

In the following subsections, all of these initial parameter selections are evaluated and adjusted to determine an optimal performance. The length and width of DD pads is varied independently and combined resulting in numerous pad sizes. Changing only one variable
allows pads a few metres long or wide to be simulated to determine the coupling limits of the topology. Optimal designs are chosen based on the coupling coefficient, pad area and cost. Each of the DD designs are referred to by the number of rows of ferrite strips and the number of cores that make up the length of each strip. The initial base pad is therefore referred to here as a 4x6 (indicating 4 strips, each comprising 6 “I” cores).

5.2. Simulated and measured results

Prior to running optimisation simulations, a measurement versus simulation experiment was undertaken to ensure the model was correct. Vertical profiles were taken and these involved increasing the air gap between horizontally aligned pads. The $P_{su}$, $k$, $V_{oc}$ and $I_{sc}$ profiles were calculated from inductance measurements taken with an Agilent E4980A precision LCR meter and are shown in Figure 5-1 below.

Figure 5-1: Comparison of measured and simulated results for 4x6 DD pads against vertical offset (a) uncompensated power, (b) coupling coefficient, (c) open circuit voltage and (d) short circuit current ($I_1 = 23\text{A}$ at 20kHz).
The results are in excellent agreement with a greatest error of 8%, which occurs in the $P_{su}$ and $V_{oc}$ profiles when the pads are very close together. The error in the practical operational range of 180-240mm is only a few percent allowing all pad optimisation to be done via simulation, with confidence.

5.3. Pad design variables

The layout of a simulation model of a DD pad is shown in Figure 3-1 below. The simulated inductance is marginally larger than measured as shown by Figure 5-1(d) and this could be because inductance is related to the area of the coils and the simulated coils use square corners and thus are slightly larger than the built ones. Square corners are used to reduce the complexity of the model and therefore the calculation time however, in practice, the minimum bending radius of the wire results in rounded corners but overall this has little effect on accuracy.

The variables considered to have the greatest influence on coupling between DD pads are shown in Figure 3-1. The length of the pad is governed by the length of the ferrite strips and the number of turns in the coils. The distance between the sides of the coil (the return portions) and ferrite has been fixed at 10mm in both the length and width dimensions. The width is determined by the number of ferrite strips, the spacing between them and the number of turns. The flux pipe was initially constructed by placing each of the turns of wire in the centre together, although it was later found by simulation that the coupling coefficient could be improved by extending the flux pipe simply by separating the wires slightly enabling a 240mm long flux pipe to be constructed. The relationship between spreading of the wires and coupling is investigated in the next section.

The thickness of the ferrite bars has the least effect on coupling because fundamental flux paths are largely unaffected. This is described in the context of circular pads in Chapter 3 section 3.4.2. Thinner ferrite strips permit lighter and more cost effective pads however saturation is more likely. A compromise is to use tapered strips that have thicker cores in the flux pipe and thinner cores at the pole faces; however this investigation is beyond the scope of this thesis and the ferrite strips are kept 16mm thick to enable easy experimental validation.
5.3.1. Number of turns and pitch of coil

The length and hence effectiveness of the flux pipe is governed by the coverage of the midsection. This is expressed as the percentage of the ferrite strip covered by the coil portions forming the flux pipe (shown above in Figure 3-1) and is dependent on the number of turns and pitch of the coil. To illustrate midsection coverage, two arbitrary pads with 18 turns are shown in Figure 5-3(a) and (b). The first pad has closely spaced coils in the midsection giving a coverage factor of 26% while the coil in the second pad has a pitch of 9.3mm covering 59% of the strip length.

Narrow coils with few closely spaced turns create short flux pipes consequently reducing coupling to a receiver at a vertical offset and combined with the low mmf this results in low power transfer. Conversely, if the coils are wound with an extremely large pitch, flux will leak out of the gaps between turns effectively resulting in a poor flux pipe [106]. Using a large number of turns has the desirable effect of increasing the length of the flux pipe however the inductance will be high making driving the pad with 240V equipment difficult. Bifilar and trifilar windings may be used but the excessive amount of copper adds weight and
makes the pad expensive. Clearly an optimal coverage of the mid section in terms of number of turns and pitch exists, and this is investigated via simulation.

A model comprising identical 4x6 DD pads was used and the number of turns of 4mm diameter wire in each coil, (N), was varied from 12 to 26 corresponding to a coverage of the midsection ranging from 20 to 60%. The graph of Figure 5-4(a) shows the coupling coefficient against coverage determined by pitch (dashed line) and number of turns (solid line). The transmitter current was set to 23A at 20kHz and the air gap was 200mm.

Models where the pitch is less than the wire diameter, shown as a hatched area to the left of the solid line in Figure 5-4(a), are presented for completeness. Practical implementation is possible if Litz wire with a rectangular cross section is used however this will undesirably add thickness to the pad. The uncompensated power and inductances for the pad models are shown in Figure 5-4(b) and (c) respectively. The highest coupling coefficient for a given number of turns is obtained when the coverage is between 45 to 50% however the highest $P_{su}$ occurs when the coverage is around 25%.

Pads with high coupling coefficients are preferable, therefore selecting a higher $k$ while allowing a slight decrease in uncompensated power is a good trade-off. In practice pads should be able to couple a desired power level (2kW in this case) within required tolerances, in this case an air gap of 200mm and greater than ±200mm horizontal misalignment from centre to centre. For most applications it is prudent to select a pad that provides $P_{out}$ with a $Q$ of 1 when the pads are vertically aligned and rely on an increase in the operational $Q$ with horizontal misalignment to ensure $P_{out}$ is maintained.

Therefore, DD pads with a $P_{su}$ of approximately 2kVA are desired and should use as little copper as possible. Two pads are preferable as they are within the cross hatched area showing a 5% tolerance boundary on the desired $P_{su}$ as shown in Figure 5-4(b). One pad is labelled as ‘A’ and has 20 turns providing ~38% coverage while the other is labelled as ‘B’ and has 22 turns with ~58% coverage. Based purely on $P_{su}$, the 22 turn pad looks more promising and from a power supply perspective both pads appear the same as they both have an inductance of ~600uH as shown by the shaded markers on the graph of Figure 5-4(c).
Chapter 5 – Optimisation of Double D power pads

Figure 5-4: Varying coverage of the midsection of identical pads. (a) Coupling coefficient, (b) $P_{su}$, and (c) $M$ and $L_2$ ($I_1 = 23$A at 20kHz).
However, selection based on $P_{su}$ gives little insight into overall system cost and efficiency. This is dependent on which side of the respective maxima of the dashed coupling curves of Figure 5-4(a) the design sits on. The 20 turn pad (A) sits to the left of its maximum while the 22 pad (B) sits on the right. Designs operating on the left are preferred as the rate of decrease in mutual inductance is less than the rate of decrease in self inductance as shown on the graph of Figure 5-4(c).

A design at the extreme left of the maxima shown in Figure 5-4(a) is problematic because the self inductance of the pad is very large requiring high terminal voltages. This can be implemented if series compensation capacitance is added to limit the inductance seen by the supply but the additional cost makes this undesirable. The coupling coefficient curves emphasise designs that have fewer turns and these are preferable to minimise the volume of copper and the associated operating loss.

5.3.2. Spacing between ferrite strips

Prior to constructing a pad for experimental validation, an investigation was undertaken using 4x6 pads with 20 turn coils and 43% coverage to determine how the spacing between ferrite strips affects coupling and peak flux density. The space between the strips was varied from 5 to 115mm corresponding to a width variation of 307mm to 637mm. The graph of Figure 5-5(a) shows the coupling coefficient and $P_{su}$ as the strips are spread further apart. The insets in the graph illustrate the minimum and maximum ferrite spacing.

The coupling coefficient reaches a maximum when the strips are spaced 95mm apart however selecting a design based on the highest $k$ is not recommended. The rate of increase in $k$ decreases as the maximum is approached and the added length of wire increases linearly (adding both cost and loss). Pads with strips spaced 15-40mm apart offer high coupling coefficients without requiring excessive wire. The open circuit voltage and short circuit current of the receiver are plotted in Figure 5-5(b). The short circuit current profile shows that the inductance of the pad initially increases at a greater rate than the mutual inductance between the transmitter and receiver but then limits. The $V_{oc}$ however increases almost linearly with the increase in the ferrite spacing (pad width) and this is due to increased mutual flux that is generated as the transmitter increases in size (the pads are driven with a constant current source thus there in an implicit increase in the energy available as discussed in Chapter 4.4.1).
Figure 5-5: Varying gap between ferrite strips (a) coupling coefficient and $P_{su}$, and (b) $V_{oc}$ and $I_{sc}$ against gap between strips ($I_t = 23A$ at 20kHz).

The ferrite structure enables arc shaped flux paths above the pads and a proportion of those couple to the receiver (mutual flux) as shown in Chapter 4. This flux is a result of current flowing in the section of coil that forms the flux pipe by covering the strip and a small portion of the coil extending past either side of the strip as shown by the cross-hatched part of the coil in the 115mm inset of Figure 5-5(a). This is due to the relatively low reluctance of ferrite. When the gap between the ferrite strips is large the distance between the extended part of the coil and the strip increases and flux flows around the coil rather than through the ferrite therefore no longer couples to the receiver. This adds to the leakage inductance and lowers
the coupling coefficient. The length of the return winding also increases further adding to the leakage inductance.

To illustrate the flux attracting effect of ferrite, both the coupling coefficient and $P_{su}$ per metre of pad width are plotted against the spacing between the ferrite strips in Figure 5-6. The greatest coupling per metre occurs with narrow pads because almost all the flux generated by the coils passes through the ferrite, however as shown by Figure 5-5(a), the overall coupling is relatively low because the pads are small resulting in poor power transfer. As the strips are spaced further apart, the $k/m$ reduces and the VA/m increases (due to greater pad size). Spaces of 15-40mm offer a practical compromise between ferrite usage and power coupled confirming the choice of a 33mm spacing for the built 4x6 pads.

![Figure 5-6: Coupling coefficient and $P_{su}$ per metre of pad width against ferrite spacing ($I_1 = 23A$ at 20kHz).](image)

As the space between strips increases, the ferrite needs to channel more flux due to flux contribution from the coil parts not directly above the strip as shown in the inset of Figure 5-5(a) potentially resulting in saturation. The flux density in the flux pipe should be as high as practically possible to ensure the highest ferrite utilisation, this is measured in VA coupled per cubic centimetre of ferrite with a given air gap.

The flux density is examined in the centre of the outer ferrite strip of the transmitter pad as shown in the inset of Figure 5-7. This location has been chosen because the constructive addition of the flux pipe and return windings results in the highest flux density found in Double D pads. Simulations show the flux density in the middle of the inner strips is 22% lower (under the same excitation). The material used is N87 and while flux densities above
200mT do increase core losses and the operating temperature, saturation only occurs if the peak flux density reaches 400mT (providing the operating temperature is below 100°C) [111].

The graph of Figure 5-7 shows the peak flux density as the space between the strips is increased. The relatively high flux densities are a result of the high number of turns in the flux pipe – each coil has 20 turns and these are added so that a current of 23A results in a total $NI$ of 920 amp turns in the centre. In practice the peak flux densities will be higher due to the operation of the chosen 2kW power supply described in [102, 113]. This means that the peak currents in the transmitter and receiver pads are twice as high as the average rms current.

The highest peak flux density in the DD transmitter is designed to be ~350mT in a relatively small volume of ferrite. Elsewhere the flux is significantly lower so the temperature throughout the pad only rises slightly above room temperature during extended operation. Slight local heating has little effect on pad operation and while additional ferrite could be added, it is unnecessary. In practice the volume of ferrite must be balanced against added weight, cost and performance.

![Figure 5-7: Peak flux density in the centre of the outer ferrite strip as the space between ferrite strips is varied from 5 to 115mm ($I_{\text{peak}} = 32.5A$).](image)

The 2kW power supply is intended to be used with a standard NZ household socket that is capable of supplying a maximum power of 2.4kW. If higher charge rates are needed, which is more likely with larger pads, additional higher current circuits can be added such as those used to power stoves or in extreme cases a three phase line may be used. Thus high power
systems would use a solid dc bus resulting in lower peak fluxes. Loss models for DD pads are considered in Chapter 6.6 where the hysteresis loss is determined by simulation and measurement.

5.3.3. Investigating width and length of pads independently

In the following sub sections the effect of the pad length and width (with reference to Figure 3-1) are investigated independently to determine the coupling limits of the Double D topology. This was done to ensure the models retain sufficient simulation accuracy, which depends on the size of the elements and surrounding air region. Ideally 3D FEM models in 32-bit JMAG should contain fewer than one million elements to ensure they fit within the computer memory and run in reasonable time. Making elements large affects accuracy thus pad models are restricted to a few meters in either length or width.

The simulations begin with pad width followed by length. Increasing the width of an initial base pad requires the insertion of equivalently spaced ferrite strips and in this analysis there is an implicit increase in inductance allowing higher power transfer because the transmitter pad is run with a constant current. The length of the flux pipe is independent of width of the pad and thus it remains constant.

Increasing the length of the pad requires an integer number of ferrite “I” cores to be added to the ferrite strips within the pad. However, doing so means the coverage of the midsection and therefore the length of the flux pipe changes. Measurements vary with changes in both variables complicating analysis. In order to investigate length only, the coverage needs to be constant as cores are added to lengthen the strips and there are two approaches for achieving this: the first is to fix the number of turns and assume the copper can be spread (by changing the cross-section of the coil or increasing the winding pitch) and the second method assumes more turns can be added and the coil is wound with a constant pitch. The results of both approaches are presented in section 5.3.3.2 and 5.3.3.3 respectively.

For the following sections the length and width refer to the maximum dimensions of the coils as shown in Figure 3-1. In practice there will be a plastic case with or without an aluminium shielding ring depending on the application. The size of the plastic casing around the pad depends on the construction technique and therefore it is not considered.
5.3.3.1. Varying the width of pads

The effect of pad width (dimension shown in Figure 3-1) on coupling has been investigated by adding ferrite strips while maintaining a consistent spacing of 33mm between them. A few sample models with various length and width configurations labelled are illustrated in Figure 5-8; note both the transmitter and receiver pads are identical for all simulations. The base pad uses four strips made up of six “I” cores giving a total ferrite length of 558m and the coils were made up of 20 turns of 4mm diameter wire. To enable a fair comparison between pads of various widths and lengths, results are plotted against actual pad dimensions. The graph of Figure 5-9(a) shows $V_{oc}$ and $I_{sc}$, while $k$ and $P_{su}$ are shown in Figure 5-9(b). In both graphs results are plotted with air gaps of 200, 250 and 300mm.

Once the pad is sufficiently wide (greater than 0.75m), the width determines $V_{oc}$ while $I_{sc}$ stays constant. This is because the rate of increase in mutual inductance closely matches that of the increase in self inductance. This is reflected by the linear increase in $P_{su}$ as well as the asymptotic limit in $k$. Notably, the rate of change of coupling is greatest for pads less than 0.75m wide and greater widths require a linear increase in material with little gain in the coupling coefficients such relatively narrow pads are preferred.

Figure 5-8: Sample of pad models described by rows and columns of ferrite, and the number of turns.
Chapter 5 – Optimisation of Double D power pads

5.3.3.2. Varying pad length with a fixed number of turns

Pads with 20 turns have been used and the length the ferrite strips was varied from 0.37m to 2.33m by increasing the number of I93 cores from 4 to 25. The pads are slightly longer due to the fixed width of the return portion of the coils (illustrated by the cross-hatching in Figure 3-1) therefore the simulated range of pad length is 0.552m to 2.505m; clearly extremely large pads are impractical however the simulation was done to observe the performance limits of the Double D topology. The results are shown in Figure 5-10(a) and (b), markers are included on both graphs to show where the 4x6 and 4x12 (with N=20) pad models exist.

Figure 5-9: (a) Open circuit voltage and short circuit current against pad width (b) coupling coefficient an uncompensated power against pad width ($I_1 = 23A$ at 20kHz).
Chapter 5 – Optimisation of Double D power pads

The number of turns has been fixed so there is no increase in mmf generated by the transmitter as the pad is made longer causing \( V_{oc} \) and \( I_{sc} \) to reach asymptotic limits. The limit in \( I_{sc} \) occurs when the pads are greater than 1.75m long. The rate of change in the open circuit voltage is continuously decreasing however the asymptote only becomes clear with pads that are significantly longer than 2.5m long. The coupling coefficient appears to reach asymptotic values of 0.42, 0.34 and 0.29 when the air gaps are 200, 250 and 300mm respectively as shown in Figure 5-10(b). The curve for \( k \) with a 200mm air gap is the flattest of the three between 2.2 and 2.5m, the remaining two are still increasing slightly.

![Diagram](image)

**Figure 5-10:** Investigating pad length with a fixed number of turns (a) open circuit voltage and short circuit current, and (b) coupling coefficient and uncompensated power as the length of the pads is increased \( (I_t = 23A \text{ at } 20kHz) \).
The plateau in the coupling coefficient that occurs with long pads as shown in Figure 5-10(b) is due to flux path lengths and these can be illustrated via simulation. The pads are referenced to the axes shown in Figure 5-11(a). The front of the transmitter pad is in the $xy$ plane and intercepts the $z$-axis at $z=0$. The $xz$ plane slices through the geometrical centres of the pads. The magnetic flux density vectors in the $xz$ planes for pairs of 4x6 and 4x12 pads are shown in Figure 5-11(b) and (c) respectively. The flux density scale has been reduced to 10mT to better show the flux in the 250mm air gap.

The flux density above the flux pipe is approximately 3mT for the 4x6 pad and this relatively high flux density increases the magnetic pressure supporting the mutual flux ($\Phi_M$) thereby enhancing coupling to the receiver. The path length of the leakage flux is illustrated by $l$ and its height is shown by $h$. For the case with 4x6 pads, $l < 2\delta_z$, therefore the leakage flux desirably travels between pole faces so it is able to support the mutual flux. There is a small void in the flux density above the $h_6$ marker and below the centre of the receiver due to the ferrite attracting flux.

A cross-section showing magnetic flux density vectors in an $xz$ plane through the centre of 4x12 pads is plotted in Figure 5-11(c). The 1.3m long pads are at ~80% of their asymptotic coupling limit as shown in the coupling curve in Figure 5-10(b) for a 250mm air gap. The reason for the limit in $k$ is that the leakage path is greater than the distance to the receiver ($l > 2\delta_z$) and in consequence, flux goes to the receiver rather than through the air gap to the opposite pole face. As a result, the flux density above the flux pipe coils to be much lower than that of the 4x6 pad and this is further illustrated by the larger void. While the coupling coefficient is higher than with the 4x6 pads, the coupling mechanism is shown to be non ideal in section 5.3.3.4. As shown in Figure 5-10(b), the asymptotic limit is 0.42 with a 200mm air gap for pads that have 4 strips of ferrite. The size of the void will increase as the pads become very long because the path via the receiver ferrite has a much lower reluctance than the path to the opposite pole face.

Mutual flux between the transmitter and receiver is enhanced by a secondary effect that is a result of the end return portions of the coil (these are shown in Figure 3-1). The flux around the return winding labelled in Figure 5-11(b) increases the magnetic potential at the end of the pad and this forces the mutual flux upward toward the receiver. If the end winding did not exist, the mutual flux would bulge out of the air gap before returning to the opposite pole face reducing coupling.
5.3.3.3. Varying pad length while maintaining constant coil pitch

The second approach to determining the effect of pad length on coupling requires constant coil pitch and thus the number of turns must increase as the pad becomes longer. The coverage and pitch were fixed to 6.1mm and 43% respectively. As the pad length is scaled linearly, the number of turns per coil is calculated from: \[ N = 3.2759c + 0.3448 \], where \( c \) is the number of I93 cores in the strip. The nomenclature is chosen so the variable ‘c’ refers to the number of ferrite columns forming the strip whereas the number of strips is the number of rows (r). The base pad uses 6 x I93 cores thus the rounded number of turns per coil is 20. The number of cores was varied from 1 to 25 resulting in pad lengths ranging from 0.14m to 3.0m (the pads are larger than in section 5.3.3.2 due to the number of turns increasing). The open circuit voltage and short circuit current are plotted in Figure 5-12(a) while the coupling coefficient and uncompensated power in Figure 5-12(b).
The coupling coefficient is 0.008 with a 200mm air gap with 4x1 pads and there is a rapid increase in $k$ until the point of inflection at a pad length of 0.62m. The rapid increase is due to the combined effect of the number of turns and pad length increasing. As more turns are added the leakage flux density above the flux pipe increases pushing flux toward the receiver (it becomes mutual flux) as shown in Figure 5-11(b).

![Figure 5-12](image)

Figure 5-12: Varying pad length with constant pitch coils (number of turns increases with length) both (a) open circuit voltage and short circuit current and (b) coupling coefficient and uncompensated power against pad length ($I_1 = 23$A at $20$kHz).

Although the mmf increases with length due to more amp turns, there will still be an asymptotic limit in $k$ because the distance to the receiver will be less than half of the leakage path length. The short circuit current also reaches an asymptotic limit because the inductance of the pad increases with length however the uncompensated power will continue to increase
with the open circuit voltage. This is because there is little change in the path lengths of the magnetic circuit with extremely long pads but there is a continual increase in the mmf as the pads are driven with constant current.

These pad length simulations assume the ferrite does not saturate as the pads are made longer however in practice saturation is likely (assuming the transmitter current is maintained at 23A/turn). Simulated pads up to a length of 1m will have peak flux densities less than 400mT, if longer pads are needed ferrite cores may be inserted in between the middle and outer rows. The exact number of cores required will need to be determined by further simulation and is beyond the scope of this thesis.

5.3.3.4. Comparing changes in pad length and width

There are three broad operational scenarios for power pads determined by the flux coupling (or non-coupling) mechanisms:

- There is little coupling between pads because the height of the flux path is too low, this occurs with short pads (<0.4m) shown in Figure 5-12(b).
- The leakage flux above a flux pipe increases the magnetic pressure forcing flux upward to the receiver (designs that do this are considered optimal).
- The path length to the opposite face is too long therefore the flux goes to the receiver and there is little flux above the flux pipe (approaching asymptotic $k$ limit).

The latter two scenarios are shown for 4x6 and 4x12 pads in Figure 5-11(b) and (c) respectively. Clearly an optimal solution exists where the pad length is increased to the point where coupling is largely due to magnetic pressure as a result of the flux pipe. Consequently, the length of the pad needs to be matched to the operational air gap and this can be determined by dividing the coupling coefficient by the pad length. The coupling coefficient per meter of either pad width or length is plotted against pad length in Figure 5-13(a), (b) and (c) for 200, 250 and 300mm air gaps respectively. The three curves on each graph are for pads with a: fixed number of turns, variable number of turns or pad width (fixed number of turns).
Figure 5-13: Coupling coefficient divided by either the length or width of the pads with peaks in the profiles labelled with air gaps of (a) 200mm, (b) 250mm and (c) 300mm \( (I_t = 23A \text{ at } 20kHz)\).

The peaks in the \( k/\text{width} \) are illustrated in Figure 5-13 by three clustered dashed vertical lines (labelled as \( w \)) traversing all three graphs and the peaks are largely independent of vertical offset – this is expected since the height of the flux paths is mainly due to pad length. The two vertical lines labelled \( 'l_{\text{fix}}' \) and \( 'l_{\text{var}}' \) on each graph indicate the length of the pad at which
the maximum coupling per meter occurs for pads that have a fixed and variable number of turns respectively. The length of the particular pad is also labelled and this is the optimal operating point because the coupling is due the fundamental flux path height, which is determined by the length of the flux pipe. As the air gap increases the peaks in $k/m$ occur with longer pads as the pair of lines move toward the right going from Figure 5-13(a) to (b) and finally to (c) further illustrating that longer pads permit higher fundamental flux paths allowing good coupling with larger air gaps.

The overall optimisation aim is to ensure the pads offer the greatest coupling coefficient for a given length with a given air gap thus designs that have dimensions corresponding to the greatest change in coupling per metre are preferable. These offer the highest performance for the amount of material used – pads that are between 0.7 and 1.1m long are considered optimal for designs that use 4 strips of ferrite. However, if the coupling is still not sufficient to achieve the desired power transfer, the width and length need to be increased as discussed in the next section.

5.3.4. Varying the area of Double D pads by changing the length and width

In the previous section the length and width of Double D pads were investigated separately to determine the coupling limits of the topology. The length of the flux pipe is determined by the coverage of the midsection as discussed in section 5.3.1 and this must increase with the length of the pads. Two approaches were used in section 5.3.3 to ensure 43% coverage, these were: assuming that the copper could be spread, and winding the coil with the same pitch and increasing the number of turns. The latter approach becomes impractical with especially large pads as they have large inductances requiring high driving voltages to obtain the desired 23A transmitter pad current. Winding approaches with parallel windings such as bifilar or quadrafilar coils can be used to lower inductance however the input current to the pad needs to increase to maintain the same NI, for example, a bifilar pad will achieve a terminal voltage reduction of 50%. Thus parallel wound coils are generally only practical if the greatest pad dimension is less than 1m.

A pad size of 1m is large in the context of EVs however designs for bus charging require significantly higher power transfer (>150kW) with large air gaps (~300mm) necessitating big pads. In order to ensure the large designs explored in this section can be built practically, the number of turns is fixed and it is assumed the copper can be spread. Forming a flux pipe with a pitch that is a few times the wire diameter will work in practice as simulated [114]. As in
the previous simulations, the coils in both the transmitter and receiver are made up of 20 turns of 4mm diameter wire. Pads up to 1.7m long can be simulated with confidence – with 43% coverage of the midsection, these pads have a flux pipe length of 654mm, which is made up of wire wound with a pitch of ~16mm. Note, the return portion is not considered when determining coverage as shown in Figure 3-1.

5.3.4.1. Simulated pad sizes

As in the preceding sections, both the transmitter and receiver pads are made identical to minimise the number of simulations, however the air gap range has been changed and pads are evaluated at separations of 100, 200 and 300mm. A 100mm air gap is included for applications where the either transmitter pad is placed on the ground (rather than being embedded in the ground) or the EV is small and has a low ground clearance. As in previous designs, the ferrite strips within the pads are made up of I93 cores and pads ranging from 4x4 to 19x12 were simulated. This corresponds to a smallest pad size of 0.391mW x 0.552mL and a largest of 1.306mW x 1.296mL. The number of columns was restricted to 12 because the cores are rectangular so the pad length increases at a far greater rate than the width (when an integer numbers of cores is used). The I93 cores are 28mm wide and with 33mm between strips, each added strip increases the width of the pad by 61mm while each core adds 93mm to the length.

In order to make fair comparisons between the sizes of pads, the actual length (plotted on the x-axis) and width (plotted on the y-axis) of pads are graphed. However, the pads are referred to by their number of rows and columns in a similar manner to matrices. The pad model is shown in Figure 5-14(a) with all the fixed dimensions such as the spacing between the strips, the distance from the strips to the coils and the coil widths labelled.

A chart with the dimensions showing all the row and column combinations of the simulated pads is shown in Figure 5-14(b). The upper half of the chart contains either the number of columns or rows of a pad while the lower half only contains rows. As mentioned earlier, the width of the pads increases less rapidly than the length with integer increments in the number of cores in the respective dimension, therefore to ensure square pads measuring ~1.3mW by ~1.3mL were considered, pads with 19 rows were simulated. The chart of Figure 5-14(b) is used to determine pad dimensions as follows, assuming an 8x5 pad, first locate ‘8’ and note the width of 0.635m as this is determined by the number of rows, then locate ‘5’ and note
0.645 m as the length, which is determined by the number of columns. The lower chart needs to be used to determine the width of pads with more than 12 rows.

The coupling coefficient at an air gap of 100 mm is plotted against pad length and width in Figure 5-15. The results are consistent with those in section 5.3.3 however the asymptotic limits in $k$ are not clear as the maximum pad dimension given here is only 1.3 m in both axes. The effect of increasing either the length or width of the pads is however easily seen. Increases in $k$ are much greater when the length is increased compared to equivalent increases in the width as the pads are grown starting from the lowest point on the surface close to the origin. This effect this also occurs at the opposite edge of the surface where extremely wide pads are made longer.

The lighter shades of grey in Figure 5-15 show higher coupling coefficients and indicate that length is the preferable variable to increase. The lowest coupling coefficient is ~0.4 and this is considered very high – pads with greater values are not practically usable as they are very sensitive to any horizontal or vertical misalignment as discussed in Chapter 3.4.3 in the context of circular pads.
Chapter 5 – Optimisation of Double D power pads

The high coupling coefficients of Figure 5-15 indicate designs smaller than 4x4 (0.391mW x 0.552mL) are recommended for 100mm air gaps but these have not been simulated due to space constraints. Rather, an emphasis is made on pad designs suitable for 200-300mm air gaps as these are considered more practical for EVs and buses that have been designed for use with transmitter pads embedded in the ground.

![Graph showing coupling coefficients](image)

Figure 5-15: Coupling coefficient against length and width for various pad models with a 100mm air gap ($I_1 = 23A$ at 20kHz).

5.3.4.2. 2D isoline plots of pads up to 1.3m square

The 3D plot of Figure 5-15 makes selecting pad designs difficult as only general trends are visible, therefore from this point onward only 2D contour plots with isolines are used. The isolines show pads that have the same graphed parameter, making changes between different designs more visible. The simulated pads are investigated using integer step increases in the number of rows and columns however plotting values such as $P_{su}$ or $k$ with isolines makes the pad size appear continuous. Therefore, to ensure feasible designs are chosen, markers (+) showing each of the simulated pads are included and the closest marker to the chosen pad parameter ensures a close to optimal design is selected. Both black and white markers have been used to enhance visibility.
The set of 2D plots in Figure 5-16(a)-(c) show the coupling coefficient for perfectly aligned pad pairs that have 100, 200 and 300mm air gaps respectively. The y-axis for all graphs is the pad width, noting the label has been removed on the right hand graphs to increase their size. The bottom left hand corner of all the graphs represents a 4x4 and the number of rows increases going up the y-axis and the number of columns increase going along the x-axis, these labels have been added to Figure 5-16(a). All 2D isoline graphs follow this definition.

Figure 5-16: Coupling coefficient against pad size at with (a) 100, (b) 200 and (c) 300mm air gaps. (d) $P_{su}$ against pad size with a 300mm air gap (for all simulations $I_1 = 23A$ at 20kHz)
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The 4x6 pads are easily found by counting the markers on the graph of Figure 5-16(b) and noting the isolines change in 0.02 increments. The 4x6 pads are shown to have a coupling factor of ~0.24 – this is consistent with the measured result in section 5.2 confirming the accuracy of the model. The plot in Figure 5-16(c) shows the uncompensated power for the simulated pads with a 300mm air gap. Pads approximately 1m$^2$ in area have a coupling factors of around 0.72, 0.48 and 0.30 at 100, 200 and 300mm separations respectively and are easily able to couple 10kVA over a 300mm air gap. The $P_{su}$ (at an air gap of 300mm) can be doubled to 20kVA with a 32% increase in area allowing adequate power for bus charging when resonated.

Coupling factors less than 0.4 are preferred for EV charging with a 200mm air gap and while greater values are possible with larger pads, they are not necessary as power transfer is sufficient. In addition, smaller pads are more cost effective as less copper and ferrite is needed. Pads less than 0.9m x 0.9m are therefore suitable for EV charging as shown in Figure 5-16(b). To show the design options more clearly, $k$, $P_{su}$, $V_{oc}$ and $I_{sc}$ are shown for a limited range of lengths and widths at an air gap of 200mm in Figure 5-17(a)-(d) respectively. In the simulation the two ‘D’ coils that make up a DD pad are connected in series resulting in very high voltages with large pads. The coils are referred to as coil ‘a’ and coil ‘b’ in Figure 4-13 in Chapter 4.3.7. In practice the individual coils are bifilar wound connected in series and this means measured voltages will be halved and the currents will be doubled.

The $P_{su}$ and $k$ isolines for all air gaps are initially very steep indicating that large changes in width are required to maintain the same coupling coefficient for short pads. This further illustrate the importance of flux paths heights – short pads have low flux path heights and therefore need to rely on the sum of ‘many poor low coupling flux paths’, which are formed by a wide pad having many ferrite strips. Conversely, the flux path height is directly related to length therefore narrow but long pads can achieve the same coupling coefficient. The optimisation of length and width is done in section 5.3.4.3.
Variations in the open circuit voltage and short circuit current against the full pad size range with a 300mm air gap are shown in the plots of Figure 5-18(a) and (b) respectively, the corresponding coupling coefficient graph is shown in Figure 5-16(c).
5.3.4.3. Selecting optimum pad sizes

Assuming there is some degree of tolerance allowed in the desired $k$ or $P_{su}$ values at a given air gap, then it is clear from Figure 5-16(a)-(d) that several pads have similar performance. Using Figure 5-16(c) as an example, if a $k$ of approximately 0.2 is desired with pads vertically aligned at an air gap of 300mm, the following pads will suffice: 17x6, 8x7, 6x8, 5x9 and 4x10. All require various amounts of material however the most optimal option is to use the pad that has the least area as this ensures the material used is minimised. From the chart in Figure 5-14(b) it can be seen that the pad that has least area is the 4x10 (0.43m²) therefore it is the most preferable design.

In order to determine optimum designs for all the simulated cases, the coupling coefficient per m² of pad area is plotted. The results for a 200mm air gap are presented in Figure 5-19(a) and for a 300mm air gap in (b). The method to ensure optimal designs are selected for operation with either 200mm or 300mm air gaps is to note all possible pads that can closely provide the desired coupling factor from Figure 5-16(b) or (d). After transcribing the possible designs to Figure 5-19(a) or (b), the pad size that sits the closest to the highest value isoline should be chosen to ensure the highest $k/m^2$. This has been done for the pads listed above that were chosen to enable a coupling factor of 0.2 with a 300mm air gap where the most optimal design is indicated by a black dot in Figure 5-19(b).
The optimal $k/m^2$ occurs with 4x7 pads with a 200mm air gap and with 4x10 pads with a 300mm air gap as shown in Figure 5-19. This indicates that pad designs are dependent on the air gap and need to be chosen appropriately. Pads that have 4 rows and 6-9 columns have the highest coupling coefficient to area ratio with an air gap of 200mm as shown in Figure 5-19(a). The isolines spread further apart moving away from the peak in $k/m^2$ indicating the coupling coefficient increases at a lower rate than the area. This is consistent with the plots of Figure 5-13 that show $k/m$ of pad width.

High coupling coefficients with narrow pads (but of sufficient length) are due to the portions of the return windings that are perpendicular to the flux pipe as labelled in Figure 5-20(a). These coil sections are referred to as the width windings and they remain at a fixed length when the pads are made wider unlike the flux pipe coils and the end return winding. The flux pattern produced by the width windings adds to mutual flux indirectly. The pattern is illustrated with magnetic flux density vectors in the $yz$ plane positioned in the middle of a pole face as shown in Figure 5-20(a).

The vectors on the $yz$ plane of only a transmitter pad energised with 23A are shown in Figure 5-20(b), note the scale has been adjusted to a maximum of 10mT to better show the flux about the pad. The direction of the flux vectors in the ferrite strips as shown here is out of the page while the current in the left and right return coils are out of and into the page respectively. The return windings increase the flux density at the edges of the pad and this
magnetic potential prevents mutual flux spilling out of the sides of the air gap between the transmitter and receiver. The effect is similar to that described in section 5.3.3.2 where the end winding helps to prevent flux bulging out the ends of the air gap. The flux path around the width winding is also explicitly labelled in Figure 5-20(b) and although a significant proportion of it does not couple to the receiver, the increased magnetic potential is beneficial.

![Diagram](image)

**Figure 5-20:** (a) YZ plane half way through pole face of a Double D transmitter pad and (b) magnetic flux density on the yz plane of (a) \(I_1 = 23\,\text{A}\).

The flux containing effect is more pronounced with relatively narrow pads because the flux path and hence magnetic potential around it remains the same regardless of width. This causes \(k/m^2\) to reduce as the pads are made wider as described earlier.

The preceding sections show that in order for Double D pads to be considered fully optimised they need to make good use of the following coupling enhancing techniques:
• The magnetic flux above the flux pipe should force the mutual flux up to a receiver (not rely on mutual flux to travel to the receiver because the distance between pole faces is too far leaving the receiver as the lowest reluctance path).

• The flux from the end winding should prevent mutual flux spilling out of the ends of the air gap.

• The width return windings should be used to form a high potential boundary around the edges of the pad.

• Pads that have high $k/m^2$ values adhere to the above points.

5.3.4.4. Cost model of Double D pads

Thus far the emphasis has been on maximising the coupling coefficient between Double D pads however cost is also a critical factor in determining the feasibility of the complete EV charging system. The two major cost components of the pads are ferrite and Litz wire and in this work standard I93 cores and 4mm diameter wire are used. The prices used to quantify the overall cost here are those paid for cores in quantities of 1000 units and Litz wire in 1000m lengths, although these costs will be far lower in a mass production scenario. Three sets of prices are used to try and account for this – the first is $4/core and $2/m and the second is $2/core and $4/m of wire and the third is $2/core and $2/m. A ratio of 2:1 has been used to determine the effect of one of the items being more expensive than the other – the only difference is the absolute cost if the cost of both items is increased proportionally.

The cost efficiency of a pad is determined by the coupling coefficient divided by the cost ($/k$) but given the cost of pads is in the range of $100s and $k$ is <0.4, the actual numbers of $k$/ are very small. Therefore, the coupling to cost ratio is multiplied by 1000 and is written as 1000$k$/$. The coupling coefficient for a set of identical pads with an air gap of 200mm is plotted in Figure 5-21(a) and the cost to coupling results are shown in Figure 5-21(b)-(d) for the cost scenarios. Optimal pads have the highest $k$/ ratio and these are on the x-axis, which is consistent with the $k/m^2$ of pad area plots of Figure 5-19(a).

Ignoring pads with both components priced at $2, pads that have ferrite at $4/core and Litz wire at $2/m are more cost effective. The cost of copper dominates the price of pads. The actual cost of a set of pads can be calculated with reference to the coupling plot Figure 5-21(a). For example, from the 1000$k$/ plot of Figure 5-21(a) and the coupling coefficient
plot of Figure 5-21(b), the cost of a set of 4x6 pads is $369 (1000x0.24/0.65) and from Figure 5-21(c) the cost is $457 (1000x.24/0.525).

![Figure 5-21](image)

Figure 5-21: (a) Coupling coefficient. Cost-performance of DD pads against size (b) 1000k$/s with ferrite at $4/core and Litz wire at $2/m, (c) 1000k$/s with ferrite at $2/core and Litz wire at $4/m, and (d) 1000k$/s with ferrite at $2/core and Litz wire at $2/m ($I_1 = 23A at 20kHz and $\delta_z = 200mm$).

5.3.4.5. Built 4x6 DD pads

The dimensions of the built pads for experimental validation are partly based on dimensions derived from the double sided flux coupler called a shaped bar flux pipe pad (presented in Chapter 4.3.1) and the optimisation processes described in the preceding sections. A coupling coefficient of greater that 0.2 to is required with a 200mm air gap to enable 2kW of power.
transfer. The chosen length of the pad to the edges of the coil, as shown in Figure 5-14(a), was 738mm, and the width was 391mm (four ferrite strips 28mm wide with 33mm spaces between them). The coils consist of 20 turns of 4mm diameter wire as they have a preferable lower self inductance than pads with 22 turns and can easily transfer the required power. The flux pipe was made to be 240mm long corresponding to 43% of the ferrite strip being covered.

The pads do not need to be shock and weatherproof for lab testing therefore a high density closed cell PVC foam block was used for the pad shell that holds the ferrite strips and coils. The was shell shaped by CNC milling, however the material properties only allow a smallest feature size of 2mm therefore a coil pitch of 6.1mm was chosen leaving 2.1mm wide lands between turns. The front cover was made out of 5mm thick PVC while the backing plate was made out of a 5mm thick aluminium sheet.

Figure 5-22: Dimensions of the built pad for testing and simulation verification.

The leakage inductance of the pad can be reduced without adversely affecting the mutual inductance by spreading the return windings; these are the portions of coil not forming the flux pipe as shown in Figure 5-20(a). However, this was not done as the power transfer is sufficient and smaller pads are desired. Similarly, coupling is improved if the distances between the coil and ferrite, as shown in Figure 5-2, are made larger however these have been set to 10mm also to minimise the pad size.

The position of the shield has little effect on coupling or loss and in the prototype was not used. The total dimensions of the pad shell include the space around the perimeter for plastic screws and are shown in Figure 5-22 above.
5.4. Summary and Conclusion

In this chapter the Double D topology, which was introduced in Chapter 4, was optimised however several constraints were applied to keep the simulation requirement reasonable. The pads were built with readily available I93 cores and 4mm diameter Litz wire. Pad models consisted of identical pads and changes made were to: the number of turns in the coils, the length of the pad and the width. Changing the length requires adding integer numbers of I93 cores to the ferrite strips while changing the width requires more strips comprising I93 cores to be added. Only integer increments were used to minimise difficulties with cutting ferrite and to limit the scope.

Prior to running the numerous simulations required for optimisation, a measurement versus simulation experiment was done to validate the simulation model. The simulated results differed from the measured by only a few percent allowing pad optimisation to be done via simulation with confidence. The built pad was used as a base and parameters were changed to determine which variable should be changed to increase performance. Firstly the number of turns and the pitch of the coils in the flux pipe were changed. The number of turns governs the \( N_I \) and this determines the output power of the receiver pad. The simulated range was from 12 to 26 turns with 20% to 60% coverage. Coils with 20 turns were chosen with a coverage of 43% as this offered a good compromise between the coupled power (2kW) and the coupling coefficient (in this work the transmitter pads are excited with a constant 23A current).

The coupling increases as the spacing between the ferrite strips that make up the length of the pad is increased but the flux density also increases. High coupling is desired but as the strips are moved further apart the ferrite needs to guide more flux which will eventually result in saturation. The coupling coefficient per metre of pad width was used as a metric and a spacing of 33mm was chosen as it offered good use of the ferrite. The peak flux density was ~300mT so the saturation is not an issue. While core loss does increase with flux density, the volume of ferrite where the peak value occurs is very small resulting in low overall loss.

Increasing the width of a pad simply involves adding more ferrite strips however increasing the length was done in two parts. The absolute length of the flux pipe determines the fundamental flux path height above the pad and this needs to vary in proportion with the length. There are two ways to achieve this, the first is spreading the copper, which means winding the coils with a larger pitch. The second method requires the coils to be wound with
the same pitch but the number of turns is increased. The latter approach is suited to small pads for operation with low air gaps due to the reduced number of turns. This becomes impractical with large pads because the excessive inductance means very large terminal voltages are required to get the desired 23A transmitter pad current. Optimal pads sizes were found by investigating the ratio of $k$ to pad length, which is dependent on the operational air gap – larger pads are more optimal with larger air gaps.

The effect of pad area on coupling was determined by adding more strips and or cores to increase the size of the pad. The results were presented on 2D plots with isolines indicating pads of equivalent parameters. Optimal pads offer the highest $k/m^2$ of pad area and these are also dependent on the air gap. The plots indicate that if more power is required then the pad length should be increased rather than the width (this should only be increased if the maximum length is reached and still more power is needed). Narrow pads offer optimal coupling due to secondary effects caused by the return windings. These increase the magnetic potential around the edge of the pad forcing the mutual flux inward and therefore upward to the receiver resulting in high coupling coefficients.
Chapter 6. Improvements to Double D pads

6.1. Introduction

In the previous chapter the size of Double D pads was matched to an operational air gap and an optimal aspect ratio was determined. In this chapter the effect of horizontal misalignment is considered and an additional coil is placed within a receiver to improve performance. Loss mechanisms are investigated followed by magnetic field leakage to determine compliance with international emission guidelines. Finally an interoperability study is done to determine how the DD topology works with well established circular pads. Here charge zones are used to show tolerance to horizontal misalignment in a practical operational scenario [115, 116].

The Double D pads used in this chapter were selected based on optimisation results in Chapter 5 and they have been chosen for two systems: a 2kW one that operates with a ~200mm air gap and a 7kW one that operates with a 125mm air gap. The latter system is favourable for scenarios where the transmitter is placed on the ground rather than being buried and it permits faster charging. Horizontal profiles are initially taken with identical pads to gauge their sensitivity to misalignment; the pads are referred to as 4x6s and they measure 0.74mL x 0.41mW and use 10 turn bifilar coils that cover 43% of the midsection. The polarised nature of the DD topology results in relatively poor tolerance to misalignment in the x-axis and therefore a magnetically decoupled coil, which is referred to as a spatial quadrature coil, is added. The added coil requires lengthened ferrite strips for optimal operation therefore the receiver becomes a 4x8.

Power loss in a DD transmitter pad is due to current in the winding, core loss in the ferrite and eddy currents in the aluminium base plate. The loss in the winding can be easily measured with an LCR meter however core losses are difficult to isolate. An approach using a simplified Steinmetz equation with coefficients determined from loss data provided by the ferrite manufacturer is applied to a simulation model. The approach is feasible given measured inductance values are within a few percent of those simulated and the ferrite is responsible for a large proportion of the inductance (by guiding flux) and therefore the flux within it is simulated reliably.

The leakage magnetic field produced by a DD system is compared to that of a circular pad system. The vastly differing coupling coefficients make direct comparisons difficult therefore
an approach where the receiver’s operational $Q$s are equal is used. In order to do this the current in the circular transmitter needs to be increased significantly. This is possible given the flux density in the ferrite of a circular transmitter pad is not very high to begin with.

Finally an investigation into interoperability is undertaken where the DD pad with the additional quadrature coil is tested above a conventional transmitter circular pad. This is essential because circular pads are well established in academia and industry and there is a high chance they may be deployed.

A highly desirable feature of the new Double D topology is that it is polarised. This allows it to be used for both stationary and dynamic charging, the objectives of which are outlined in the following chapter.

### 6.2. $P_{su}$ profiles of Double D pads

Horizontal $P_{su}$ and coupling profiles are used to determine the sensitivity and tolerance limits of power pads. These are created by measuring the open circuit voltage and short circuit current at a receiver pad as it is moved horizontally above a transmitter pad operating with rated current. The air gap is fixed at the expected operational value. Double D pads are polarised therefore a full horizontal profile requires measurements in one quadrant and this is cumbersome in the development stage. Initial measurements showed that DD pads can be accurately characterised with only two profiles in a similar manner to the oval pads of Chapter 3 (profiles in the $x$-axis (length) and the $y$-axis (width)). Offsets that have both $x$ and $y$ components lie within the bounding $x$ and $y$ profiles.

Measured horizontal profiles were taken using the 4x6 transmitter and receiver pads developed in Chapter 5. The profiles of the uncompensated power and coupling coefficient are presented in Figure 6-1(a) with an air gap of 200mm. The open circuit voltage and short circuit current are shown in Figure 6-1(b). The pads have a coupling coefficient of 0.23 and this enables the desired 2kW to be transferred easily when the pads are vertically aligned. Assuming a maximum operational $Q$ of 6 is allowed, the maximum horizontal offset at which full power can be delivered in the $x$-axis is only 160mm but full power delivery is possible with offsets of at least 200mm in the $y$-axis.

This polarised nature makes orienting the pads difficult for practical EV charging systems as drivers naturally have better control in one direction. Controlling the position in the
Chapter 6 – Improvements to Double D pads

forward/backward direction is easier when parking in lots or garages indicating the pad should be installed with the \( x \)-axis along the length of the car. However, as shown in Chapter 7, doing so would not permit extension to Roadway Powered Electric Vehicles (RPEV).

![Figure 6-1: Horizontal profiles (a) \( P_{\text{su}} \) and \( k \), and (b) \( V_{\text{oc}} \) and \( I_{\text{sc}} \) (\( \delta_z = 200 \text{mm} \) and \( I_1 = 23 \text{A} \) at 20kHz).](image)

6.2.1. Null in \( x \)-axis power profile

The sharp drop in coupling in the \( x \)-axis shown in Figure 6-1(b) is due to a null that exists at an offset of 260mm as illustrated in the \( V_{\text{oc}} \) profile of Figure 6-2 (note the receiver DD coils are connected in series resulting in higher voltages). At the null the mutual coupling is reduced to zero because the flux is entering and exiting only one of the receiver coils. Nulls determine the gradient of the power profiles therefore impose similar performance restrictions on the DD topology as they do on the circular pads described in Chapter 3, section 3.3.

![Figure 6-2: Horizontal \( V_{\text{oc}} \) profile of a DD receiver in the \( x \)-axis (\( \delta_z = 200 \text{mm} \) and \( I_1 = 23 \text{A} \) at 20kHz).](image)

All power profiles must pass through the null regardless of the \( P_{\text{su}} \) when the pads are centred making systems operating with relatively small air gaps very sensitive to any horizontal
Chapter 6 – Improvements to Double D pads

misalignment. As shown in Chapter 5.3, $P_{su}$ increases at a greater rate than the length of the pad at a given air gap (when pads are less than 1m long) therefore making pads longer to shift the null is impractical because systems become more sensitive to horizontal misalignment. The null is intrinsic to the DD coil structure and a practical approach to improving performance in the x-axis is to add another coil that couples flux when the DD coils do not.

6.3. Adding a quadrature coil to a Double D receiver

An isolated DD transmitter produces arc shaped flux paths that initially start vertically (at the pole face) and then become horizontal (above the flux pipe) and finish vertically (in the opposite direction at the returning pole face). DD receiver coils couple mainly horizontal flux in the same manner as the transmitter generates it but the directional change in the flux path means that at some position there will be no coupling. For a pair of 4x6 DD pads satisfactory operation is possible up to an offset of 160mm in the x-axis because the low reluctance of the ferrite in the receiver directs flux through the coils. The flux is largely horizontal for offsets in the y-axis but slowly decreases in magnitude as the receiver is moved away, there is no null or forced flux cancellation enabling full power delivery with offsets of at least 200mm.

6.3.1. Operation of the quadrature coil

A quadrature coil can be added within the winding area of the DD coils to keeping the external dimensions of the pad the same. A Double D pad with an added quadrature coil is referred to as a DDQ and one is shown in a conceptual diagram of a receiver pad with three strips of ferrite in Figure 6-3(a). The DDQ receiver pad is shown upside down to better illustrate the quadrature coil – during operation the coupling side faces the transmitter (clarified in the conceptual IPT system of Figure 2-4 in Chapter 2 where the receiver is mounted upside down under the EV). In this case the added coil is considered to be spatially in quadrature because it is only active with horizontal offsets that have a component in the x-axis.

The position of the quadrature coil ensures it is decoupled and this can be shown by assuming the DD coils in a receiver are excited by a transmitter resulting in current $I_2$. The physical arrangement means there is no net flux through the quadrature coil due to the DD receiver coils as shown by the winding diagram with simplified coils in Figure 6-3(b) where the dots and crosses represent magnetic flux out of and in to the page respectively. Sections of the quadrature coil are closely coupled with the return windings of both the DD coils but due to
the change in the current directions with the coil structure, the net flux through the quadrature coil is cancelled.

There are effectively two coupling flux cancellation cases. The first is for a section of the quadrature coil that is coupled to the return portion of the DD coil in the y-axis; this coupling is cancelled by flux in the opposite direction on the other end of the pad as shown in Figure 6-3(b). The effect is similar for coupling between the quadrature coil and sections of the DD coils in the x-axis however the cancelling flux is diagonally opposite as shown. The second case is for coupling around the flux pipe – the flux leaving one pole face is equal to the flux returning at the opposite pole face resulting in no net flux and hence no coupling.

Referring to flux above a transmitter pad as either horizontal or vertical is useful for describing the operation of DD or quadrature coils. The DD coils capture primarily horizontal flux while the quadrature coil is designed to capture vertical flux however, it is noted that both horizontal and vertical fluxes are surface integrals of each other.

In order to maintain the same flux path through the receiver ferrite as it is moved in the x-axis, slightly lengthened ferrite strips are required. The flux path remains largely the same but flux passes through the DD or the quadrature coils or both. The original DD pad is a 4x6 and the strip length has been increased by two I93 cores so the DDQ receiver becomes a 4x8. As the added ferrite cores are placed under the return portions of the receiver DD windings, they only add 6mm to the length of the pad (shown later in Figure 6-5).

The operation of the quadrature coil is shown by magnetic flux density vectors on an xz plane through an energised transmitter with an open circuited receiver positioned with an air gap of 125mm; the plane is referenced to the axes shown in Figure 6-4(a). The flux plot of Figure 6-4(b) shows a perfectly aligned receiver pad where there is only net flux (Φ<sub>M-DD</sub>) through
the DD receiver coils resulting in maximum coupling and they supply the full output power. The flux going into and out of the quadrature coil is identical resulting in no coupling.

When the receiver is laterally offset by 190mm in the $x$-axis, the quadrature coil is ideally positioned to capture the vertical flux as shown in Figure 6-4(c) and is able to supply the full output power. The DD coil nears a null in its $x$-axis VA profile and the null occurs at $\delta_x=260$mm as shown in Figure 6-2. The flux pattern above the transmitter (and through the receiver) is reasonably consistent as $\delta_x$ is increased due to the longer strips of ferrite in the receiver allowing better coupling between the DD transmitter coils and the quadrature receiver coil. The output power is a result of the mutual coupling to both the DD and the quadrature coils as the receiver moves between the two coupling extremes.

![Diagram](image)

**Figure 6-4:** (a) DD transmitter and receiver with reference axes. Magnetic flux density vectors in the $xz$ plane of a DDQ-DD system (b) horizontally aligned and (c) misaligned with a $\delta_x$ of 190mm ($\delta_z=125$mm and $I_1=23$A).
6.3.2. Optimisation of the quadrature coil

The arc shaped flux path means the size of the quadrature coil needs to be chosen so that the cancellation in the induced voltage is minimal (i.e. the amount of flux entering and exiting the area of the coil should be minimised). Flux cancellation is more likely with a large coil however if the coil is too small there will not be enough net flux resulting in poor overall performance, clearly there is an optimal size and this also depends on the size of the transmitter. An investigation of quadrature coils with different sized transmitters is however, beyond the scope of this thesis.

The width of the quadrature coil (shown in Figure 6-5(a)) was made the same as that of the DD coil. This dimension is not affected by the flux cancellation discussed above and the coil should be as wide as possible to increase flux capture but without increasing the size of the pad. In addition, the thickness of a built pad should not increase due to the addition of the quadrature coil and the arrangements shown in Figure 6-5(a) and (b) enable the quadrature coil to be fitted under the DD coil return sections and brought up in the pole face [115]. For simulation purposes, the quadrature coil is placed above the DD receiver coils resulting in a simple planar coil structure (similar to Figure 6-3(a)) that requires far fewer elements enabling faster calculation times.

![Figure 6-5: Quadrature coil length range with a minimum (a) 410mm and a maximum of (b) 570mm.](image)

Coupling between the DD transmitter and quadrature coils is dependent on both the length of the quadrature coil and the position of the receiver pad. Therefore the optimisation process involves taking $P_{su}$ profiles in the $x$-axis direction for quadrature coils of various widths. The profiles were done with a 125mm air gap and the results are shown in Figure 6-6(a).

The highest $P_{su}$ is achieved with a 450mm long coil however a length of 410mm was chosen given the $P_{su}$ is very close and there is an overall reduction in the length of Litz wire of 1.6m (20 turns but 40 ‘wires’ are shortened by 40mm). The longest quadrature coil possible given geometrical constraints is 570mm as shown in Figure 6-5(a). The flux cancellation is
relatively high as shown by the poorest peak $P_{su}$ in Figure 6-6(a) but when the offset is extreme (>250mm) one edge of the coil is still relatively close to the transmitter enabling better coupling. The 560mm long coil couples 2190VA with a $\delta_x$ of 280mm while the 410mm long coil couples 1780VA with the same $\delta_x$. The $V_{oc}$ and $I_{sc}$ for the chosen 410mm quadrature coil are shown in Figure 6-6(b) and the coupling coefficient in Figure 6-6(c).

![Figure 6-6](image)

Figure 6-6: (a) $P_{su}$ profile of quadrature coils of various widths, (b) open circuit voltage and short circuit current and (c) coupling coefficient for a 410mm long quadrature coil ($\delta_z = 125$mm and $I_1 = 23$A at 20kHz).

### 6.4. Combined DD and quadrature power profile

The total power output of the DDQ controller is supplied by the independently tuned outputs of the DD and quadrature coils as shown in the schematic of Figure 6-7. The quadrature coil is independently tuned (with $C_{2Q}$) and rectified before being added to the rectified output of
the DD. This circuit is similar to that presented in [82, 117] where a quadrature coil is added to a receiver for a distributed system. The rating of the switched mode controller for a DDQ receiver is similar to that of a standard DD receiver as the quadrature output complements the DD – it makes up for the drop in DD output and without additional tuning elements, it is not able to exceed the output of the DD coil.

The $P_{su}$ profiles in either the $x$-axis or $y$-axis of the DD and quadrature coils are shown in Figure 6-8(a). The horizontal tolerance of the DD receiver in the $x$-axis is significantly improved (better than in the $y$-axis) with the addition of the quadrature coil.

The coupling coefficients of the DD and quadrature coil against horizontal offset are shown in Figure 6-8(b). In the case where the DD and quadrature coils outputs are added, coupling is due to an effective coupling coefficient and this is referred to as $k_{eff}$, it is calculated from the individual contributions of the DD and quadrature coils as shown in (6.1) where the VA is the input VA to the transmitter pad.

$$k_{eff} = \frac{\sqrt{P_{suDD} + P_{suQ}}}{VA} \quad (6-1)$$
Figure 6-8: Horizontal profiles in either x or y-axes for the DD and quadrature coils (a) $P_{su}$ and (b) $k$ ($I_1 = 23\, \text{A}$ at 20kHz and $\delta_z = 200\, \text{mm}$).

6.5. Power loss in power pads

Power pads are designed to be as efficient as possible so that a good design has a real impedance component that is very small compared to its imaginary component. Accurately determining power loss is difficult because small phase errors can lead to large inaccuracies. In Chapter 4 the native quality factor, $Q_L$, was measured with an Agilent E4980A precision LCR meter and used to calculate the power loss in various power pad topologies. The LCR meter determines the quality factor based on small-signal excitation and this approach is generally only useful for relative comparisons of pad topologies. For example, the $Q_L$ of the flux pipe dropped from 260 to 86 with the introduction of aluminium shielding indicating a
large increase in loss in the aluminium. The large change in $Q_L$ provides a good indication that power loss will increase as aluminium is brought close to a powered pad however the absolute value of $Q_L$ should be determined when the pads are operated with their rated current. All power pads studied in this thesis use ferrite for field shaping and this non-linear material can lead to smaller practical $Q_L$ values than those measured with an LCR meter [118].

There are three materials that cause power loss in pads; these are ferrite, Litz wire (copper) and the aluminium shielding. The greatest of these three losses is in the copper, however the other two are not insignificant [32, 119]. Measuring the loss in an operating transmitter pad is not practically possible due to the large VA and relatively low real power dissipation. A typical Double D pad has a $Q_L$ of 350 therefore the real part of the impedance is only 0.29% of the total impedance resulting in a phase angle of 88.84°. Assuming 5% accuracy is required for the measurement, a phase resolution of 8.2m° is required. Current probes typically have a phase shift of around 1° and therefore are unsuitable.

The approach used here is to model the loss in the ferrite and combine that with a combination of measurements taken on a pad without ferrite and or aluminium. The power loss in the Litz wire is due to resistive, skin and proximity effects and if the wire is measured in isolation, a relatively accurate resistance can be measured at the operating frequency. The loss in the aluminium shielding plate is due to induced eddy currents and this can be measured relative to the coil with no ferrite present. The ferrite will alter the copper and aluminium losses due to field shaping and this is be investigated.

6.5.1. Loss in ferrite

The core loss ($P_c$) is divided up into three components as shown in (6-2) where $P_h$ is the hysteresis loss, $P_e$ is the eddy current loss and $P_r$ is the residual loss. The residual loss is a result of magnetic relaxation and dimensional resonances which cause the calculated eddy current losses to be smaller than measured. The error increases with frequency however such losses only dominate at MHz frequencies and as the power pads are operated at 20kHz, this is not of any significance [120].

$$P_c = P_h + P_e + P_r$$ (6-2)

The energy lost due to hysteresis can be found by integrating the exciting magnetic field (H) with respect to the flux density (B) for a complete cycle and multiplying it by the volume of
ferrite. This approach is difficult in practice because the area of hysteresis loop depends on the magnetic field strength and this varies with position along the ferrite strips. This creates an infinite number of small location dependent hysteresis loops within the main loop and such data is not usually provided. Also, the shape of the hysteresis loop is temperature dependent, the upper half of a loop for N87 material is shown in Figure 6-9(a) at a temperature of 25°C. At 100°C the loop is smaller indicating lower loss but saturation occurs at 380mT.

![Figure 6-9: (a) Dynamic magnetisation curve for N87 material at 25°C (b) core loss versus frequency at different flux densities [111].](image)

There are two categories of models that describe the loss characteristics of ferrite; these are physical models and behavioural models. Physical models, such as Jiles-Atherton (J-A), are based on impedances to domain walls caused by pinning sites as they move during magnetisation. Pinning sites are caused by non-magnetic inclusions in the material, voids or inhomogeneous stress and are intrinsic to ferrite. The J-A model is able to accurately generate sigmoid shaped hysteresis loops for all magnetic field values however the coefficients in the differential equations that describe the magnetisation need to be determined [121, 122].

Behavioural models such as the Stoll and Steinmetz are based on calorimetric measurements and are the most popular methods for determining loss in ferrite [123, 124]. The Steinmetz equation is shown in (6-3) where $P_c$ is the power loss, $V$ is the volume, $f$ is the frequency of the sinusoidal excitation and $\hat{B}$ is the peak flux density. For common power ferrites the exponent values are: $\alpha = 1.2-1.9$ and $\beta = 2.3-3$. The actual values of N87 material can be
determined by examining the power loss versus volume graph of Figure 6-9(b). Data is not provided for operation at 20kHz but the loss curves are smooth allowing easy extrapolation.

\[
\frac{P_c}{V} = k_f B^\beta \quad (6-3)
\]

### 6.5.2. Simulated loss in a transmitter pad

The approach used here is to apply a simplified Steinmetz equation to a simulation model of a Double D transmitter pad. The pad will only be operated at 20kHz therefore the frequency dependency has been removed and incorporated into the constant, \( k_f \) as shown in (6-4) below. The values of \( k_f \) and \( \beta \) are 4197879 and 2.21 at 25°C (from Figure 6-9(b)).

\[
\frac{P_c}{V} = k_f B^\beta \quad (6-4)
\]

The elements created by JMAG FEM software within a model are tetrahedrons and these convex solids have four triangular faces that meet to form four vertices, which are referred to as nodes. The FEM solver computes the magnetic flux density vectors (specified by 6 values which are the real and imaginary components in the \( x, y \) and \( z \) axes) at each node and these can be added together to determine the magnitude of the flux density for a given element. Core loss is calculated by searching for elements of ferrite material and applying the simplified Steinmetz equation as shown in (6-5). The number of ferrite elements is specified by \( m \) and there are typically 100,000 such elements in a large model, \( V_n \) is the volume of the element in m\(^3\) for which the loss is being computed.

\[
P_c = \sum_{n=1}^{m} k_f B_n^\beta V_n \quad (6-5)
\]

The 3D meshing process in the FEM software determines the size of the elements however this process is not deterministic and the element sizes vary within materials and at material boundaries as the simulator increases the size of non critical elements to reduce simulation time. The volume of an irregular tetrahedron can be calculated based from (6-6) where the coordinates of the four nodes (1-4) are specified by \((x_{1-4}, y_{1-4}, z_{1-4})\).
The loss calculation algorithm begins with finding an element of ferrite material, summing the flux density vectors at the four nodes and multiplying by $\sqrt{2}$ to determine the peak flux density. The volume of the element is calculated and the loss equation (6-5) is applied. The loss of the given element is added to the total loss and then the search begins for the next element of ferrite material up to a total of $m$ elements. The process is shown by (6-7).

$$P_c = \sum_{n=1}^{m} k_f B_n^\beta \frac{1}{3!} \begin{vmatrix} x_{1,n} & y_{1,n} & z_{1,n} & 1 \\ x_{2,n} & y_{2,n} & z_{2,n} & 1 \\ x_{3,n} & y_{3,n} & z_{3,n} & 1 \\ x_{4,n} & y_{4,n} & z_{4,n} & 1 \end{vmatrix}$$  \hfill (6-7)

The results of the ferrite loss model are shown for a 4x6 transmitter pad for various currents in Figure 6-10(a). The crosses indicate the simulated currents (with loss noted) and a trend line is also plotted. The core loss can be determined as follows: $P_c = 0.0965I_1^{2.2}$, where $I_1$ is the rms transmitter pad current. The exponent for the trend line is consistent with the Steinmetz exponent, $\beta$, which is 2.2 and is expected because the flux density within the ferrite strips increases linearly with current. The chosen 4x6 pads are operated with a current of 23A and the expected core loss is 102W.

The flux density on a contour line beginning at the centre of the outer ferrite strip and travelling along the half of the length of the outer strip (in the $x$-axis) is plotted in Figure 6-10(b). The pads can be operated with higher currents if more power is desired however the maximum should be 30A or saturation may occur (shown on the BH plot of Figure 6-9(a)).

The flux density plot of Figure 6-10(b) indicates proper operation of the flux pipe, the total spread of the coils is 240mm and only half of the coverage is shown (the flux pipe consists of two 120mm wide coil sections). The flux density is high under the coil (up to a distance of 120mm) and drops slowly moving outward along the length of the strip from the centre whereas is drops relatively rapidly at the pole face, which begins at 120mm. The flux leaves one pole face and heads toward the other but the magnetic pressure created by the flux on the shorter path lengths (between inner parts of the pole faces) forces flux outward resulting in a relatively low drop in flux density in the pole face.
Chapter 6 – Improvements to Double D pads

6.5.3. Loss in copper and aluminium

The power loss in copper and aluminium have been grouped together as they are considered the winding loss and given the symbol $P_w$. This loss cannot be determined by simulation due to the complex magnetic field interactions. The winding loss is made up of three components, these are due to the fundamental resistivity of copper and frequency dependent effects such as proximity and eddy currents that occur in the coil, and similarly as mirror currents in the aluminium backing plate. The coil is made up of several turns of Litz wire that is made up of insulated strands that are twisted together forming bundles and these are twisted to form the wire.
There are strand-level and bundle-level proximity and skin effects that occur within individual strands and the bundle respectively. With proper construction each bundle occupies every other bundles position along the length of the wire minimising bundle level proximity and skin affects, therefore the current ($I_1$) is distributed evenly among the bundles. The strand level skin effect is reduced by choosing a strand diameter smaller than the skin depth at the operating frequency, leaving strand level proximity effects as the dominant loss mechanism. These strand level proximity effects are due to internal magnetic fields created by strands or by the external magnetic field acting on the strand [125, 126].

Simulating winding loss would require each of the 810 strands that make up the Litz wire to be individually modelled requiring a level of complexity and computation power that is not currently feasible. Also, the induced currents in the aluminium backing plate need to be accounted for requiring a few skin depths to be simulated. In consequence, an alternative but suitably accurate approach is to measure the ac resistance of a coil in situ to account for all of the above effects. Small signals generated with a precision LCR meter can be used because the winding loss occurs in linear materials unlike core loss. The test current from the LCR meter is a few mA and from the graph of Figure 6-10(b) it can be inferred that the flux density will only be a few $\mu$T resulting in a very small core loss.

Initial winding loss measurements were made on the 420mm circular pads described in Chapter 3.2 because these can be easily disassembled allowing various measurement combinations such as: the coil only, or coil and ferrite, or coil and aluminium, or coil, ferrite and aluminium. The loss due to eddy currents induced in the aluminium backing plate could not be explicitly measured because the ferrite screens a significant proportion of the underside area of the coil (clearly only the field below the coil is able to create loss in the aluminium). In practice circular pads would never be operated without ferrite due to poor power transfer caused by extremely low inductance. Measurements show the coil of the 420mm diameter circular pad on its non-magnetic former has twice the inductance in air than it does when placed in an aluminium case without ferrite.

The winding loss, $P_w$, is shown in (6-8) where $I_1$ is the transmitter pad current, $F_r$ is a factor which relates the ac winding resistance to the dc resistance and $R_{dc}$ (the dc resistance of the coil).
Chapter 6 – Improvements to Double D pads

\[ P_w = I_1^2 R = I_2^2 F_r R_{dc} \]  

(6-8)

The measured parameters of the various pad configurations are shown in Table 6-1 where the winding loss is calculated assuming the current is 23A at 20kHz. Here the label ‘coil’ refers to measurements of the coil held in its non-magnetic former away from any magnetic or conductive materials, the label ‘Fe’ means the strips of ferrite have been placed in the former and the label ‘Al’ indicates the coil former has been placed the pad case. The label ‘Fe+Al’ means the former with ferrite has been placed in the aluminium pad case. The loss with 23A dc is 14.1W and this increases by 3.7W to 17.8W for a coil in air at 20kHz. This 21% increase in resistance is due to proximity and eddy current effects and these are not insignificant. As shown, the coil in air has an inductance of 52μH and this increases dramatically in the presence of ferrite whereas it drops when the aluminium is present. The coil loss doubles if it is placed in the aluminium pad case (consists of a backing plate and ring as shown in Chapter 3), whereas the presence of ferrite mitigates this loss.

<table>
<thead>
<tr>
<th>Config.</th>
<th>L (μH)</th>
<th>Q_L</th>
<th>R_{ac} (mΩ)</th>
<th>P_w (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coil</td>
<td>52</td>
<td>194</td>
<td>33.6</td>
<td>17.8</td>
</tr>
<tr>
<td>Fe</td>
<td>78</td>
<td>273</td>
<td>34.6</td>
<td>18.2</td>
</tr>
<tr>
<td>Al</td>
<td>23</td>
<td>38</td>
<td>76.3</td>
<td>40.4</td>
</tr>
<tr>
<td>Fe+Al</td>
<td>70</td>
<td>142</td>
<td>60</td>
<td>31.7</td>
</tr>
<tr>
<td>dc loss</td>
<td></td>
<td></td>
<td>26.3</td>
<td>14.1</td>
</tr>
</tbody>
</table>

TABLE 6-1 PARAMETERS OF 420MM CIRCULAR PAD MEASURED AT 20KHZ (I = 23A)

The graph of Figure 6-11 shows \( F_r \) for various pad configurations against frequency. The minimum measurement frequency the LCR meter allows in 20Hz and the value at this frequency is considered to be \( R_{dc} \) as the frequency effects are negligible. The ac resistance factor of the 4x6 DD pad described in Chapter 5 with and without the aluminium backing plate is plotted against frequency in Figure 6-12.

The ac resistance, \( Q_L \), winding loss and resistance factor are listed for various frequencies in Table 6-2. A DD pad with the aluminium plate present has an \( F_r \) of 1.44 at 20kHz whereas it is 2.2 for the 420mm circular pad. The increase is mainly due to the distance between the coils and the aluminium backing plates, both pads have coils that sit on ferrite strips but the circular pad uses 10mm thick cores whereas the DD uses 16mm thick cores that sit on a 4mm buffer. As such, the coil is a further 10mm away from the aluminium resulting in lower eddy currents and hence loss. The winding loss without the backing plate is 82W at 20kHz and this increases by 16W to 98W when the backing plate is added.
6.5.4. Loss model

The 4x6 DD pads in this thesis are designed for an EV charging system that operates at 20kHz. Here the data from sections 6.5.2 and 6.5.3 are combined to create a suitable loss model at this frequency. The loss in the DD pad is the sum of the winding and core losses ($P_{loss} = P_c + P_w$) and this is plotted in Figure 6-13 as a function of current (assuming the pad uses “single-filar” coils, with 20 turns as shown in Figure 5-2). It should be noted that here the winding loss is calculated based on $R_{ac}$ measurements with an LCR meter and while the exciting current is small (mA), it should not be totally dismissed. There will be a slight traversal of the hysteresis loop that means $P_w$ may include some core loss. In consequence, the calculated $Q_L$ using the combined simulated core loss and measured winding loss will...
Chapter 6 – Improvements to Double D pads

have some overlap so that actual loss will likely be lower than the presented theory predicts. The calculated winding and theoretical core losses using this technique are predicted to be 98W and 102W respectively when the pad is operated with a current of 23A resulting in a total loss of 200W. The measured $Q_L$ with an LCR meter is 400 however the $Q_L$, calculated by the $\text{VA}/P_{\text{loss}}$, is 196. This is the absolute lowest value of $Q_L$ possible and in practice due to the overlap of measurement in core loss, a $Q_L$ somewhere between 200-300 is likely.

The results of Figure 6-13 show that with pad currents less than ~20A the core and winding losses are similar however the core loss increases at a greater rate. The efficiency of a system will decrease if the track current is increased to transfer more power because core loss is proportional to $I_1^2$ but $P_{\text{su}}$ is proportional to $I_1^2$.

![Figure 6-13: Total, winding and core loss in a 4x6 DD pad operating at 20kHz ($I_1$ is the rms current).](image)

6.5.5. Measured pad loss

The approach used here to roughly determine the loss in a DD pad is to excite an equivalent inductance and measure the input power to the system, then connect a fully compensated DD pad and measure the input power again. The increase in power over the base level will largely be the loss in the pad however there will be some increase in the loss in the IPT power supply and compensation network due to the additional power dissipated in the DD pad and tuning effects. This loss is not expected to be large given the real power is in the range of ~100W whereas the power supply and its resonant network are rated for 2kW. An increase of 20% is assumed here which as shown, matches the theoretical analysis above, however if the percentage loss in the power supply is greater than 20%, then the effective $Q_L$ will be higher.
Chapter 6 – Improvements to Double D pads

The experimental set up is shown in Figure 6-14(a). An Agilent 6813B power supply is used and it is able to measure the input power to the system. The compensated DD pad is added in Figure 6-14(b). Due to component ratings this experiment is done with a currents of 15A and 20A per turn in the DD pad (noting the pad used is bifilar wound therefore the input currents are 30A and 40A).

![Figure 6-14: Pad loss experiment (a) measuring base loss and (b) measuring loss with compensated DD pad added.](image)

The input voltage ($V_{in}$), current ($I_{in}$) and power ($P_{in}$) to the IPT supply (shown above in Figure 6-14) are listed in Table 6-3 where the two base values are for the loss in the power supply and equivalent inductor at the two current levels. The power losses are presented in Table 6-4 where $\Delta P$ shows the power difference with and without the DD pad and its compensation capacitor (here 80% of the power dissipation is assumed to be in the added components).

The measured loss difference here also includes the loss in the capacitor (comprising parallel individual capacitors) shown in Figure 6-14(b) as the full current passes through it. This loss, $P_{cap}$, is calculated from datasheet ESR values and is subtracted from $P_{loss}$ along with the winding loss ($P_{w}$). The simulated core loss values are 40W and 75W with currents of 15A and 20A respectively (shown in Figure 6-10(a)). The results are considered to be in good agreement although the simulated values are conservative.

Further work is needed to properly characterise the loss modes in the DD pads. These preliminary results show there is additional core loss in the pads that cannot be accurately determined with small signal excitation from an LCR meter – the power loss is considerably greater than expected when the measured $R_{ac}$ value is used.
## 6.6. Compliance with ARPANSA and ICNIRP guidelines

The IPT EV charging system should comply with guidelines from the International Commission on Non-Ionising Radiation Protection (ICNIRP). ICNIRP stipulates that the general public should not be exposed to body average rms flux densities greater than 27\(\mu\)T in the frequency range of 3kHz-10MHz [89]. Local spot measurements are allowed to exceed this 27\(\mu\)T limit provided the spatial average remains within the guidelines. The Australian Radiation Protection and Nuclear Safety Agency (ARPANSA) have based their reference levels on earlier (1998) ICNIRP guidelines and suggest taking the average exposure level at four points on the human body: the head, chest, groin and knees. The spot limits are allowed to be 27.3\(\mu\)T, which is \(\sqrt{20}\) greater than the body average of 6.1\(\mu\)T, providing the body average is still met. More information regarding these exposure standards can be found in Chapter 2.

### 6.6.1. Leakage flux around Double D pads

The symmetrical nature of circular pads means that only a single line is needed extending outward from the middle of the air gap to characterise the worst case leakage flux as shown in Chapter 3.8.3. However, as DD pads are polarised, the flux density needs to be considered in a plane situated between the transmitter and receiver pads. To determine the plane on which leakage was most significant, a simulation was run with a 200mm air gap and the transmitter energised with a current of 23A.

---

### TABLE 6-3 INPUT POWER TO IPT POWER SUPPLY

<table>
<thead>
<tr>
<th>Desc.</th>
<th>(V_{in}) (V)</th>
<th>(I_{in}) (A)</th>
<th>(P_{in}) (W)</th>
<th>(I) (per turn)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Base</td>
<td>215</td>
<td>0.23</td>
<td>49</td>
<td>15</td>
</tr>
<tr>
<td>Base</td>
<td>290</td>
<td>0.3</td>
<td>87</td>
<td>20</td>
</tr>
<tr>
<td>DD pad added</td>
<td>215</td>
<td>0.69</td>
<td>148</td>
<td>15</td>
</tr>
<tr>
<td>DD pad added</td>
<td>290</td>
<td>0.94</td>
<td>273</td>
<td>20</td>
</tr>
</tbody>
</table>

### TABLE 6-4 POWER LOSSES

<table>
<thead>
<tr>
<th></th>
<th>(I = 15A/\text{turn})</th>
<th>(I = 20A/\text{turn})</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\Delta P)</td>
<td>99</td>
<td>186</td>
</tr>
<tr>
<td>(P_{\text{loss}})</td>
<td>79</td>
<td>148</td>
</tr>
<tr>
<td>(P_{w})</td>
<td>42</td>
<td>74</td>
</tr>
<tr>
<td>(P_{\text{cap}})</td>
<td>2</td>
<td>3.5</td>
</tr>
<tr>
<td>(P_{c})</td>
<td>35</td>
<td>71</td>
</tr>
</tbody>
</table>

---
The reference axes are shown in Figure 6-15(a) where the length of the pads is in the $x$-axis, the width is in the $y$-axis and the lid of the transmitter (not shown) is at $z = 0$. A contour showing the magnetic flux density in the $xy$ plane midway in the air gap ($z = 100$) is shown in Figure 6-15(b) noting the red indicates a density of $50\mu$T. A 0.5m radius circle is drawn centred on the transmitter and it appears that leakage is much worse in the $x$ direction, however the line segments of length $l$ which are added to the figure show that the leakage is similar at equivalent distances away from the pad edges in the length and width axes. The line segment, which is placed at the corner of the pad at an angle of $45^\circ$ shows leakage is less ‘off axis’, so the leakage in both the $x$ and $y$-axes is all that is needed to characterise worst case leakage around a DD pad.

A flux density contour in the $xz$ plane is shown in Figure 6-15(c). Notably, the leakage flux extends below the transmitter pad, but as this will typically be below ground level this flux is not likely to present a problem to humans. In practice the chassis of the vehicle will naturally provide shielding above the receiver or shielding will be added to reduce potential power loss. The degree of shielding depends on the type of material, the effect of steel can only be measured rather than simulated since large surfaces with small skin depths are likely to be present in a real vehicle requiring an impractically large number of elements for accurate simulation. Due to the variability of materials, the main concern in this thesis is the flux present in the $xy$ plane because people could be exposed to this.
6.6.2. Energy in the leakage magnetic field

The cyclic storage of energy in the inductance of a power pad results in an alternating magnetic field about the pad. The flux density contours of Figure 6-15(b) and (c) indicate energy is mainly stored in between the pads and in the air gap extending in the $xy$ plane. Clearly only the energy that can be effectively extracted by the receiver is useful in practical applications however, considering energy allows the percentage of the leakage magnetic field created by a transmitter to be determined. In this work the leakage field is considered to be any magnetic field that is not within the boundary edges of the pads – the flux within the air gap creates coupling thus is useful. The energy stored by an inductor ($W$) is equal to the energy stored in the magnetic field as shown by (6-9) where $\mu$ is the permeability of the volume ($V$) of the material being considered [112].

$$W = \frac{1}{2}LI^2 = \frac{B^2}{2\mu}V$$  \hspace{1cm} (6-9)

The flux density drops to approximately 10µT at a distance of 500mm from the edges of the vertically aligned pads as shown in Figure 6-15(b), therefore there will be little energy stored
in the volume of air significantly further away. As such this is considered the outer boundary for the leakage flux zone. In order to determine the energy stored in a particular area using JMAG FEM software, a unique material needs to be specified; in this case a 300mm high non-magnetic non-conductive block of material extending 500mm further from the pad edges in the length and width axes is used (as shown in Figure 6-16). The pads are symmetrical about the $x$ and $y$-axes thus the total energy is four times the energy in the block. The amount of energy is determined by summing the energy in each element in a similar manner to the loss modelling method in section 6.6.2.

The results are presented in Table 6-5 for a current of 23A where $W_L$ is the total energy stored by the transmitter pad, $W_1$ is the energy in the leakage field (in the volume four times the size of the block of Figure 6-16). The leakage magnetic field in the block surrounding the pad is 0.31% of total field created by the transmitter pad. At an operational frequency of 20kHz this means there is a power availability in the leakage field ($P_l$) of 9.6W, this corresponds to a leakage field power density ($W_l \rho$) of 15Wm$^{-3}$.

![Figure 6-16: One quarter of the leakage field zone.](image)

<table>
<thead>
<tr>
<th>Block ($\mu_r = 1$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>200</td>
</tr>
<tr>
<td>82</td>
</tr>
<tr>
<td>500</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>L</th>
<th>583.5 μH</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_L$</td>
<td>154 mJ</td>
</tr>
<tr>
<td>$W_1$</td>
<td>0.478 mJ</td>
</tr>
<tr>
<td>Leakage</td>
<td>0.31 %</td>
</tr>
<tr>
<td>$P_l$</td>
<td>9.6 W</td>
</tr>
<tr>
<td>$W_l \rho$</td>
<td>15 Wm$^{-3}$</td>
</tr>
</tbody>
</table>

**Table 6-5 Energy in the leakage field of DD pads**

### 6.6.3. Double D leakage compared with circular pads

The magnetic field leakage around 2kW systems using DD pads and circular pads are investigated via simulation, this approach is viable given measurements taken with a Narda ELT-400 exposure level tester (probe area of 100cm$^2$) differed from simulated results by ~5% for the circular pads in Chapter 3 and in [127]. Compliance with magnetic field leakage guidelines or standards are difficult to determine because leakage fields can be investigated with pads being either centred or with horizontal offset. The ferrite in the receiver attracts
flux therefore leakage increases slightly as the receiver is offset horizontally leaving part of the transmitter exposed. In addition, the coupling coefficients of DD and circular pad systems are very different with the same air gap therefore comparing their leakage fields cannot be done directly due to different receiver operational $Q$s (these are required to ensure the same power output of both the systems).

To minimise the effect of coupling variations and account for operation with horizontal offset the magnetic field leakage is investigated assuming both receiver pad topologies have equal operational $Q$s and provide equal output power. Both the 4x6 DD and 700mm diameter circular pad systems have similar areas and self inductances, and both were designed to transfer 2kW with a 200mm air gap. However, the DD pads offer significantly better coupling and therefore allow far greater tolerance to horizontal misalignment than the circular pads. Circular pads have a null in their horizontal power profile with an offset of 260mm however when DD pads are offset by this amount in the $y$-axis, they are still able to transfer 43% of the available power when they are perfectly aligned.

The $y$-axis is chosen for comparison because the DD pads are more sensitive in this direction and therefore the leakage field will be higher. The 700mm diameter circular pad allows the system to transfer 2kW with a radial horizontal tolerance of 130mm and therefore operation at this horizontal offset is used to compare field leakage against a DD system. Any larger offset and the comparison becomes invalid as the current in the circular pad will have to be increased substantially to ensure any useful power transfer due to the null.

With a transmitter pad current of 23A the $P_{su}$ of a circular pad offset by 130mm is 338VA and for a DD system it is 1255VA when offset by the same amount in the $y$-axis. The current in the circular transmitter pad needs to be increased by a factor of $\sqrt{3.7}$ to 44.3A to ensure both pads have the same $P_{su}$ and therefore operational $Q$. Although this study is done for comparison purposes, the current in a circular pad can be increased by using wire with a greater cross-sectional area. in the ferrite strips Saturation is unlikely as the flux density in the ferrite strips increases linearly with current ($B_{max}$ with 32.5A is 113mT) therefore with 44.3A the peak flux density will be 217mT. A comparison where the current in a DD transmitter pad is reduced is still fair but is unlikely in practice as the transmitter will be operated at the rated current to ensure the designed power will be transferred to maximise the cost efficiency.

Three measurement axes are required to fully characterise the worst case leakage for both DD and circular pads, because horizontally misaligned pads are no longer symmetrical, a fourth
axis is not needed because the DD receiver is offset in the y-axis only. The origin of the measurement axes is in the middle of the air gap centred on the transmitter pad as shown in Figure 6-17(a) and (b) for circular and DD transmitter pads respectively.

![Figure 6-17: (a) Measurement axes for DD pads and (b) measurement axes for circular pads.](image)

The graph of Figure 6-18 shows the simulated leakage flux, where the direction along an axis is indicated by – or + with respect to the pad diagrams of Figure 6-17(a) and (b). For this 130mm offset, the DD system has more magnetic field leakage than circular pads after a distance of 0.6m from the transmitter centre regardless of axis. At a distance of 1.5m from the centre the DD leakage is ~4µT compared to ~1µT for the circular pads and these values are well within the ICNIRP and ARPANSA limits. A spot flux density of 27µT is reached at a distance of ~0.84m in the +x direction for DD pads and at ~0.62m if the offset is in the +y direction – the latter is similar to the circular pad. These distances are within half of the width of a small vehicle so that both pad topologies should easily comply with leakage standards at this power level.

The energy in the leakage field is relatively small compared to the energy transferred between the pads as shown in section 6.7.2. The relatively poor coupling between the misaligned circular pads means the current has to be increased resulting in a far greater flux density in the air gap. For circular pads the flux density at the origin of the axes shown in Figure 6-17(a) is 3.3mT whereas it is only 1.6mT for the DDs (shown in Figure 6-18). This is a very important because DDs are far less likely to heat stray metallic objects between the pads because the heating energy is proportional to $B^2$. Double Ds present a much lower safety risk as the available heating energy is less than a quarter that of a circular pad for the same power transfer.

The magnetic field leakage does not change substantially with this 130mm horizontal offset and it can be assumed that leakage flux increases in proportion with transmitter current. The resonant current in the receiver pad coil also contributes to the leakage field but the VA is
relatively low compared to the transmitter given operational Qs are constrained to less than 6, and this has been found to add only a small amount to the leakage [127]. If the above comparison was undertaken at larger offsets, the leakage field from the circular pads would be significantly greater than the DD-DDQ. As noted earlier, the field leakage simulations have been done with pads in air, however in practice the EV body and chassis will absorb energy and further attenuate the leakage flux.

![Figure 6-18: Leakage magnetic flux density for DD and circular pad systems.](image)

### 6.7. Interoperability

The advantages of IPT charging are well recognized and there has been significant development of systems in academia and industry as referenced in Chapter 2. At present circular pads are the most common even though they fundamentally offer poor performance and are not suitable for RPEV applications as discussed in Chapter 7. In this section an interoperability investigation is done to determine how a DDQ receiver works with a circular transmitter and how a circular receiver works with a DD transmitter. The investigation is done via simulation given a model containing a 700mm diameter circular transmitter and 4x8 DDQ receiver (with a quadrature coil length of 560mm) was successfully compared with real measurements to ensure accuracy [115]. As shown in Chapter 3.6, the performance of the 700mm diameter circular pads improves significantly if the shielding ring is moved further away from the ferrite spokes. In order to ensure the best performance of the circular pads (both in conventional and interoperable systems), the ring was positioned 40mm away from the ferrite resulting in a total diameter of 800mm (including 10mm for the thickness of the
ring). The quadrature coil width was set to 410mm as this offers the best compromise between copper use and coupling as shown in section 6.3.2.

The three possible combinations referred to by ‘receiver on transmitter’ for interoperability are: circular on circular, DDQ on circular and DDQ on DD. Interestingly, the quadrature in a DDQ receiver is effectively circular therefore gives good indication of how a circular receiver with linear strips of ferrite will operate with a DD transmitter (a true circular pad has radial strips of ferrite). The first series of simulations investigates the vertical profiles of the three pad combinations with and without horizontal offset, which is required to compare the output of the quadrature coil because it does not couple any power when vertically aligned above a DD transmitter.

Horizontal offset is also required for a DDQ receiver above a circular transmitter, as there is no net flux through the DD coils when centred (coupling is only due to the quadrature coil). The horizontal offsets at which the vertical profiles are taken are at the position of maximum coupling for the given coil and this has been determined by taking separate horizontal profiles. The 2D horizontal profiles are used to create charge zones in section 6.7.3, these show the area in which full power can be delivered in a battery charging application assuming the receiver pad is constrained to an operational $Q$ of 6.

### 6.7.1. Vertical profiles

Vertical profiles of $P_{su}$ and $k$ for air gaps ranging from ~100mm-250mm are shown in Figure 6-19(a) and (b) respectively assuming the transmitter and receiver pads are perfectly aligned ($\delta_x = \delta_y = 0$). The DDQ-DD combination offers significantly better power transfer than either of the other two. Although the DDQ receiver pad has not been specifically designed for operation with a circular transmitter it still achieves an almost identical performance noting the quadrature provides all of the output power when centered.

The coupling between the quadrature coil of a DDQ receiver and a circular transmitter is comparable to that of a circular receiver because the coils are similar in size. The quadrature coil measures 450x391mm and the circular transmitter is 416mm in average diameter. The VA output of the DDQ on a circular transmitter is close to that of a circular on circular design regardless of vertical offset therefore the DDQ is considered completely interoperable. If circular transmitter pads are present in a charging location, a car with a DDQ receiver can magnetically couple with this transmitter and if operating at the same tuned frequency can
feasibly receive almost an identical full power charging rate as an EV with a circular coupler specifically designed to operate with that same transmitter pad. The DDQ could easily be modified to further improve its coupling with a circular transmitter, but this is left as future work.

![Figure 6-19: Vertical profiles for different pad combinations (a) \( P_{su} \) and (b) \( k \) \( (I_1 = 23A \text{ at 20kHz}) \).](image)

When a DDQ receiver is centered above a DD transmitter the DD coils are in their maximum coupling position whereas when the DDQ receiver is offset by 180mm in the \( x \)-axis, the quadrature coil is in its maximum coupling position. When a DDQ receiver is offset by 200mm in the \( x \)-axis above a circular transmitter the DD coils are in their maximum coupling position. The vertical \( P_{su} \) profiles in Figure 6-20(a) demonstrate the DDQ receiver provides significantly better coupling above a circular pad than a matched circular receiver. A circle on circle system (shown in Chapter 3) has a \( P_{su} \) of 97VA when offset by 200mm in the \( x \)-axis with an air gap of 200mm whereas a DDQ receiver couples 680VA under the same conditions as shown in Figure 6-20(a). The coupling coefficient for the DDQ receiver above circular and DD transmitters is shown in Figure 6-20(b).
6.7.2. Power profiles with horizontal offset.

Horizontal VA profiles for the three pad combinations pads are shown on 3D plots that illustrate the output of the individual and summed receiver coils (total output) in Figure 6-21(a)-(e). The vertical air gap for all simulations is set to 125mm and $I_1$ is 23A at 20kHz. The $P_{su}$ profiles of Figure 6-21(a)-(e) are arranged such that Figure 6-21(a) is the profile of a circular pad on a circular pad and two sub plots on the left show results for a DDQ excited with a circular transmitter while the sub plots on the right show the DDQ receiver operating above a DD transmitter.

The profile for a circular on circular system shown in Figure 6-21(a) clearly indicates the power null that is intrinsic to circular pads. All circular profiles pass through this null regardless of the $P_{su}$ when the pads are horizontally aligned and this imposes a fundamental tolerance limit that makes them less ideal for stationary charging without alignment means and unsuitable for RPEV as the power falls to zero before reaching the next pad.

The profile for a DDQ excited by a circular transmitter is shown in Figure 6-21(b) with the outputs of the individual coils labelled. The peak $P_{su}$ is similar to that achieved with a circular receiver (shown by Figure 6-21(a)) and occurs when the receiver is vertically aligned. This is expected because the quadrature coil approximates a circular receiver and therefore provides very similar coupling. The horizontal tolerance is significantly better in the x-axis due to the contribution from the DD coils as these are almost able to couple as much power as the quadrature coil.
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The 2D horizontal VA profile of Figure 6-21(d) confirms the DDQ is completely interoperable with a circular transmitter. The thickness of all the pads is the same however the surface area of the DDQ is 0.32m$^2$ and the circular pad is 0.38m$^2$. Given the overall material requirements for a DDQ are only marginally greater than the circular, it would be preferable to have a DDQ receiver over a circular rather than a circular over a circular because it offers far greater tolerance to horizontal misalignment.

The $P_{su}$ profiles for the DD and quadrature coils of a DDQ receiver are shown in Figure 6-21(c) when it is excited with a DD transmitter. The VA output and horizontal tolerance of the DD coil alone are significantly better than that possible with a circular pad system. Here the quadrature coil contributes significantly to the overall output when the receiver is misaligned and delivers close to its maximum output as the DD enters its null. The annotated part of the surface where $\delta_x$ is large (and $\delta_y$ is small) is the VA output of the DD after it has passed through the null.

The total VA coupled by both the DD and quadrature coils when excited by a circular transmitter (Figure 6-21(d)) and the DD transmitter (Figure 6-21(e)) are also shown for comparison. These combinations produce far greater charge zones than possible with circular receivers. The power profiles are both very smooth in the $x$-axis due to the balanced profiles of the individual coils shown in Figure 6-21(b) and (c). These profiles demonstrate that the DDQ receiver offers significantly better performance than a circular on circular receiver and preferably should be used in EV charging systems.
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6.7.2.1. Open circuit voltage and short circuit current horizontal profiles

The open circuit voltage and short circuit current for the circular on circular and DDQ on DD systems are shown in Figure 6-22(a)-(f).

Figure 6-21: Power profiles for receivers with different or combined coils on different transmitter pads: (a) circular on circular, (b) DD and Quad. Rx. on circular Tx., (c) DD and Quad. Rx. coils on DD Tx., (d) DD+Quad. Rx. on circular Tx. and (e) DD+Quad. Rx. on DD Tx ($I_1 = 23A$ at 20kHz).
Figure 6-22: (a) $V_{oc}$ and (b) $I_{sc}$ horizontal profiles of a circular receiver on a circular transmitter, (b) DDQ receiver on a DD transmitter (a) $V_{oc}$ of DD coils, (b) $I_{sc}$ of DD coils, (c) $V_{oc}$ of Quad. Coil and (d) $I_{sc}$ of Quad. coil ($\delta z = 125$mm, $I_1 = 23$A at 20kHz).
The null in the $V_{oc}$ profile for the DD coils of a DDQ receiver occurs at an offset of 260mm in the $x$-axis as shown in Figure 6-22(c). The horizontal profiles of the DD and quadrature coils above a circular transmitter are plotted in Figure 6-23(a)-(d). The lengthened strips of ferrite permit the quadrature coil to couple power at extreme offsets in the $x$-axis as shown in Figure 6-23(c) and (d). When a circular pad is similarly offset above a circular transmitter there is very little coupling as shown in Figure 6-22(a) and (b).

![Horizontal profiles of a DDQ receiver on a circular transmitter](image)

**Figure 6-23**: Interoperability - horizontal profiles of a DDQ receiver on a circular transmitter (a) $V_{oc}$ of DD coils, (b) $I_{sc}$ of DD coils, (c) $V_{oc}$ of Quad. coil and (d) $I_{sc}$ of Quad. coil ($\delta_z = 125$mm, $I_1 = 23$A at 20kHz).

### 6.7.3. 7kW charge zones

Charge zones define the physical operating area where the desired power can be delivered given a particular air gap and operational $Q$. In this work a maximum $Q$ of 6 was assumed.
and the air gap was set to 125mm. The results of a DDQ receiver pad operating on DD transmitter can be compared with this same pad operating on circular pad in Figure 6-24(a) and (b) respectively. Both plots show the charge zone possible if a circular on circular system were used for comparison.

Notably the physically smaller DDQ-DD pads significantly outperform the circular pads and a DD coil alone provides an oval charge zone 350mm wide and 440mm long, which is large enough to enable parking without electronic guidance. Either the quadrature or DD coil can be used to supply the full output power in the regions where the DD and quadrature charge zones overlap. The region outside the explicit DD and quadrature charge zones (indicated by DD+Q in Figure 6-24(a)) shows the output of either coil is not enough to provide the desired 7kW but when both coils are combined the power output is ≥7kW. The small DD ‘zones’ when δx is very large show the DD has passed through its null and is able to supply the required 7kW.

The charge zone for a DDQ on a circular pad is shown in Figure 6-24(b) – this is a far larger zone than that possible with circular pads only. The DDQ receiver is considered to be completely interoperable with systems based on circular pads and as shown an EV will fundamentally have more tolerance.
Figure 6-24: 7kW charge zones for different pad combinations ($Q_{\text{max}} = 6$) (a) circular on circular and DDQ on DD, and (b) circular on circular and DDQ on circular ($\delta_z = 125$mm, $I_1 = 23$A at 20kHz).
6.8. Conclusion

Horizontal $P_{su}$ profiles taken on a DD-DD system show the DD receiver has relatively poor tolerance to misalignment in the $x$-axis due to a coupling null that exists at 260mm. Despite this the DD pads still offer far better performance than circular pads. To make up for this sensitivity, a magnetically decoupled coil, referred to as a spatial quadrature coil, was added to the DD receiver and this new receiver was referred to as a DDQ. The length of the quadrature coil was optimised by taking power profiles in the $x$-axis and it was found for a 4x8 structure that a length of 410mm provided a good compromise between the peak power delivered and the length of copper required.

Power loss in a DD transmitter pad is due to current in the coil, core loss in the ferrite and eddy currents in the aluminium base plate. The loss in the winding and base plate can be measured using the ac resistance with an LCR meter, however core loss cannot be measured as the LCR meter uses small signal excitation. Copper loss is due to the dc resistance of the wire together with ac factors such as proximity and eddy current effects. Although these are minimised by the use of Litz wire their effects are not negligible. The core loss has been determined by applying a simplified Steinmetz equation, with coefficients determined from loss data provided by the ferrite manufacturer, to a simulation model. The core loss for a DD transmitter operating with a current of 15A is has been roughly measured to be 35W while the simulation results indicates the loss is 40W.

The leakage magnetic field produced by a DD system was compared to that of a circular pad system with both receivers operating at the same $Q$, noting the circular transmitter current was increased. This enables a fair comparison given the coupling coefficients vary considerably. With an air gap of 200mm the $k$ for the DD is 0.23 whereas it is 0.16 for circular pads. Both systems were easily able to meet ICNIRP guidelines and ARPANSA standards and while the drop off in flux density from the edge of the circular pad is greater, the DD pads are still preferable. The heating energy available to foreign objects sitting in the air gap is proportional to $B^2$ and because the DD pads transmitter throws flux high toward the receiver (reflected by the high coupling coefficient) less current is needed to get the same power transfer compared to circular pads. In consequence the heating energy available between DD pads is less than one quarter of that which would arise with circular pads resulting in significantly safer systems.
Circular pads are the most common in the literature and industry therefore an interoperability study was done with a DDQ receiver above a circular pad. Charge zones, which illustrate the area in which full power can be delivered, show the DDQ is completely interoperable with traditional circular pads and actually offers far greater tolerance than a circular receiver. A DDQ receiver is the preferred type and is suitable for both stationary and roadway powered operation as discussed in the next chapter.
Chapter 7. IPT systems for roadway powered electric vehicles

7.1. Introduction

The work done in preceding chapters considered the design of couplers for stationary EV charging systems. In this chapter the design of coupling topologies that can continuously transfer power to a moving vehicle are investigated. These vehicles are referred to as Roadway Powered Electric Vehicles (RPEV) and the power may be used for propulsion as well as charging the battery with the ultimate aim of range extension. At present the battery is the most expensive component in an EV and it must have a large capacity in order to achieve a comparable range to conventional gasoline powered vehicles (~400km). Continuous power transfer means EVs only need to have a range sufficient to reach the nearest powered section of road. The reduced battery capacity coupled with such a roadway powered option can make EVs as cost effective as conventional vehicles [6].

This chapter begins with looking at traditional track based distributed systems in the context of the demanding requirements of RPEVs, these are: large power transfer (10s of kW), large horizontal tolerance across a powered lane and a large operational air gap. Single phase tracks are initially considered but this topology is extremely sensitive to horizontal misalignment therefore not suitable for an unguided EV. Poly-phase tracks are therefore investigated as an alternative. The existing three phase track topology is shown to be non-ideal and a new topology, referred to as a uni-polar track, is proposed [128, 129].

A small scale laboratory prototype of a uni-polar track is built for testing the topology and to calibrate simulation models. Subsequently, the system is scaled up via simulation to meet the power and tolerance requirements of an EV. Initial results show that ferrite is required under the track to enhance power transfer. Using a large slab of ferrite is impractical and not cost effective considering hundreds of kilometres of roadway are ultimately required. A similar approach is used here to remove excess ferrite as in Chapter 3, where circular pads were optimised.

In the second part of this chapter, a radically different approach to realising an RPEV system is presented using the DD couplers developed in Chapter 4. This topology is referred to as a quasi-lumped distributed type because it consists of discrete transmitter pads distributed
along the road to create a continuous magnetic field like that of a traditional distributed system. Only the transmitter pads under the vehicle are switched on enabling power transfer to a chassis mounted DDQ receiver (developed in Chapter 6). The advantages of such a system are the high degree of modularity, compliance with magnetic field exposure guidelines and a universal receiver, which can be used for both stationary and dynamic coupling.

Appropriately sizing the pads to couple the 20-60kW required by an RPEV and determining the optimal spacing to minimise resource use requires a considerable amount work and is a thesis in itself! In consequence, the work done here is largely a scoping exercise to investigate the feasibility of such a quasi-lumped distributed system. DD transmitter pads (4x6) and a 4x8 DDQ receiver pad have previously been built and these are used to make a short section of road. $V_{oc}$ and $I_{sc}$ measurements are taken to determine the VA profiles that show the feasibility of the system and also aid in validating a simulation model. Following that, full power profiles are simulated along and across a section of roadway to determine the shape of a zone in which the desired power can be coupled.

7.2. Power and tolerance requirements

The power required by a moving vehicle depends on the driving cycle, vehicle size and use, for example it may be used for either propulsion, charging the battery, limiting the peak power requirement from the battery to increase its life, or any combination of the three options. This results in numerous cases where different power transfer levels are required making determining the capacity of an IPT system a challenge. These issues have been addressed in [6, 7, 20] where the energy requirements of small (Honda Insight), medium (Chevrolet Impala) and large vehicles (Ford Explorer (SUV)) on different driving cycles have been investigated. The driving cycles are defined by the US Environmental Protection Society and are the Urban Dynamometer Driving Schedule (UDDS) and Highway Fuel Economy Test (HWFET). The study assumed power transfer levels of 20, 40 and 60kW were possible and the results show the proportion of powered to unpowered road varies according to the vehicle type and driving cycle. In summary, a range of 480km was attained by supplementing the power required by an EV via IPT therefore based on those results similar power levels (20-60kW) are considered in this thesis.
The IPT system needs to provide full power to the vehicle within a certain horizontal tolerance about the centre of the lane. Automatic guidance systems are not considered suitable therefore full power needs to be transferred across a width suitable for an average driver. Staying within a narrow zone requires constant control action causing driver fatigue but if the zone is made as wide as the lane the system will no longer be cost effective due to the excessive amount of material required (namely copper and ferrite).

In this thesis the horizontal tolerance zone is based on the minimum desirable lane width of 3.5m for highways and motorways in NZ. This value has been chosen according to the road’s annual average daily traffic and peak hour traffic volumes [130]. Assuming a medium sized car is 1.8m wide, there is a distance of 0.85m either side of the car before encroaching into the adjacent lane (or shoulder). Taking approximately half (0.85/2) of the distance to either lane leaves a horizontal tolerance of 0.4m either side of the centre of the lane or an absolute horizontal tolerance of 0.8m. Further work is needed to accurately determine country dependent horizontal tolerances. An air gap of 200mm is considered here.

7.3. Track based RPEV systems

Distributed systems naturally lend themselves to RPEV applications because they allow power transfer at any position along the track, these are considered in this subsection (7.3). The quasi-lumped distributed system using DD pads is presented in section 7.4.

7.3.1. Single phase tracks

Single phase tracks are not able to practically meet the tolerance requirements of RPEVs even with quadrature receivers. The receiver designs presented in [81, 82, 131] were 144mm wide and used with a vertical offset of 20mm above a track with conductors spaced 100mm apart. The receiver is very close to the conductors therefore can take advantage of the changes in the magnetic field direction as described in Chapter 2.3.3. This system cannot be practically scaled up to transfer 20-60kW because magnetic field strength decreases with the inverse square of the operational air gap \(1/d_z^2\). If the system is scaled up by a factor of 10 the receiver air gap is increased to a desirable 200mm and the track conductors are placed 1000mm apart. However, the track current must increase by a factor of 100 to maintain the same flux density and hence power transfer. The excessive copper requirement combined with the extremely large voltages required to obtain these high track currents make such systems impractical.
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The distance between the conductors in a single phase track must increase in proportion with the operational air gap. If the conductors are relatively close together for a given air gap there will be little net flux through the receiver. In the extreme case where the conductors are close together and the receiver is positioned far away, there is no coupling as there is effectively no net current relative to the receiver (illustrated by the contour in Chapter 2, Figure 2-9(b)).

7.3.2. Poly-phase tracks

Poly-phase tracks have been developed to improve the horizontal tolerance of IPT systems and these use conductors that are excited by currents that differ in phase due to electrical separation and geometrical layout. Single phase tracks produce a magnetic field that changes in magnitude as the field vectors only exist on a line at any given point about the track. As such, the field is considered to be stationary but time varying. However, poly-phase tracks produce fields that change in both magnitude and direction with time allowing quadrature receivers to be continuously excited allowing greater power transfer [81, 131]. Poly-phase tracks have a greater number of conductors for a given track width compared to single phase tracks resulting in a smoother power transfer profile; single phase tracks in comparison produce a peaky profile [103, 132].

7.3.3. Three phase tracks

Three-phase tracks are considered in this thesis and the most common topologies found in the literature are the bipolar and unipolar, as shown in Figure 7-1(a) and Figure 7-1(b) respectively [103]. The length of the track is referred to as $T_l$ and the width as $T_w$. The bipolar track is named as such because it comprises six conductors that leave the power supply and return to it therefore the receiver is exposed to both forward and return currents (with reference to the SS’ line). An advantage of the bipolar topology is that the return conductors effectively generate magnetic fields in the opposite direction due to opposite currents without the need for an additional power supply. This is shown by the vector diagram of Figure 7-1(b) and results in adjacent conductor currents that differ by 60°. This topology creates a travelling field across the width of the track in a similar manner to windings in a cage induction motor.

A major disadvantage of bipolar tracks is the unequal mutual inter-phase coupling due to differing loop areas as explained in [133]. For example, loops A and B share more common area that A and C. Given the mutual flux is proportional to area and the areas differ, there is
net coupling between phases that causes power to be fed back into the supply. This in turn causes the dc bus voltage to vary, which must be managed if control problems are to be avoided [133].

Two methods have been presented in [133] to reduce the inter-phase mutual inductance, these are conductor positioning and flux compensation. The position of the conductors can be adjusted to minimise the coupling between the inner phase (B in Figure 7-1(a)) and the outer phases (A and C). The distance between the outer phases naturally results in low coupling. This approach is shown to have a detrimental effect on the power transfer profile as the distances between phases is no longer uniform resulting in an inconsistent magnetic field shape.

Flux compensation involves intentionally introducing opposite mutual coupling between the equally spaced phases at the beginning of the track by passing the conductors through toroidal cores. This lumped mutual coupling is then cancelled by the distributed inter-phase cross coupling along the track. Flux compensation is not practical for large scale implantation because of the added cost and complexity of installing the toroidal cores. Further limitations of this topology are discussed in section 7.3.3.1 below.

The layout of a longitudinal unipolar three phase track is shown in Figure 7-1(c). The receiver is only exposed to forward currents from the power supply (shown in Figure 7-1(d)) as the conductors are star connected at the end. This topology still has the disadvantage of inter-phase coupling and there are half as many conductors for a given track width compared to a bipolar track. This results in a poor power transfer profile across the width of the track making this topology unsuitable for RPEV applications.
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7.3.3.1. Transverse unipolar track

RPEVs require high power levels as well as large horizontal tolerance and both the bipolar and longitudinal unipolar topologies are not practically scalable to meet these requirements. A track topology referred to as a transverse unipolar or just unipolar is presented in this thesis. The track comprises interleaved conductors wound across the width of the roadway as shown in Figure 7-2. The physical directional changes of the conductors mean that only three forward currents from the supply create three more effective currents as shown by the vector diagram in Figure 7-1(b). This creates the phase sequence shown at the bottom of Figure 7-2 (with reference to the SS’ direction). Unipolar tracks are referred to by the number of periods of the conductors and the track illustrated is a 2.5 period design, a 3 period track would need to be terminated on the opposite side.

The vehicle travels along the length of the unipolar track therefore the horizontal tolerance is determined by the width of the track as labelled in Figure 7-2. As discussed earlier, increasing the width of any of the conventional track topologies requires moving the conductors further apart resulting in a reduction in the magnetic flux density above the track. The power transfer zone for a bipolar track is shown by the hatched area in Figure 7-1(a) and its width cannot be increased practically limiting the use of this topology for RPEV applications.
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The power zone possible with a unipolar track is illustrated by the hatched area in Figure 7-2, and changes in its width are independent of the spacing between the conductors (pitch, \( p \)). The main advantage of this topology is a track that can be made as wide as desired with only a linear increase in copper loss – the track current does not need to be increased. A second advantage is there is almost zero mutual inductance between any two phases due to the repeating nature of the layout. This eliminates the need for positioning or flux compensation methods.

The unipolar track requires slightly more wire for a given length than a bipolar track due to the edges however the advantages outweigh the added cost. Wider tracks are more cost efficient in terms of coupling area for a given track length because the length of the curved edges is constant. In Figure 7-2 the edges have been made semi-circular to simplify the simulation model however in practice straight edges with small corners will reduce the length of wire required.

![Figure 7-2: Layout of a unipolar track with power zone shaded.](image)

The flux pattern about the track is similar to that of a linear induction motor and is a result of the constructive addition of the magnetic flux around each conductor (due to the phase sequence shown in Figure 7-2). However, due to the salient nature of the track the travelling magnetic dipoles vary in size, as shown in Figure 7-3. A similar effect is seen across the width of a bipolar track but the dipoles are more irregular due to the edge effect because there are only six conductors.

Travelling dipoles are advantageous as they reduce coupling nulls and also allow the use of quadrature receivers. Quadrature receivers use two coils that are able to couple with both horizontal and vertical flux components. As a dipole travels, the flux direction changes simultaneously exiting both coils resulting in greater coupled power. Quadrature coils in
systems with stationary fields have position dependent excitation – and usually only one coil is used to couple power in a given position.

A 3D simulation was done to show the magnetic dipoles produced by a unipolar track, the results, looking into the section line SS’ of Figure 7-2 are shown in Figure 7-3 (a) to (c) for phase angles of 0°, 30° and 60° respectively in reference phase A. As the currents between each of the conductors (+A, -C, +B, -A, +C, -B) differ by 60° only three plots are required to fully show the movement of the dipole through the distance of the track pitch. The dipoles expand and contract as they move along the length of the track, the minimum dipole width is less than the track pitch (as shown in Figure 7-3(a)) and the greatest width is almost twice the track pitch as shown in Figure 7-3(b). Regardless of the dipole width they are always $3p$ apart and this means that ideally the length of the receiver should be matched to the track to ensure the coupled power is maximised as investigated in section 7.3.4.

The unipolar track has similarities to a linear induction motor, however the speed of the travelling dipoles is not comparable to that of the vehicle. The speed can be calculated from $v=6fp$, where $f$ is the operational frequency of the system. Assuming practical values of 20kHz and a track pitch of 0.2m, the poles travel at 24km/s whereas a vehicle typically travels at 27m/s on a highway. Thus for simulation and design the receiver can be considered stationary.

The same phase sequence is maintained when the end of a unipolar three phase track is interleaved with the beginning of the next track. This is ideal for RPEVs as continuous power can be guaranteed for long section of road provided the power supplies are synchronised. There is a limit on the length of any track topology that can be driven with a power supply as the inductance increases requiring very large voltages to obtain the desired track current. However, investigating the length limit of tracks is beyond the scope of this thesis.

It is envisaged that such a unipolar track system will be completely modular with power supplies powering fixed lengths of track in the order of 10s of metres. This type of system is extremely reliable, if some sections get damaged or a few power supplies fail, the powered roadway (in the range of km) will still function at close to full capacity.
7.3.3.2. Built lab size track

In order to test the concept of the unipolar track topology and different receiver topologies, a section of track measuring 2.2m long by 0.795m wide with a pitch of 65mm was built as
shown in Figure 7-4. The track was made up of 5 periods of each phase to minimise unbalanced interphase coupling due to end effects. The test track was constructed with mains cable rather than Litz wire for ease of prototyping however this limited the maximum track current to 22.5A at 38.4kHz or it would heat excessively. The inductance values across the input terminals are well balanced, the values are: A-C 17.6μH, A-B 18.8μH and B-C 17.7μH.

**Figure 7-4: Built three phase unipolar track.**

### 7.3.4. Investigation of unipolar tracks

Several parameters affect the output power of a distributed system, however the ones considered to have the greatest effect are: receiver length and thickness, track pitch and the design of the ferrite structure underneath the track. The width of the track is not investigated as this can be increased if more horizontal tolerance is required.

There are numerous receiver topologies and considering each type is exhaustive and not necessary. A simple flat bar receiver is investigated here, and the optimisation is done both ways – first the track is matched to a receiver then a receiver is matched to a track. It is noted such a simple receiver will only couple power when the magnetic field is horizontal, illustrated by the labelled flux vectors in Figure 7-3(b), however the power output can be increased by using a quadrature receiver as presented in [81].

#### 7.3.4.1. Matching receiver length to track to pitch – varying P_l

The layout of the simulation model used to match the receiver or pick-up length (P_l) is shown in Figure 7-5 with the receiver placed in the middle of a track (both length and width wise). The width of the track is determined by the length of the straight midsection conductor (T_{cw}) and the pitch. Semicircular ends are used thus the total width is: \( T_w = T_{cw} + 3p \). Note only 1.5 periods are shown however a 5 period track was used in this simulation. The pitch is set to 65mm, air gap to 60mm and the receiver was 125mm wide (P_w) and 20mm thick.
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The changing distance between the cables and receiver results in undulations in the power profile as the receiver is swept along the length of the track. To account for this, the receiver is simulated in 5 positions, one at the centre (as shown in Figure 7-5) and two forward and two backward (a total movement range of 400mm) and the average coupled power is used. The length of the receiver has been varied from 100 to 400mm.

Figure 7-5: Simulation model of a flat bar receiver $I = 22.5A$ at 38.45kHz.

The graph of Figure 7-6 shows the uncompensated power of the receiver as it is made longer on a track with a pitch of 65mm. The coupled VA is at a maximum when the receiver is $\sim 4p$ or 260mm long. This result is consistent with the flux density vectors shown in Figure 7-3(b) where the outline of a matched receiver has been added. The distance between the dipole pairs is three times that of the track pitch and with 80% of the ferrite covered by the receiver coil, almost all of the available horizontal flux is captured. 80% of the optimal receiver length is $\sim 200$mm and $3p$ is 195mm confirming the maximum coupling condition. Increasing the length of the receiver ($P_1$) and hence coil length beyond $4p$ is undesirable. The majority of the flux in the ferrite of the matched receiver illustrated in Figure 7-3(b) is in one direction (to the left), but as a consequence of the extended length, a part of the receiving coil now also has field components that are in the opposite direction (to the right) and are out of phase. These additional out-of-phase field components produce an opposing voltage in the last few turns of the coil which results in a drop in VA (once the receiver length is greater than $4p$) as shown in Figure 7-6.

The ferrite utilisation efficiency is usually determined by dividing the $P_{su}$ by the volume of ferrite but given the width of the receiver is constant, the ferrite utilisation can be compared using VA/cm of receiver length. Designs that have a $P_1$ that is $\sim 2.8p$ are found to offer the best coupling for a given volume of ferrite however at present, designs that offer the highest coupled VA are preferred.
7.3.4.2. Matching receiver length to track to pitch – varying both $P_l$ and $p$.

The effect of increasing the track pitch for receivers of various lengths is shown in Figure 7-7 where $P_l$ has been varied from 140mm to 380mm and is normalised to the track pitch. The peak in coupled power occurs at receiver lengths that are 2.2 to 2.4 times the track pitch.

Clearly, larger receivers and larger pitches permit more power transfer, however this result appears to conflict with the previous simulation results in Figure 7-6. The power output of a 260mm long receiver on a 120mm pitch track is almost double that of the same receiver on a 65mm pitch track. The difference in coupling is due to flux cancellation and the fundamental
flux path height above the track. A track that has a pitch of 65mm has almost two dipoles for an optimal length receiver. When a dipole is at the centre of a receiver as shown in Figure 7-8(a) there will be no induced voltage. With a relatively long receiver above a relatively short pitched track there is more flux cancellation (for more percentage of the time period over which the flux vector sweeps 360°). This results in lower power transfer. If the same receiver is used on a track with a pitch of 120mm, there will be less cancellation in the coupling due to the greater distance between the dipoles, as shown in Figure 7-8(b), resulting in less cancellation because the distance to the next dipole is greater.

The above reasons along with the fundamental flux height above the track (described in the context of lumped systems in Chapter 4) influence power transfer. A higher fundamental flux path means there is more flux further away from the track resulting in better coupling. In a distributed system the height of the flux path is determined by the pitch of the track. Having a low pitch means the height of the flux path above the track is limited and this is approximated by $h$ in Figure 7-8(a). Note the height is shown under the track for convenience, in practice the flux about a track without ferrite is symmetrical. Doubling the pitch doubles the flux path height, as indicated in Figure 7-8(b). This allows significantly improved power output for a given track current and volume of receiver ferrite.

In practical systems, receivers that couple with all three of the phases are preferred to ensure load sharing as [114]. As discussed, each phase effectively is driven by two legs of a three phase inverter bridge (due to isolation transformers) and if the receiver couples to only two of the phases, these must be able to supply the full output. In consequence, components with 50% higher ratings are required. A receiver that is four times the track pitch in length will couple to all three phases resulting in proper load sharing among the legs in the inverter bridge, or between the three inverters if each phase is independently driven with an H-bridge.
7.3.4.3. Effect of pick-up width

The uncompensated power and ferrite utilisation are plotted against the width of a receiver (dimension $P_w$ in Figure 7-5) which is varied from 10mm to 300mm in Figure 7-9. Again the receiver is evaluated in five positions to average out the effect of varying distances from the conductors as discussed in section 7.4.3.1. The increase in $P_{su}$ is reasonably linear because the rate of increase in $M$ and $L_2$ is the same ($P_{su}$ is proportional to $M^2/L_2$) however the ferrite utilisation decreases with wider receivers. This is due to end effect which is a result of coupling due to flux from sections of track conductor not directly under the receiver getting attracted to the ferrite and passing through the coil [44]. As the receiver is made wider the fixed contribution to the coupling created by the end effect becomes less significant compared to the total coupling. The end effect is responsible for the initial large negative gradient of the VA per unit width of receiver curve.

Figure 7-8: Magnetic flux vectors illustrating nulls with optimal receivers on tracks with small and large pitches. The fundamental flux path heights are shown by $h$ ($I = 22.5A$)
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7.3.4.4. Scaling up system

The above simulations were done in the context of a small scale laboratory track however a practical RPEV system may need to deliver 20-60kW with an air gap of 150-200mm. Toward that aim, all subsequent simulations are done with a 175mm air gap and the track current is increased to 250A as industrial IPT systems use currents in the range of 100-300A.

In order to conceptualise the size of a system suitable for RPEVs, a simulation was done where the pitch of a track was varied from 50-380mm and the length of the 300mm wide receiver was adjusted to be four times the track pitch. This investigation was done with and without a solid 10mm thick ferrite sheet under the track. The sheet of ferrite maximises power transfer, however in the following section ferrite strips are used to optimise the power transferred for a given volume of track ferrite in the interests of cost efficiency.

The $P_{su}$ and $P_l/P_{su}$ against variation in the pitch of the track are shown in Figure 7-10. The increase in coupled power for a track with and without ferrite is relatively linear with pitches greater than 150mm. The low reluctance of the ferrite sheet under the track causes it to act like a flux mirror. In air, the flux lines are concentric circles around a conductor and the path is comparatively long resulting in a lower flux density 175mm above the track. When a ferrite sheet is added, the flux lines become semicircular and the path length is reduced significantly resulting in an increased flux density and hence improved power transfer.

Based on the results of Figure 7-10, a track pitch of 250mm has been selected with a receiver length of 1m. As the pitch of the track determines the width (shown in Figure 7-5) the width
becomes 1350mm. The $P_{su}$ produced by this arrangement is 11kVA and with a $Q$ of 6, the power output is 66kW, which meets the requirements of RPEVs.

![Figure 7-10: Increasing track pitch with a matched receiver, $I = 250$A at 38.4kHz and $\delta_z = 175$mm.](image)

7.3.4.5. Adding ferrite strips under the track

Simulations were done to determine the effect of placing continuous ferrite strips underneath and along the length of the track. The simulations were done with either 5 strips as shown in Figure 7-11(a) or 10 strips (not shown). The width of the strips was varied from a minimum of 5mm to a maximum (dependent on the number of strips) resulting in a solid sheet of ferrite. The results are shown in Figure 7-11(b). There are two families of curves, one for $P_{su}$ and the other for a volumetric comparison based on the volume of the track ferrite. The intersection with the $y$-axis is for a track without ferrite and the markers represent the annotated widths.

Using more strips enables higher power transfer for a given volume of ferrite as shown in Figure 7-11(b). The effect is more pronounced with narrow strips (volume of ferrite <6000cm$^3$) because each strip attracts flux from part of the conductor that is not directly above the strip (illustrated by the shaded area in Figure 7-11(a)) and a portion of this flux is then able to couple to the receiver. There are more such flux attracting areas with more strips and this increases the power coupled for a given volume of track ferrite. The effect reduces with wider strips as shown by both $P_{su}$ curves (of Figure 7-11(b)) converging on the point where a solid sheet of ferrite is used, in addition the $P_{su}$/volume ferrite decreases.
7.3.4.6. Using ferrite sheets of variable relative permeability

Constructing a track with embedded ferrite strips may be difficult in practice as the vibration can create small cracks that lower the relative permeability of the strip affecting power transfer. Also, it may be more desirable to use a layer of powdered ferrite material rather than discrete strips to lower cost and improve reliability. The ferrite powder may be sandwiched between plastic sheets before being cast into the concrete foundation for the road.

A simulation was done where the relative permeability of a solid sheet of material under a track was varied from 1 to 3000. The results are shown in Figure 7-12 – the $P_{su}$ is ~11 kVA with relative permeability of 2400 matching the results of Figure 7-10 and Figure 7-11.
curve of Figure 7-12 can be used to match the ferrite strip designs presented in Figure 7-11, for example, a solid sheet of material with a relative permeability of 400 achieves the same performance as 10 evenly spaced 20mm wide strips.

![Figure 7-12: Varying the relative permeability of a solid sheet under the track, $I = 250A$ at 38.4kHz and $\delta_z = 175mm$.](image)

7.3.4.7. Performance of the selected system

A conceptual IPT system for RPEVs is tested via simulation based on the previous results. A track pitch of 250mm and a conductor length ($T_{cw}$, dimension shown in Figure 7-5) of 800mm were selected resulting in a total track width of 1550mm. Eight equally spaced 20mm wide ferrite strips were placed under the track. The receiver is the same as that shown in Figure 7-11(a) and the operational air gap is 175mm.

A power profile along the length of a section of track with a maximum horizontal offset of 400mm is shown in Figure 7-13, noting the $y$-axis begins at 5kVA. The directions of movement are referenced to the axes shown in Figure 7-11(a). Only profiles with positive offsets in the $y$-axis are required as the track is effectively symmetrical. The ripple in the power profile when $\delta_y$ is relatively large is due to the curved end sections at the edge of the track. The peaks occur when the receiver is above two overlapping curves and the troughs occur when it is over one conductor however, the ripple is only 200VA and is not likely to affect the regulated output power.

The simulated system meets the 800mm horizontal tolerance requirement assuming a $Q$ of 6 is allowed. The minimum $P_{su}$ is 6.2kVA at the edge of the horizontal tolerance boundary as shown in Figure 7-13 and this enables a power transfer level of at least 37kW.
The results in the preceding sections show that a unipolar three-phase track is able to meet the power and tolerance requirements outlined in this thesis with a simple receiver topology. Using a quadrature receiver will not improve the horizontal tolerance but will enable ~50% more coupled power.

Figure 7-13: Performance of a 1m long receiver with lateral offset, $I = 250$A at 38.4kHz and $\delta_z = 175$mm.
7.4. Double D roadway system

As shown in the preceding sections, distributed systems are naturally suited to RPEV applications because they allow continuous power transfer within a given zone and are easily scalable. However, using such systems in an urban setting may be problematic due to the potential of exposing people to stray magnetic fields. Magnetic field exposure guidelines are introduced in Chapter 2 and discussed in the context of circular pads in Chapter 3 and Double D pads in Chapter 6. In essence, the general public should not be exposed to a magnetic flux density greater than 27\(\mu\)T. The inherently low coupling coefficients in a distributed system mean that large track currents (~100s of amps) are required to achieve the desired level of power transfer. The receiver only couples to a small section of the track and the remaining majority is likely to create a freely radiating magnetic field that makes complying with exposure guidelines difficult. This necessitates a different approach to the magnetic coupling used in RPEV systems.

In this section, the Double D couplers introduced in Chapter 4 are evaluated as part of a quasi-lumped distributed system. It is envisaged that DD transmitter pads will be appropriately spaced and buried along the centre of a lane enabling continuous power transfer to a DDQ receiver as outlined in [134]. In a discretised system, only the pads that are under a car will be activated thereby eliminating any additional stray magnetic fields along the road. Using DD transmitter pads in the road and garage offers the advantage of using a DDQ receiver that is suitable for both stationary and dynamic powering.

Charge zones in Chapter 6.8.3 show that the DD-DDQ system can provide continuous power transfer in the \(y\)-axis (width direction), this allows the transmitter pads to be placed side by side creating a continuous magnetic field. The DD pads are 410mm wide and as shown, 7kW can easily be transferred when the DDQ receiver is offset by 205mm in the \(y\)-axis. At this point the DDQ receiver is also effectively offset from an adjacent transmitter by 205mm (i.e. the receiver is positioned midway between two transmitter pads). When offset by 205mm, the receiver is likely to couple more than 7kW due to the constructive contribution from both pads, this allows the transmitter pads to be positioned in the road with a gap between them.

Unlike DD pads, circular pads are not polarised, however a set of identical transmitter and receiver pads will have a coupling null when horizontally offset by 38% of the pad diameter (as discussed in Chapter 3.3) therefore this topology is unsuitable for RPEVs. The coupled power will drop to zero before the receiver is above the next transmitter resulting in a pulsed
output and the average power delivered will be far lower than that possible with DD pads. In addition the power null means there is very little horizontal tolerance across a lane necessitating extremely high tracking accuracy, which is only possible with automation.

In this work the magnetic aspects of a quasi-lumped distributed system are investigated using 4x6 DD transmitter and the 4x8 DDQ receiver pads developed in Chapter 6. Larger pads are required for a practical RPEV system in order to transfer the necessary 20-60kW, however scaling up these power pads requires a comprehensive investigation that is far beyond the scope of this thesis. In addition, the spacing between transmitter pads needs to be optimised to ensure the power coupled is sufficient while minimising the number of pads, however this is also left as future work. The approach used here is to energise three transmitter pads and take $V_{oc}$ and $I_{sc}$ measurements to form a VA profile using a DDQ receiver, and then compare the results with a simulation model. Subsequently a full 3D simulation is run to obtain a complete power profile above the pads to determine the tolerance possible with the given pads.

7.4.1. System concepts

In practical roadways it is not usually advisable, or allowable (for safety reasons) to bury dc or mains cables directly beneath the roadway surface. As such, the envisaged technique for energising the DD transmitters is to use a double coupled system. The system consists of the following components:

- A buried three-phase line that runs alongside the road.
- Power supplies that produce a 20kHz track current.
- A double conductor single phase track buried in the centre of the lane that is typically 100-200m long.
- An intermediate coupler (which is similar to the ‘S’ pick-up described in Chapter 1 and [135]).
- A decoupling circuit.

The connection of these components is shown in Figure 7-14 for a section of roadway beginning with a power supply; the coupling system is modular therefore the ‘unit’ repeats to the right. Installation of such a system is simple – the units can be made offsite and then mechanically slotted into place as the road is constructed.
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The current in the transmitter pad is controlled by the decoupling circuit, these are able to communicate and synchronise the turning on and off of the transmitter pads to ensure the receiver on the vehicle sees a continuous magnetic field. The details of the decoupling circuit are not presented here as this thesis is only concerned about the magnetic coupling between multiple transmitters and a receiver, but it could encapsulate either simple tuning elements and an ac switch, or a bridge rectifier and H-bridge inverter.

![Diagram of a double coupled DD roadway system](image)

**Figure 7-14: Layout of a double coupled DD roadway system**

### 7.4.2. Measurement vs. simulation

Measurements are firstly taken to investigate the shape of the power profile across multiple transmitter pads and then these are compared to a simulation model to verify its accuracy. Here three transmitter pads are used to form a section of roadway (since using two pads will not produce accurate results because the end effect will affect the power profile). Using three pads allows a receiver to be placed above the centre pad – power transfer is mainly due to this pad but the surrounding energised pads will alter the shape of the magnetic field thereby enhancing power transfer as discussed in section 7.4.5.

The simulation model is shown in Figure 7-15(a) with the aluminium backing plates and pad cases removed. To maintain consistency with the axis definitions in Chapters 5 and 6, the lane is positioned in the y-axis and this is the direction of vehicle travel. The RPEV requires horizontal tolerance in the x-axis (across the length of the pads), and this particular pad orientation is chosen because the DDQ receiver has better tolerance in this direction due to
the contribution of the quadrature coil (presented in Chapter 6). This desirably results in an increased width of the power transfer zone and the length is simply increased by adding more double coupled transmitter pads along the lane.

The direction of the current in each transmitter pad is shown by a white arrow in Figure 7-15(a) and this must be consistent among all transmitter pads, if not, there will be no usable power output from the receiver when it is positioned between two transmitters. The magnetic field directions above the flux pipes (shown in Figure 4-13 in Chapter 4) will be opposite resulting in no net flux through the receiver coils causing a coupling null. The pad pitch \( P_p \) is the distance between the centres of the transmitter pads – this dimension is used because it is not affected by the dimension of the pad cases.

The built pads with a DDQ receiver on a test jig are shown in Figure 7-15(b). In this work the transmitter pads each are driven with 11.5A at 20kHz with individual power supplies that are synchronised by a master controller as described in [136]. A potential issue when running multiple transmitter pads is cross coupling, this means energy is fed from one pad to another and is described in the context of three phase tracks in section 7.3.3. Early experiments showed that there is very little cross coupling if the transmitter pads are slightly spaced apart as shown in Figure 7-15(b), and there is no cross coupling if a receiver is present as it extracts energy.
Measured and simulated $P_{su}$ profiles along the $y$-axis (with reference to Figure 7-15(a)) are shown in Figure 7-16(a) for transmitter pads positioned with a pitch of 525mm. Note the measured results have been scaled up by a factor of 4 ($23^2/11.5^2$) as the results were taken with each pad driven with 11.5A whereas the current in each pad was 23A in the simulation. The transmitter pads are illustrated on the bottom of the graph for size reference. The dashed line shows an average uncompensated power of 2kVA is available over a ~1.1m long section of roadway, this confirms the suitability of the DD topology for RPEV applications.

The results are very close with the largest error being ~5% when the receiver is over the last transmitter (Tx.3). The simulated results are exactly symmetrical about the middle transmitter pad therefore the deviation is most likely due to measurement error and may be due to the
wooden beams sagging or to a slight increase in the transmitter pad current as a result of component variation.

A $P_{su}$ profile for a receiver pad with a horizontal offset of 80mm in the $x$-axis is shown in Figure 7-16(b). Noting that the system is symmetrical about the $y$-axis thus an offset with $x=-80$mm (with respect to Figure 7-15(a)) will look identical.

![Figure 7-16](image)

Figure 7-16: (a) $P_{su}$ profile along centre of transmitter pads ($x=0$mm) and (b) $P_{su}$ profile with horizontal misalignment ($x=80$mm) for both profiles $I_1 = 23A$ at 20kHz for all transmitter pads and $\delta_z=200$mm.

The measured and simulated results are again in excellent agreement. The $P_{su}$ of the DD coils is now reduced and the quadrature coil is active. The output from the quadrature coil will continue to increase as the horizontal offset in the $x$-axis increases reaching a peak when $x=220$mm (as shown in the 3D power profiles in Chapter 6.8.2). The simulation model is
accurate which in turn allows further investigation of various pad pitches to evaluate the full power zone (with offset in both x and y-axes).

7.4.3. Measured VA with various pad combinations

Measured VA profiles along the x-axis (shown in Figure 7-15(a)) were taken to determine how various combinations of powered pads affect power transfer. The power profile of Figure 7-17(a) was taken above a single powered transmitter pad (Tx.1) with and without Tx. 2 present (it is positioned with a pitch of 525mm). These results indicate that an adjacent unpowered pad has very little effect on the power transferred and further illustrates the desirable lack of cross-coupling.

The VA profile of Figure 7-17(b) shows the three remaining cases of the powered pad combinations (noting the asymmetry of the on-off-on curve is due to either coupling or current variation in Tx. 3). Here a dashed line has been added at the peak of the off-on-off curve while the vertical lines show the centres of each of the three transmitter pads. The $P_{su}$ (when the receiver is at 0mm) is the highest when transmitter pads 1 and 2 are on (on-on-off curve). This is due to increased coupling that is a result of an indirect contribution from the flux around the return portions of the coils in Tx. 2 as discussed in section 7.4.5. This also causes the peak in $P_{su}$ (606 VA) to be slightly off centre as shown in Figure 7-17(b). When all three transmitter pads are powered the peak is exactly above the centre of Tx. 2 as shown in Figure 7-16(b) and the (appropriately scaled down) value is 666VA, which is higher than the case where only the first two transmitters are powered due to indirect flux contribution from Tx. 3.
7.4.4. Power profile with various pad pitches

Simulated VA profiles along the y-axis for transmitter pad pitches of 425, 475, 525 and 575mm are shown in Figure 7-18. The pitches correspond to a gap between the return portions of the transmitter pad coils (shown in Figure 7-15(a)) of 34, 84, 134 and 184mm. The pads were constructed with curved sides (shown in Figure 7-15(b)) to fit series compensation capacitors inside if needed therefore the power profiles with gaps of 34 and 84mm could not be measured. The complete range of simulated transmitter pad pitches results in a continuous power profile however the peak-peak variation is larger when the pads
are spaced further apart. The length of road that can be powered for a given pad pitch is shown by the vertical lines at the left of the graph in Figure 7-18 assuming the minimum $P_{su}$ is determined by the trough in the respective power profile. Transmitter pads placed close together allow more power to be transferred however the cost of the roadway increases as more pads are required.

Each of the VA profiles of Figure 7-18 has a main peak in $P_{su}$ when the receiver is centred over the middle transmitter pad and slightly lower peaks when the receiver is roughly centred over the first and last transmitters. The difference between the main peak and the lower peaks reduces as the pitch increases. Two examples are shown in Figure 7-18, one is for a pitch of 425mm and indicated by $\Delta_{425}$ and the other is for a pitch of 575mm and labelled as $\Delta_{575}$. This variation in the main and lower peaks is shown as a percentage change from the main peak in for all the simulated cases in the insert of Figure 7-18. This variation is due to favourable flux generation from adjacent pads and is discussed in the following section.

Figure 7-18: Power profiles ($\chi=0$) for various pad pitches, $I_1=23$A at 20kHz in all transmitter pads and $\delta_z=200$mm.

7.4.5. Flux in cross-sections of energised transmitter pads

The $NI$ generation is the same for all pad pitches simulated in Figure 7-18 however the peak $P_{su}$ is significantly higher with a pitch of 425mm compared to all other simulated pitches. This is due to flux about the return portions of the DD transmitter pad coils that are adjacent to each other, these are labelled in Figure 7-19(a). The magnetic flux density vectors looking into cross-section ‘A’ of Figure 7-19(a) are plotted for transmitter pads positioned at a pitch...
of 425mm in Figure 7-19(b). The direction of the currents in the parts of the coils in between the pads are opposing as shown by the superimposed dots and crosses. The flux about each section of coil (sliced by plane ‘A’) adds constructively creating a high flux density zone on the long edges of the pads shown by dashed lines in Figure 7-19(b). This high flux density ‘boundary’ forces more of the flux created by the flux pipe (shown in Chapter 4.3.7) to the receiver thus increasing power transfer. This is an indirect beneficial improvement in coupling but it diminishes as the transmitter pads are moved further apart.

There is no neighbouring pad at the outer edge of the third pad shown at the left of Figure 7-19(b), the lack of magnetic pressure causes the flux to collapse and in consequence the flux created by the flux pipe collapses similarly resulting in the 12% drop in $P_{su}$ when the receiver is directly above the last transmitter as shown in Figure 7-18 (for the 425mm curve at an offset along the road of 850mm).

Magnetic flux density vectors for transmitter pads positioned with a pitch of 575mm looking into cross-section ‘A’ of Figure 7-19(a) are plotted in Figure 7-19(c). Although the flux between the neighbouring coils adds constructively, the relatively large distance between them of 184mm results in a relatively low flux density (compared to Figure 7-19(b)). This forms ‘softer boundaries’ around the central transmitter pad and reduction in $P_{su}$ when the receiver is moved above either of the adjacent ones is only 6%.
7.4.6. Complete power profiles

The VA profiles taken in the preceding sections only considered displacement along the central $y$-axis without any horizontal offset (shown in Figure 7-15). However offsets that have both $x$ and $y$ components are required to fully determine the power transfer profile. Numerous simulations were run where a DDQ receiver pad was moved in 40mm increments
in both axes over a zone measuring 0.4m by 1.5m over three transmitter pads positioned with pitches of 425mm and 575mm. The power profiles for the individual DD and quadrature coils in a DDQ receiver are plotted as well as the total output (DD+Quadrature) in Figure 7-20(a)-(d). The graphs have been oriented to match the viewing direction of the pads shown in Figure 7-15.

The VA profiles of the DD and quadrature coils are shown in Figure 7-20(a) where the transmitter pads are positioned with a pitch of 425mm. There is no power output from the quadrature coil when \( x = 0 \)mm however it couples power with any offset. When the offset by in the \( x \)-axis is more than \( \sim 180 \)mm the quadrature coil couples more power than the DD coil, this is shown by the intersecting surfaces in Figure 7-20(a). The maximum simulated \( x \)-offset
is 200mm yet the quadrature coil reaches its peak coupling with an offset of 220mm, as such the peak is not visible in these profiles.

The VA profile obtained with the three transmitter pads positioned with a pitch of 575mm is shown in Figure 7-20(b) noting the axis only goes to 3kVA. The profile is peakier and the maximum \( P_{su} \) is lower than that obtained when the transmitter pads are placed with a \( P_p \) of 425mm, however spacing the transmitter pads further apart allows a greater length of the road to be energised for a given number of pads. This trade-off requires considerable investigation and is not undertaken here but left as future work. The total VA profiles (DD+Q) are plotted in Figure 7-20(c) and (d) for pad pitches of 425 and 575mm respectively. The contribution from the quadrature coil smoothes both profiles in the \( x \)-axis.

### 7.4.7. Charge zone characterisation

Charge zones were introduced in Chapter 6.8.3, and these 2D plots show the area in which the desired level of power transfer can be achieved given an operational air gap and a maximum receiver operational \( Q \). It is noted that in practical EV battery charging applications the output voltage of the power flow controller will be fixed (~300V for example). Therefore, to deliver the required power, the current \( (I_{sc}) \) of the receiver must be sufficiently high. The charge zones assume the current from the receiver is adequate and the number of turns on the DD and quadrature coils has been adjusted so the currents match. In practice doing this will affect the coupling slightly due to differing coil geometries but the work done here is a scoping exercise to prove the viability of the quasi-distributed lumped system.

Charge zones in this section assume a power level of 3.5kW and the transmitter pads are positioned with pitches of 425mm and 575mm. A maximum receiver operational \( Q \) of 6 is allowed for the DD and quadrature coils. The power zone possible with the DD coils of a DDQ receiver (\( P_p \) of 575mm) is the lighter shaded area shown in Figure 7-21(a). The shape of the zone reflects the position of the transmitter pads as greater tolerance across the lane is possible when the receiver is moved across any of the pads. Note the zone is slightly wider with extreme offset (\( x>150\text{mm} \)) above the middle pad (Tx. 2), this is caused by neighbouring pads containing the flux as described in section 7.4.5 above. The power zone possible with the quadrature coil only is shown in Figure 7-21(b). In practice the edges of the zones will be rounded but the artefacts in the simulated results are due to the 40mm increments used.
The power zone possible with the power output from the DD coils with transmitters positioned at a $P_p$ of 425mm is shown in Figure 7-21(c). The power zone is rectangular and approximately 300mm shorter than the charge zone of Figure 7-21(a). The power zone created by the quadrature coil is shown in Figure 7-21(d). With extreme $x$-offsets the closely positioned pads allow the desired power to be coupled using only the quadrature coil.

![Figure 7-21: 3.5kW charge zones (a) DD coils ($P_p=575mm$), (b) quadrature coil ($P_p=575mm$), (c) DD coils ($P_p=425mm$) and (d) quadrature coil ($P_p=425mm$), $I_t=23A$ at 20kHz in all transmitter pads and $\delta_z = 200mm$.](image)

The total output of a DDQ receiver is the sum of the power output from the DD and quadrature coils as described in Chapter 6.4. The full 3.5kW power zone possible with transmitter pads with a $P_p$ of 575mm is shown in Figure 7-22(a). The output of the DD coil makes up the largest proportion of the power zone (although the difference between the DD
and Q contributions is exacerbated because full analysis should consider variations to ±400mm rather than ±200mm). In the areas referred to by ‘DD or Quad.’ the output power can be supplied by either of the coils whereas in the areas indicated by ‘DD+Quad.’ the total output power is made up from contributions of both the DD and quadrature coils.

The power zone for transmitter pads positioned with a pitch of 425mm is shown in Figure 7-22(b). There is a large overlap where the output power can be furnished by either the DD or quadrature coils. For the two cases simulated, a pitch of 575mm is considered to be more optimal given the desired 3.5kW can be transferred over a longer zone permitting fewer transmitter pads per unit of road length. The width of the power transfer zones (across the lane) is at least 400mm, given the quadrature coil only reaches its peak in coupling when offset by 220mm in the x-axis, thus the power zones are likely to extend to ±300mm from centre.

A suggested approach for optimising the pitch is to keep moving the transmitter pads apart until holes appear in the full power zone (points at which full power cannot be delivered at all). However, due to the indirect coupling gain as a result of flux around neighbouring pads, this means that a full optimisation process will require at least five pads and only the results obtained when the receiver moves across the centre pad can be used. That is, a receiver must couple to a transmitter pad that is supported by a neighbouring pad that is also supported by a neighbouring pad to ensure the ‘boundaries’, shown in Figure 7-19(b), exist.

The above results show a system using the DD topology is a completely feasible approach, but significant future work is required to understand and optimise this system.
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7.5. Summary and Conclusion

Conventional track topologies used in distributed systems are not able to meet the challenges inherent in RPEV systems without electronic driver assistance – large amounts of power (20-60kW) need to be coupled over a 200mm air gap with large horizontal tolerance. In the first section of this chapter it was shown that a three phase bipolar track cannot practically meet the horizontal tolerance requirement of 800mm. In order to ‘widen’ the power transfer zone the conductors need to be spread and doing this results in a drastic drop in the magnetic flux density above the track. To offset this, the current needs to increase in proportion with the square of the change in conductor spacing. This means large currents are required and the cross-sectional area of the conductors needs to be increased resulting in additional cost.

A topology referred to as a unipolar three phase track was presented which consists of transversely laid conductors that are wound back and forth across the width of the road. This track effectively decouples the width of the power transfer zone from the current. A small scale 2.5m long track was made to investigate the topology and to validate a simulation model. A simple bar receiver with 80% of its length covered with the receiver coil was used and matched to the track pitch. A receiver that is ~4 times the track pitch was found to couple the most power and this desirably results in even phase loading.

A scaling simulation was done where the pitch of a track was increased with a matched receiver. Conductors spaced further apart permit higher flux paths and hence power transfer.
A pitch of 250mm was selected corresponding to an optimised receiver length of 1m, however it was found that ferrite was required underneath the track to increase the power transfer to levels acceptable for RPEVs. The $P_{su}$ without ferrite was 2.5kVA and this increased to 12kVA when a solid ferrite sheet was added under the track. However, such a solid sheet of ferrite is resource intensive and in consequence, ferrite strips were investigated as a suitable alternative. Using ferrite strips with $1/3^{rd}$ of the total volume of a solid sheet provided 80% of the $P_{su}$.

The simulation model of the unipolar track was made 1.6m wide and 5m long. It was able to provide a power output of at least 37kW with 800mm of horizontal tolerance at an air gap of 175mm assuming a maximum receiver operational $Q$ of 6 was allowed. The power can typically be boosted by 50% using a quadrature receiver enabling a power output of up to 55kW, which is at the upper limit of the desired RPEV power range in this thesis.

In the second part of this chapter a quasi-lumped distributed system using a string of DD transmitters and a DDQ receiver was presented. This topology was validated by measurements using a wooden jig with three 4x6 transmitters and a 4x8 receiver. Simulated flux plots were used to show that neighbouring transmitter pads aid in creating a high flux density boundary along the edges and this favourably increases power transfer but this effect diminishes as the transmitter pads are spaced further apart.

Full 3D power profiles were simulated with three transmitter pads positioned with pitches of 425mm and 525mm. Positioning the pads closer together results in a smoother power profile however the cost of a length roadway increases (assuming a fixed power transfer level). The profile produced with pads spaced further apart is peakier but the receiver controller will naturally operate to compensate for this variation by dynamically adjusting its circuit $Q$ over a wider range. As far as the output load is concerned, it sees an equally smooth power output irrespective of the separation providing the peaks in $Q$ can be tolerated.

The work on the quasi-lumped distributed system in this thesis was largely a scoping exercise and scaling the pads to achieve the required 20-60kW was not done. Determining the optimal pad pitch also requires a considerable amount of work however the basic results presented in this chapter form the foundation for future investigation.
Chapter 8. Conclusion and Future Work

8.1. General conclusion

Magnetic couplers have been investigated and developed in this thesis with the aim of creating cost effective designs that are suitable for both stationary and dynamic EV charging. In order to achieve this, the couplers should: provide good coupling with a large operational air gap for ground clearance, have large tolerance to horizontal misalignment for easy parking and be small and light for easy installation and to ensure high overall vehicle efficiency. Hitherto, couplers found in the literature and industry are not able to meet these demanding requirements in a cost effective way however, the Double D power pads developed in this thesis provide the closest approach. The work in this thesis has been published in two journal papers, four conference papers and four patents [100, 101, 106-108, 115, 116, 128, 129, 137].

The thesis began with a brief history of electric vehicles and the details of the SAE J1772 contact based charging standard. Plugs and sockets compliant with J1772 may have safety features such as a proximity detection pin and shrouding, but successful opportunity charging requires many plug-unplug cycles still creating a source of anxiety for the EV owner. Charging stations in public areas are prone to vandalism and the plug and cable will require maintenance due to frequent use. IPT charging systems are inherently more durable because the primary side can be cast in concrete so it is protected from the elements and ill-natured people. Because IPT is ideally suited to such tasks, it has been successfully employed in a variety of consumer and industrial and applications as presented in chapter one, with power transfer levels ranging from 1W to 30kW.

A general overview of IPT systems was presented in Chapter 2 where an emphasis was made on the importance of having power pads with high coupling coefficients and native quality factors. Generally, high coupling coefficients (~0.2) at the desired operational air gap (~200mm) ensure the necessary power transfer while a high native quality factor ($Q_L$) of the pads ensures good system efficiency. Magnetic field exposure guidelines produced by ICNIRP and ARPANSA were presented as well as the reasoning for the general public and occupational exposure limits.
8.1.1. Circular power pads

Circular power pads are the most common type of coupler found in the literature and industry but these typically use solid ferrite discs that are heavy and fragile. Improvements over existing couplers were made in Chapter 3 by purposefully positioning radial ferrite strips made up of discrete cores while still maintaining a suitably high \( k \) [100, 101]. Discrete cores allow the pads to flex and twist without damaging the ferrite structure, which is important for long term reliability when mounted on an EV. An optimisation procedure was applied where each variable of interest was swept from its minimum to maximum value given geometrical constraints using a 420mm diameter pad as the starting point. It was found that the strips should be made as long as possible and an average coil diameter 57\% of the pad diameter provided the highest \( k \). Despite this optimisation, the 420mm diameter pad under consideration was not able to provide enough coupling to transfer the desired 2kW over a 200mm air gap (it only allowed a 40mm air gap). Consequently, a pad sizing investigation was done by scaling this pad structure and it was found that pads 900mm in diameter were required to achieve a \( k \) of 0.2 with a 200mm air gap.

Coupling can be improved by adding extra ferrite but this needs to be done in the most effective manner, therefore a power to volume metric (in terms of VA coupled per cubic centimetre of ferrite) was used to compare several designs. Preferable options should have several narrow ferrite strips rather than a fewer wide ones. Pads 700mm in diameter were built with 12 strips each comprising three I93 cores (due to material availability). These had a coupling coefficient of \(~0.15\) with a 200mm air gap enabling 2kW of power transfer within a 130mm tolerance radius from the centre of the transmitter, however this is not sufficient for parking an EV without guidance. In the second part of Chapter 3 it was found, by simulation, that the aluminium shielding ring could be moved further away from the ends of the ferrite strips to improve coupling but this undesirably increases the pad diameter.

8.1.2. Polarised pads – developing the Double D

In Chapter 4 the shape of the fundamental flux paths above a circular pad were investigated. These largely remain the same regardless of optimisation – the flux path height is proportional to \( \frac{1}{4} \) of the pad diameter [106-108, 116]. In consequence, the pads must necessarily become very large to maintain the same coupling coefficient as the air gap is increased. In recognition of this, a radical shift was made and shaped bar pads were investigated as an alternative. These pads have a preferable fundamental flux path height
proportional to half of the length of a pad, which is twice that of an equivalent sized circular pad. Initially a coil was wound at the centre of a shaped bar and with a 200mm air gap, $k$ was measured to be only 0.07. The low coupling is due to a short reluctance path around the coil therefore most of the ferrite is redundant. The flux pipe was developed to address this and it consists of two coils connected magnetically in series and spaced appropriately apart along the length of the shaped bar. Flux pipe pads 586mm long were able to achieve a $k$ of ~0.2 with a 200mm air gap (vs. 0.15 for 700mm diameter circular pads).

Despite their high coupling factors, flux pipe pads cannot be used in practical EV charging applications because they are effectively flattened solenoids and produce equal amounts of flux out the front and back. This means loss increases dramatically if any shielding is used – about 500W is lost in the shielding on the back of the transmitter pad alone in a 2kW system. Auxiliary coils were added to the flux pipe to try and improve the front to back flux ratio however the best solution was to avoid producing the back flux in the first place.

In order to produce a single sided flux coupler, coplanar coils were placed on a ferrite sheet with the adjacent coil sections made straight and the remaining parts made semi-circular to minimise wire length. This results in two back to back ‘D’ shaped coils leading to the name Double D. The DD topology has a fundamental flux path height proportional to half of the length of the pad and notably adding an aluminium backing plate had little effect on $Q_L$, which indicates that the flux is largely constrained to the coupling side as desired. The coupling factor was found to be 0.215, however the intra-pad coupling factor was only 0.154 indicating the coupling was not due to the flux pipe effect as expected but due to the sections of coil on the perimeter of the pad.

Ferrite was removed from the outer edges of the pads during the design refinement process and this increased the intra-pad coupling coefficient to 0.47 and $k$ to 0.24 – this is achieved with less than half of the original amount of ferrite. The mode of coupling between the two ‘D’ shaped coils is unique to this topology. The flux in the ferrite is largely unidirectional and allows ferrite strips to be used to reduce cost as discussed in the next sub-section.

A second approach was used to verify the enhanced coupling provided by the DD topology by examining the flux distribution about these pads compared to a circular pad. The DD transmitter pad was shown to have more energy stored further away from the coupling face compared to a similar sized circular transmitter pad. As coupling systems are magnetically
symmetrical, an identical receiver is able to extract energy from further away confirming the significantly better coupling coefficients achieved.

8.1.3. Double D and optimisation

The Double D developed in Chapter 4 used a solid sheet of ferrite – this is similar to the circular couplers found in the literature that use a disc. The optimisation process applied to circular pads in Chapter 3 showed similar coupling can be achieved with ferrite strips resulting in a significant reduction in weight and hence cost. A similar process was applied to DD pads in Chapter 5 using a model comprising identical DD transmitter and receiver pads. Variables such as the spacing between the ferrite strips, the number and length of strips and number of turns were varied. A 2D design surface, created by varying both the length and width, enabled optimal designs to be selected by dividing the coupling coefficient by area. 4x6 pads were selected and these measure 770mmL by 500mmW and have a $k$ of 0.23 with a 200mm air gap.

8.1.4. Double D quadrature receiver

In Chapter 6 horizontal coupling profiles were taken in the two main axes (length and width) of DD pads and clearly showed the polarised nature of the topology. Performance was better in the width axis than the length axis for a given offset due to a power null that exists in the length axis at an offset of 35% of the pad length. A spatial quadrature coil was added to the secondary pad to compensate for this power null resulting in a significant improvement to tolerance to misalignment in the length axis [115, 116]. The quadrature coil is tuned and rectified independently and then added to the output of the DD rectifier – the rating of the secondary power flow controller does not need to be increased as the outputs are complementary.

8.1.5. Compliance with magnetic field exposure guidelines

DD pads have considerably better coupling coefficients than circular pads making a direct comparison of leakage magnetic field unfair. The 700mm circular pads evaluated in Chapter 3 coupled an uncompensated VA of 338VA with a horizontal offset of 130mm (200mm air gap) whereas the DD pads of Chapter 6 coupled 1255VA under the same conditions. In order to make a fair comparison, the current in the circular transmitter pad was increased accordingly so that the $P_{su}$ of both receivers is 1255VA. When compared under these conditions, the DD pads have slightly worse magnetic field leakage given they are only able
to meet the ARPANSA 27\(\mu\)T spot limit at a distance of 0.84m from the centre whereas circular pads meet the same limit at 0.62m. A key difference is the flux density in the operational air gap as it is able to heat metallic objects – the DD pads have one quarter the heating potential of circular pads thus are considered better overall.

8.1.6. Interoperability

An interoperability investigation was done in Chapter 6 where a DDQ receiver was operated over a circular transmitter pad. Charge zones show the area in which full power can be delivered (with a fixed air gap) assuming the maximum receiver operational \(Q\) is limited to 6 [115]. The DDQ receiver was shown to be completely interoperable – the quadrature coil provides the output power when the pads are vertically aligned (it is approximately circular) while the DD coils provide power when the receiver is offset. The charge zone produced is larger than that possible with a circle on circle system.

8.1.7. Using a three phase track for RPEVs

Power and horizontal tolerance requirements for RPEVs were introduced in Chapter 7 where small, medium and large cars were considered. A transverse unipolar three phase track ("unipolar") was investigated as a transmitter while using a simple flat bar receiver. The advantages of the unipolar topology are that there is balanced inter-phase coupling and the horizontal tolerance is decoupled from the input current (consequently the track can simply be stretched) [128, 129]. Other track topologies such as the bipolar and lateral unipolar require an increase in current to maintain the same flux density and hence power transfer resulting in higher standing loss. After verifying a simulation model with a laboratory scale prototype, a series of simulations were done where the track and receiver were scaled up to meet the 20-60kW as required by a moving EV. A receiver 1m long on a track with a pitch of 250mm was selected and is able to supply 37kW with 800mm of horizontal tolerance.

8.1.8. Using DDs to power RPEVs

In the second part of Chapter 7 a unique approach to realising an RPEV system was presented using only DD transmitters and a DDQ receiver to form a quasi-lumped distributed system. Unlike circular pads, DD pads have a continuous power profile when the receiver is offset in the width direction. This enables a string of transmitter pads to be laid side by side allowing continuous power transfer to a receiver. This concept was tested using three independently powered DD transmitter pads and a DDQ receiver. A simulation model was created and
verified using practical laboratory measurements. This enabled various charge zones to be investigated in the simulator at two transmitter pad pitches (425 and 575mm). A power level of 3.5kW was chosen and both pad pitches allowed a horizontal tolerance of at least 400mm however using a larger pitch enabled a 1.4m long section of roadway to be formed whereas the section was only 1.1m long if the smaller pitch was used. As shown the DDQ receiver is suitable for both stationary and dynamic charging and therefore is the first genuine universal receiver proposed. As the work in this thesis was only a scoping exercise to determine the feasibility of the quasi-lumped distributed system, significant further work is needed to scale up the pads for a practical RPEV system as described in section 8.3.3.

8.2. Suggestions for future work

This thesis mainly focussed on the development of polarised couplers and while some optimisation was done, there is still a considerable amount of future work required to fully understand the Double D topology. Future work is needed on pad sizing, magnetic field leakage attenuation and pad scaling and spacing for roadway applications as described below.

The pad size investigation and optimisation in this thesis was done with identical vertically aligned pads and the size of the charge zone (horizontal tolerance) was determined by the size of the pads. The 4x6 DD and 4x8 DDQ receiver pads produced an elliptical shaped charge zone (shown in Chapter 6) however for some applications (such as robots) circular zones may be preferable. This will require a change in the aspect ratio of the pad and this is suggested as future work.

The transmitter pad is likely to be installed in the ground for several years, especially if used in public parking areas therefore it must be future proof. In this context, power transfer and operational tolerance are the most important factors. Further investigation is needed to determine the optimal size of the transmitter pad since it may be required to charge both a small car with a small air gap at low power or a large SUV at high power over a much larger air gap. As a result, systems with asymmetric pad sizes need to be investigated. It is envisaged that a small car can use a small pad to naturally reduce coupling while a big car can use a large pad however the horizontal tolerance may be compromised.

Semiconductor switches are constantly improving and will soon enable operation at higher frequencies than the conventional 20 or 40kHz, noting an upper limit of 150kHz is imposed.
by magnetic field exposure guidelines [94]. Higher frequencies allow more power transfer but finer strands may be required in the Litz wire and all this warrants further investigation.

The operation of a new pad topology derived from the DD pad has been recognised by the author based on a similar design for a primary transmitter described in [138] and it is called a bipolar receiver [137]. The bipolar receiver works with a DD transmitter and provides comparable performance to a DDQ receiver while using less copper. The bipolar receiver requires extensive investigation but is left as future work due to time constraints however the concept is described in the subsequent sections.

8.2.1. Bipolar receiver pad

A bipolar receiver above a DD transmitter is shown in Figure 8-1(a). The bipolar receiver pad consists of two coils (A and B) that are magnetically decoupled and therefore can be individually tuned and rectified before being connected to a power flow controller [137]. The dimensions of the bipolar receiver pad are shown in Figure 8-1(b); this is referred to as a 6x9 receiver because it has 6 strips of ferrite made up of 9 of I93 cores. The coils sit on top of each other in the simulation model but in practice the overlapping outer edge of coil B will be fed under coil A (similar to the construction of the quadrature coil photographed in Chapter 6) therefore the pad is the same thickness as a DDQ receiver.

In this work the length and width of the bipolar receiver are fixed therefore the two dimensions that can be varied to reduce the mutual coupling between the coils are the mid-space and coil spacing in the x-axis (csX) as shown in Figure 8-1. The mid-space essentially describes the overlap between the two coils however, this needs to change as csX is adjusted (by moving the edge of the coils inwards so that the ferrite strips protrude).
The intra-pad mutual coupling (the coupling between coil A and B) is measured by exciting one coil and measuring the induced voltage in the other, which is given by: \( V_{\text{induced, } A} = j\omega M I_B \). When little or no voltage is induced the mutual coupling is close to zero. The troughs in the graph of Figure 8-1 show the mid-spaces that result in low mutual coupling for \( csX \) values ranging from 80-160mm. When the coils are brought in they become smaller therefore the mid-space reduces accordingly.
A 6x7 DD transmitter pad was chosen to excite a bipolar receiver because it is roughly the same physical size as shown in Figure 8-1(a). Horizontal VA profiles were taken in the x-axis for all valid csX and mid-space combinations. The total VA output of the receiver is shown in Figure 8-3 and is the sum of coil A and coil B.

The VA profile of the pad with a csX of 160mm is the smoothest and there is a slight improvement in the VA coupled with horizontal offsets greater than 230mm. However, a csX of 80mm provides the best overall performance and this is expected given the coils are the largest and therefore couple the most flux. A csX of 80mm and a mid-space of 115mm are used for the remaining simulations in this thesis.

The operation of the bipolar receiver seems counterintuitive, therefore to aid the explanation the VA of the individual coils as well as the summed output in both the x and y-axes are plotted in Figure 8-4. The inset shows the position of the coils looking down on the system.
shown in Figure 8-1(a). The bipolar receiver is symmetrical about both the \(x\) and \(y\)-axes therefore the output of both coils is the same when the pads are vertically aligned.

As the offset is increased in the \(x\)-axis, the output of coil A reduces because the flux produced by the DD transmitter enters and exits the inner area of the coil – when \(x=180\text{mm}\) there is no net flux through the coil resulting in a coupling null as shown in Figure 8-4. The case is the opposite for coil B, it couples more flux as the offset is increased in the \(x\)-axis because the ferrite in the receiver ensures the flux paths produced by the DD remain constant. This is discussed in the context of a DDQ receiver in Chapter 6. The output of coil A and B are complementary and when added result in a smooth power profile in the \(x\)-axis (labelled as \(A+B(x)\)) shown in Figure 8-4.

Both of the bipolar receiver coils couple the same amount of flux when the offset is in the \(y\)-axis therefore have the same VA profiles. As such the \(P_{su}\) profiles are identical and are labelled as A or B in the graph of Figure 8-4. The profiles showing the total VA output in both the \(x\) and \(y\)-axes are labelled as \(A+B(x)\) and \(A+B(y)\) respectively in Figure 8-4. These VA profiles show the bipolar receiver has a favourably symmetric VA profile that is similar to a circular pad (although with far better coupling). The maximum simulated horizontal offset is 300mm and it appears the output of coil A is increasing rapidly indicating better tolerance however further analysis is left as future work.

![Figure 8-4: Output of the individual and summed coil in both axes (\(\delta_z=250\text{mm}\) and \(I_1=20\text{A}\) at 20kHz).]
8.2.1.1. Comparison of DDQ and bipolar receivers

A 6x7 DD transmitter is used to excite both the bipolar and DDQ receivers and the 3D VA profiles for coils A and B of a bipolar receiver are shown in Figure 8-5(a) while VA profiles for an equivalent sized DDQ (6x9) receiver are shown in Figure 8-5(b) (noting that here the vertical scale goes to 3kVA). The VA coupled by the individual coils in a bipolar pad are equivalent in magnitude and have rotational symmetry about the z-axis (180°). Both of the coils contribute power to the load with horizontal offsets less than approximately 100mm whereas with a DDQ receiver the DD coil provides most of the output in that scenario – the quadrature coil is redundant and only becomes useful with extreme offset in the x-axis.

A bipolar receiver may be more desirable in over-coupled situations; this occurs when the receiver gets too close to the transmitter and the coupled power is too high. The coils are mutually decoupled therefore one can be “switched off” thereby reducing power to the load – this makes the receiver very versatile. Such a mode of operation is not possible with the DDQ receiver because the quadrature coil can only be used if there is enough offset in the x-axis. The vast difference in VA coupled between the DD and quadrature coils (as shown in Figure 8-5(b)) means that the controller for a DDQ receiver will need to be significantly overrated to deal with over-coupling. Further investigation is required on this mode of operation but is left as future work.

The overall power coupled by the bipolar and DDQ receivers is the sum of the individual outputs of the coils and is shown in Figure 8-6(a) and (b) respectively. The profiles look...
largely similar however there is a slightly quicker drop off in VA coupled by the bipolar in the $x$-axis.

![Chart 8-6](image)

**Figure 8-6:** Total output of (a) bipolar receiver and (b) DDQ receiver ($\delta_x=250\text{mm}$ and $I_t=20\text{A}$ at 20kHz).

Charge zones illustrate the area in which full power can be delivered assuming a limited receiver operational $Q$ (for each coil). In this case the desired power is 5kW and $Q \leq 6$. The charge zones for the bipolar and DDQ are shown in Figure 8-7. The bipolar receiver provides a similar charge zone to the DDQ receiver, however the latter is expected to have greater tolerance to offset in the extreme $x$-axis. Simulation with offsets greater than $\pm 300\text{mm}$ are needed but again these are left as future work.

The DDQ receiver requires 87.4m Litz wire whereas the bipolar requires only 73.8m of the same wire. The bipolar receiver requires 16% less copper than a DDQ and is therefore a cheaper topology (as the ferrite needed is the same) and it has a unique mode of operation that allows it to work in over-coupled situations. It is therefore considered a practical alternative to the DDQ but more investigation is needed to fully determine the advantages and disadvantages of both topologies.
8.2.2. Magnetic field attenuation techniques

A system using DD pads is able to comply with magnetic field exposure guidelines when transferring 2kW as shown in Chapter 6 – in fact field levels are comparable to those produced by a circular pad operating under the same conditions. However, if larger pads are used to higher power levels (~10kW) compliance may become a problem and field attenuation techniques may need to be considered. Ferro-/ferromagnetic or conductive material can be used to shunt or absorb some of the leakage flux. The effect of iron (a typical EV chassis) can only be determined by measurement as FEM software cannot resolve such small elements (to account for skin effect) in physically large models.

Initial 2D simulations were done to investigate the effect of extending the ferrite in the receiver and adding aluminium shielding. This is considered to be simulated accurately because the FEM software treats the 5mm thick sheet as an ideal layer (no skin depth) and in practice at 20kHz there are ~12 skin depths therefore the aluminium is effectively impermeable to flux in this case.

The first simulation was done with a transmitter and receiver made up of strips comprising eight I93 cores. The magnetic flux density contour is shown in Figure 8-8(a) where the receiver is open circuit and has no aluminium backing plate. The view of the pads is in the xz plane with reference to the insert. The circle has a radius of 1m and the red triangle indicates...
the start of the circle and should be ignored. The flux density on the illustrated circle is ~10μT and reasonably consistent around the entire circumference.

An I93 core was added to either end of the receiver (10 cores in total) and the results are shown in Figure 8-8. The additional cores provide favourable flux paths and help contain the leakage at the sides of the operational air gap and above the receiver. There is still a significant amount of flux under the transmitter pad but this is unlikely to present a problem if the pad is buried – this is a preferable option as the aim is to reduce the flux out the sides of the pads.

The effect of an aluminium backing plate extended by 100mm either side of the receiver is shown in Figure 8-8(c). This results in a significant reduction in the leakage flux at the sides of the pads. The line segment $l$ in Figure 8-8(b) and (c) are the same length and this indicates the aluminium sheet desirably pushes flux further below the transmitter pad. These simulations have been done in 2D due to time constraints however it is expected that extending the ferrite will have a beneficial effect but this requires further investigation with 3D models.
Figure 8-8: Investigating leakage flux with 2D simulations in the $zx$ plane (a) Identical Tx. and Rx., (b) ferrite in Rx. extended and (c) aluminium shield added ($I_1=23A/turn$, 20 turns/coil and 2 coils/Tx.)
8.2.3. DD roadway pad sizing

The feasibility of using DD transmitter pads and a DDQ receiver pad for RPEVs was evaluated in Chapter 7 however the power transfer level was only 3.5kW. Significantly higher power levels with greater tolerance to horizontal misalignment are required for practical applications. The pads need to be scaled up and appropriately spaced to minimise resource use. It may be possible to use pads with different aspect ratios – the return potion of the coils (as shown in Chapter 4) remain the same length regardless of pad width therefore wider pads contain a greater proportion of ‘active’ copper. Doing this could also have the desirable effect of reducing the undulations in the power profile in the direction of travel.

A similar pad sizing investigation needs to be done with the bipolar receiver for both stationary and dynamic charging applications.

In conclusion, this thesis has formed a solid foundation for further work on these single-sided flux couplers.
References


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