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Design of a Variable-Phase Contactless Energy Transfer Platform using Air-Cored Planar Inductor Technology

PROEFSCHRIFT

ter verkrijging van de graad van doctor aan de Technische Universiteit Eindhoven, op gezag van de rector magnificus, prof.dr.ir. C.J. van Duijn, voor een commissie aangewezen door het College voor Promoties in het openbaar te verdedigen op maandag 15 maart 2010 om 16.00 uur

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Abstract

Design of a Variable-Phase Contactless Energy Transfer Platform using Air-Cored Planar Inductor Technology

Contactless Energy Transfer (CET) describes the process in which electrical energy is transferred among two or more galvanically isolated electrical circuits or devices by means of magnetic induction (magnetic energy). The potential applications can range from the transfer of energy between low power home and office devices to high power industrial systems. Medical, marine, and other applications where physical electrical contact might be dangerous, impossible or at the very least problematic, are all prospective candidates for using CET.

The work in this thesis mainly concentrates on fundamental concepts of a CET system that consists of multiple PCB inductors arranged to form a platform surface. When a mobile electronic device, embedded with a similar “power receiving” inductor, is placed on the platform surface, power is transferred from the platform to the mobile device through magnetic induction. Small low profile air-cored PCB inductors are used, since they can easily and cheaply be produced, and due to the small size and low profile easily be embedded in different devices without adding much weight or volume to the existing designs. As such, the CET platform attempts to remove the different cables, plugs and adaptors used to power and recharge various mobile electronic devices.

Firstly, a library of analytical and numerical methods and procedures for modeling planar inductors constructed as copper tracks on printed circuit boards are presented. Here methods for calculating the DC- and AC-resistances, mutual inductances, self-inductances, magnetic field intensity, and parasitic capacitances are developed. Using these models, the maximum excitation frequency of the inductors are estimated in respect to their self-resonating frequencies. The purpose of this “library of techniques” is not only to model the inductors used in this work, but to act as a single repository for information on modeling similar planar PCB inductors for various future CET applications. The developed models are verified through measurements and FEM simulations, and they all show good agreement.

Secondly, the development of the theory and the synthesis of a variable-phase CET system is performed. The platform is designed, based on user requirements, to transfer approximately 8 W of power from a cluster of three primary windings to a load connected to the secondary winding, over a 5 mm air gap. In this way about 5 W is available to load devices connected to the secondary winding and circuit after rectification and down conversion. The primary inductor platform consists of a matrix of hexagon spiral windings with radii of 12.5 mm, while the secondary windings have radii of 20 mm. The system operates at a frequency of 2.777 MHz, and is designed to induce a minimum voltage of 10 V (RMS) in the secondary winding, in order to ensure that at least 5 V - 6 V DC is available to the load devices. The CET platform is implemented
in a prototype system. This high-frequency power electronic system is modularly designed and consists of several subsystems including a FPGA controller, quadratic buck converters, high-frequency half-bridge MOSFET drivers, half-bridge inverters, voltage and current estimators, as well as winding commutation circuits.

Thirdly, in order for the CET system to locate the valid devices placed on the platform, an innovative load detection scheme is developed. The load detection scheme “scans” the primary windings while estimating their equivalent impedances through measurements. By comparing and detecting differences between the measurements and previously calibrated values, the system can find and locate the valid devices. Furthermore, this process has the added advantage of being able to distinguish between valid CET devices, metallic materials, as well as soft-magnetic materials placed on the platform surface. Objects like, soft-drink cans, lighters, pens, coins, and certain soft-magnet containing devices, which might have been accidently placed on the platform, can thus be detected, and the activation of the windings closest to it, avoided.

Fourthly, the CET platform investigates the efficiency of a novel stray magnetic field shielding methodology using destructive wave interference. In contrast to the more conventional methods of shielding using magnetic- and conductive materials, this method uses the amplitudes and phases of the currents in the primary windings, to alter the distribution of the magnetic field above the windings, resulting in a reduction of the stray magnetic field. The system is designed to operate in one of three modes: single-phase, three-phase and variable-phase mode. During single-phase mode, all activated primary windings are excited with in-phase currents. Here, no additional reduction of the magnetic field is expected. During three-phase mode, the primary windings are excited with currents of equal amplitude but with 120° phase shifts. Here, the magnetic fields of the three primary windings undergo destructive wave interference, attenuating the stray magnetic field. During power transfer, however, the secondary winding current still contributes to the field in a constructive manner. The variable-phase operational mode attempts to further reduce the stray field by taking into account the current in the secondary winding during power transfer. Here, the windings inside the activated winding cluster are excited with precalculated currents of different amplitudes and phases, which will give the minimum stray magnetic field values for given secondary winding placements and power transfer levels. The simulation results show that both the three-phase and variable-phase modes are able to reduce the stray magnetic fields (for the same amount of power transferred to the load resistor) compared to the single-phase mode. Attenuation results of up to 12 dB are calculated.

Finally, the power transfer capability, the operation of the load detection scheme, and the stray magnetic field shielding through destructive wave interferences are all verified through measurements performed on the prototype system. Furthermore, the prototype system is able to transfer 5 W of power (DC) to a load connected to the secondary circuit, in all three operational modes, and all measurement results show good agreement with calculated values.
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1

Introduction

“Education is the most powerful weapon which you can use to change the world.”
- Nelson Mandela

1.1 Energy and Energy Transfer

The word energy, originates from the Greek word ἐνέργεια - energeia, meaning “activity or operation” and ἐνεργός - energos, meaning “active or working”. Energy is a scalar physical quantity that describes the amount of work that could be performed by a force. Energy is an attribute of objects and systems that is subject to a conservation law. Several different forms of energy exist to explain all known natural phenomena. These forms include (but are not limited to) kinetic, potential, thermal, gravitational, sound, light, elastic, and electromagnetic energy [1].

Without a definite reference frame, the term contactless energy transfer can be used to describe the transfer of energy between various energy states and between different objects. The word contactless infers some sort of remote action, so that the transition between energy states would occur over a certain physical distance. A few examples of such CET systems are the transformation of light energy from the sun into heat energy which warms up the earth (Fig. 1.1 (a)), as well as the transfer of sound energy emitted from an audio speaker into movement of the eardrum (kinetic energy) inside the human ear [2] (Fig. 1.1 (b)).

In the context of electrical systems, the term CET takes a more definite form. A few examples of electrical CET systems are shown in Fig. 1.2. These include: the conversion of light and heat energy into electrical energy inside a solar cell [3], the conversion of electrical energy into light energy in a light bulb [4],
the transformation of electrical energy into electromagnetic energy in antennas [5], as well as the conversion of sound energy into electrical energy in acoustic microphones [6].

![Figure 1.1](image1.png)

**Figure 1.1:** (a) Infra-red light energy emitted from the sun is converted into heat energy which heats up the earth. (b) Sounds in the form of pressure waves are transformed into kinetic energy inside the human ear.

![Figure 1.2](image2.png)

**Figure 1.2:** Examples of electrical CET systems. (a) A solar cell, (b) a light bulb, (c) an electromagnetic antenna, and (d) an acoustic microphone.

### 1.2 Contactless (Inductive) Energy Transfer

Contactless energy transfer, as used in this thesis, refers to a specific form of electrical CET, and describes the process in which electrical energy is transferred among two or more galvanically isolated electrical circuits by means of magnetic induction (magnetic energy). Due to the ambiguity in the term CET, there exist many different terms in literature to describe this action. Some of these include, contactless inductive energy transfer (CIET), contactless power transfer (CPT), inductive power transfer (IPT), inductive energy transfer (IET), wireless power transfer (WPT), wireless energy transfer (WET), as well as the recently coined term Witricity [7, 8] (from wireless and electricity).
The potential applications for such a technology can range from the transfer of energy between low power home and office devices to high power industrial applications. Medical, marine, and other applications where physical electrical contact might be dangerous, impossible or at the very least problematic, are all prospective candidates for using CET.

One of perhaps the most popular examples of commercially available CET systems today, is the range of Sonicare electrical toothbrushes from Philips [9]. The Sonicare toothbrushes are fitted with small batteries, which need to be periodically recharged as the toothbrushes are used. Since toothbrushes are mostly used in bathrooms, in the vicinity of water, open electrical contacts used to recharge them, could cause short circuits or electrical shocks when they become wet. This could pose a possible danger to the user of the toothbrush, and can also damage the electronic circuitry inside the toothbrush. To solve this problem, this range of toothbrushes uses CET to transfer power from the charging station to the toothbrush, thus avoiding open electrical contacts and the dangers they pose. Figure 1.3 shows the open CET toothbrush charger as well as a charger which has a toothbrush inside it. Here, the toothbrush has a designated position inside the charger and can only be recharged when the toothbrush is docked inside it. The charging station transfers approximately 16 W of power to the toothbrush, and is a good example of a low power CET system [10].

An example of a high power CET system, is the multi-level contactless motion system presented in [11]. In this system, a matrix of primary coils is used to levitate and move a magnet-plated platform. Essentially, the platform flies above the primary coil matrix with a theoretically unlimited stroke in the planar directions. Levitation and movement are achieved through the repulsive forces acting between currents in the primary coils, and the magnets embedded in the floating platform. Systems like these are used as high precision machines in lithography, wire bonding, as well as pick and place machines. Here the physical isolation of the platform from the “real world” can reduce the disturbances and vibrations on the platform, allowing higher levels of accuracy compared to existing non-
levitated systems. In order to operate machines placed on the floating platform, power is transferred from the primary coils, to secondary “power receiving” coils mounted on the platform. In this way the system is kept “contactless”. Here, power in the range of several hundreds of watt is transferred to the platform. Due to the use of multiple primary coils, this system has a high degree of freedom regarding the positioning of the platform and the transfer of power. Figure 1.4 shows an impression of this system.

Figure 1.4: A multi-level contactless motion system with integrated CET [11].

From these two, and various more examples, it is seen that CET can be used as a viable alternative to the use of regular plugs, wires, and cables, and the limitations and problems they might pose.

1.3 CET Platforms for Charging Mobile Devices

Electronic devices like mobile phones, PDA’s, multimedia- and music players, laptops and many more, are used daily by countless amounts of people all around the world. These devices are fitted with batteries which allow them to operate independently and without drawing power continuously from the power utility network. These devices, however, need to be periodically recharged, since their batteries can only store a finite amount of energy. Most of these devices operate with relatively low DC voltage levels (typically 5 V - 12 V) compared to the high utility voltage of 240 V AC (120 V AC in the USA), and almost always require an AC-to-DC converter (also called a charger) to recharge their batteries.

Currently, there are millions, if not billions of different cellular phone sets in use all over the world [12]. There are at least a dozen different cell phone manufacturers, and numerous different versions and varieties of phones available. Many of these phones have unique chargers with specific power ratings and unique connectors suited only for the phone it is sold with. The same can be said
about portable music players, PDA’s, laptops, and various other mobile electronic devices on the market today. Finally, if we add all these chargers together, the total amount of chargers on this planet, today, is really mind-blowingly huge!

In domestic and office environments, a person can own and operate various different electronic mobile devices. Having to deal with all the chargers, plugs, and wires however, can be quite bothersome and irritating. Overloaded wall sockets, connected with numerous chargers can also be dangerous, or at the very least, an eyesore. From a consumer point of view, charging these devices with only one universal charger would be an improvement and would greatly simplify the recharging process. Simultaneous charging of two or more devices, however, would still require multiple chargers. An even better solution would be, if a charger could recharge multiple devices at the same time, without even plugging them into sockets, but by simply placing them close to the charger itself.

This is what CET platforms aim to achieve: To effortlessly power and recharge multiple different consumer electronic devices without using any plugs, chargers, or connectors, but by transferring energy through magnetic induction. Various CET charging platforms for these applications have been proposed in literature.

In [13, 14, 15], different CET mobile phone-battery charging platforms are proposed. In these systems, a single circular spiral winding inside a charging platform transfers power to a similar winding embedded inside a mobile phone, through inductive coupling. The phone has a designated position on top of the platform, so that the windings have perfect overlap for maximum coupling. The design does take into account small variations in air gap sizes and winding misalignments, but due to the use of only single primary and secondary windings, it does not have the ability to deal with large lateral displacements. The phone is thus restricted to a certain area wherein it can be charged, and the charger can only charge one mobile phone at a time. Figure 1.5 (a) shows a graphical representation of a mobile phone with a single winding, and its relative placement to the charging platform. One way to improve the phone positioning restriction, is to increase the primary winding radius, as shown in Fig. 1.5 (b). This would also make enough space for charging multiple load devices. This type of solution has not been investigated much in literature, possibly because it would prove inefficient when charging only single devices, due to the increased losses in the larger primary winding. The magnetic field would also cover a larger area, where it could undesirably couple with other electronic devices, and increase the possibility of exposure to people.

A popular choice in literature for increasing the freedom of CET device placement and for allowing multiple devices to be charged at the same time, is by using CET platforms with multiple primary windings.

In [16, 17], a multi-winding CET battery charging platform is presented. In this system, multiple hexagon spiral windings produced on printed circuit boards are used to transfer power from the platform to a receiving winding embedded inside a cellular phone. The primary windings are arranged in a special three-layer hexagon matrix configuration which produces a uniform magnetic field intensity above its surface. Due to the unique shape of the magnetic field intensity,
receiver windings placed above and in parallel to the charging platform, experience an unvarying induced voltage and stable power transfer even with relative movement in the plane. The placement of the cellular phone for charging is thus not restricted to any certain position or orientation, but can be placed randomly on the platform. The system thus has a higher degree of freedom, and possible increased user comfort, in regards to the positioning and the amount of the cellular phones, placed on the platform. However, one drawback of this system is that to maintain the uniform magnetic field, all the primary windings should be energized constantly. This dramatically reduces the platform’s efficiency, since all the activated windings even the ones not coupled with the mobile phone, will dissipate power due to their winding resistances. Furthermore, all the windings continuously generate magnetic fields, also at those positions where no devices are present. These stray magnetic fields could undesirably couple with other electronic devices in the vicinity, possibly causing them damage or interfering with their normal operations. Figure 1.6 shows a graphical representation of the three-layer hexagon spiral matrix CET charging platform.

Moreover, in [18], a desk with a cord-free power supply is presented. This system uses multiple circular coils embedded into a desk to transfer power to devices placed on top of the table surface. The primary coils can be individually activated, and only a few primary windings closest to the load device are activated at a time. This system has a high degree of freedom regarding the placement of the secondary devices. Unlike [16, 17], however, this system does not make used of a special magnetic field profile, but makes use of the existing inductance path between the primary and secondary coils, to send signals that indicate the presence of a secondary load device. This method however, requires extra circuits and intelligence on the secondary device in order to work.

Figure 1.5: (a) A graphical representation of a CET platform with single primary and secondary windings. (b) A large primary winding could allow multiple secondary CET devices to be charged.
1.4 Commercially Available CET Charging Platforms

Commercial mobile electronic manufacturing companies constantly seek new and innovating technologies to incorporate into their devices in order to give their products advantages over those of their concurrents. Mobile electronic device charging though the use of CET is one of the technologies which companies hope, will not only make their products more user friendly by removing the need for cables and power adaptors, but also more appealing to buyers in the ever increasing and demanding consumer electronics market.

Recently, a handful of innovating companies have started researching, producing, and selling CET charging platforms for charging existing and new mobile devices, through means of magnetic induction. However, due to the lack of existing standardization in this field, it resulted in various different CET charging devices, all claiming to be better than the other. Moreover, companies keep their intellectual property and technologies secret, which makes it very difficult to compare their systems. The products from these companies are all very similar, consisting of a primary charging platform and small secondary receiver devices, which can be plugged into the mobile devices, for charging. A few companies have gone as far as retrofitting some available mobile device casings with secondary windings, and also created sleeve-like transparent enclosures, with integrated coils and circuitry. A few of these companies are:

- Powermat [19],
- Ecoupled (formerly known as Slashpower) [20],
- Wildcharge [21],
- WiPower [22],
• Mojo Mobility [23],
• and PowerCast [24].

A list of additional companies that specialize in CET charging platforms and application can also be found online [25].

One example of an existing integrated CET charger, is the Palm Touchstone CET charging platform [26], which can be used to recharge various PDA devices produced by the same company. Figure 1.7 (a) shows the Palm Touchstone CET charging device, while Fig. 1.7 (b) shows a Palm Pre PDA docked to the CET charger. Furthermore, Fig. 1.7 (c) shows a Wildcharge CET charging platform with various mobile phones placed on it.

![Figure 1.7: (a) The Palm Touchstone CET charging device, with (b) the Palm Pre PDA docked to the charger. (c) A Wildcharge CET charging platform with various mobile phones on it.](image)

Although a handful of different CET charging platforms are commercially available, without a clear set of operational rules and standards, a large scale implementation of this technology is not very probably. This is due to possible compatibility problems that could arise between the different systems available from the various companies. Recently however, a consortium of large international companies have started a process of standardization of the CET technology for charging mobile electronic devices through magnetic induction. However, up to now they have released only limited information about the intended technology and standards they aim to use [27].

### 1.5 The Variable-Phase CET Platform

One of the objectives of this thesis is to develop and implement a variable-phase CET platform.

The variable-phase CET platform is a specialized implementation of the CET platform concept, and consists of multiple hexagon spiral windings arranged to form a matrix of windings. When a mobile electronic device, embedded with a similar “power receiving” winding, is placed on the CET platform surface, power is transferred from the platform to the mobile device through magnetic induction.
The primary and secondary windings are produced as copper tracks on printed circuit boards, in order to make them as small, cheap, and as thin as possible, so that they can easily be embedded into different devices, without adding extra weight or volume to them.

Multiple primary windings are used in order to obtain a high degree of freedom regarding the placement, orientation, and amount of secondary devices to charge, without the drawbacks posed by using large primary windings (such as increased power loss in the windings). In this way, only a few primary windings are used at a time, limiting the magnetic fields to active power transfer locations only.

The variable-phase CET platform is not intended for commercial use, or to replace any of the commercial CET charging devices available on the market, but rather, to serve as a research platform which is used to investigate some novel and fundamental CET-related concepts, which could possibly be used to improve future CET charging platforms.

The variable-phase CET platform operates by activating a cluster of three primary windings, located closest to the mobile device, for transferring power. In order for the CET platform to power load devices placed on its surface however, these load devices must first be located. To accomplish this, an innovative load detection scheme is developed. Unlike the load detection scheme presented in [18], this novel scheme does not make use of complex signalling circuitry, but instead monitors the primary windings’ equivalent terminal impedances. By doing so, the variable-phase CET platform can not only locate valid load devices, but also distinguish between soft-magnetic material and conductive material placed on the platform surface. Non CET-enabled devices, like keys, soft-drink cans, coins, pens, and certain magnetic devices, which might accidently be placed on the CET platform, can thus be detected, and the excitation of primary windings close to them, avoided.

Furthermore, many conventional CET platforms use a combination of thin ferrite plating and conductive copper sheeting to shield the internal electronic circuitry in the CET devices, from the alternating magnetic fields which are used to transfer energy. The variable-phase CET platform however, contains no soft-magnetic or conductive shielding materials. Instead, the variable-phase CET platform investigates the effectiveness of a third, alternative method of magnetic field shielding, using destructive wave interference. By exciting the three primary windings inside a activated winding cluster, with currents of different amplitudes and phases, it is shown that the magnetic field, especially the stray field, further away from the windings, undergo destructive wave interference, resulting in a reduction in the magnetic field values. Shielding is thus attained, even though no additional materials are used.

The variable-phase CET platform is designed to transfer up to 8 W of power to the secondary winding, over a maximum air gap size of 5 mm. The system operates at a frequency of 2.777 MHz, and is designed to induce a minimum voltage of 10 V (RMS) in the secondary winding, in order to ensure that 5 V - 6 V DC is available to the load device after it is rectified and down converted.
1.6 Research Goal and Objectives

The goal of this thesis is to develop and implement a variable-phase CET platform. To meet this goal, several objectives are formulated:

- **Modeling planar inductors on printed circuit boards**
  The first objective of this thesis is to build up a library of analytical and numerical methods and procedures for modeling planar inductors constructed as copper tracks on printed circuit boards, so that the use of FEM software can be minimized or avoided completely. The purpose of this “library of techniques” is not only to model the inductors used in the variable-phase CET platform, but to act as a single repository for information on modeling similar planar PCB inductors for various future CET applications.

- **Development of a load detection scheme**
  The second objective is to develop a load detection scheme which can be used to locate valid mobile devices placed on the CET platform, and also distinguish them from other non-CET-enabled objects that might have been accidently placed on the platform. This scheme should not use signalling circuitry, but rather monitor and measure the impedances of the primary windings in the CET platform, in order to detect the presence of the load devices.

- **Investigate the effectiveness of magnetic field shielding through destructive wave interference**
  The third objective of this thesis is to investigate the efficiency of a novel magnetic field shielding methodology using destructive wave interference. In contrast to the more conventional methods of shielding using magnetic and conductive materials, this method uses the amplitudes and phases of the currents in the primary windings, to alter the distribution of the magnetic field above the windings, resulting in a reduction of the stray magnetic fields.

- **Synthesis and implementation of the variable-phase CET platform**
  The final objective of this thesis is first to develop the theory, and synthesize the variable-phase CET platform based on given user requirements, in order for it to transfer the appropriate amount of power to the mobile devices. Second, is the development of the hardware necessary to implement the variable-phase CET platform in the form of a prototype system. The implemented prototype should then be used to verify the various operations of the CET platform through experimental measurements.

1.7 Organization of the Thesis

This thesis consists of six chapters (including this introduction). The contents of each chapter are briefly described hereunder.
The inductors used in the variable-phase CET platform are produced as copper tracks on printed circuit boards. **Chapter 2** is dedicated to introducing the reader to planar inductors and to build up a library of analytical and numerical methods and procedures for modeling them.

The purpose of **Chapter 3** is to synthesize the variable-phase CET platform. Here the CET platform user requirements are combined with the planar PCB inductor modeling methods presented in Chapter 2, and together with the inductive power transfer details and CET inductor- and platform particulars presented in this chapter, are used to calculate the electrical parameters and inductor geometric- and lumped parameters which will allow for the successful transfer of power to the mobile devices. Furthermore, the load detection process as well as the effects of destructive wave interference of the magnetic fields are calculated and discussed.

**Chapter 4** is dedicated to the implementation of the variable-phase CET platform, and to discuss, in detail the various design considerations taken in developing it. Here, a system overview of the complete CET platform is presented, and the CET prototype design is broken down into its individual subsystems, and together with their interactions, discussed. The primary-side and secondary-side subsystems and circuits as well as the primary current amplitude and phase controller are developed.

Experimental verification of the various operations of the variable-phase CET platform is performed and presented in **Chapter 5**. Here, firstly, the operation of the load detection process is verified. Afterwards various voltage-, current-, as well as magnetic field measurements are performed on the CET platform, in order to verify the theoretical results obtained in previous chapters.

**Chapter 6** contains the general conclusions from the results of the previous chapters and the recommendations for future research.

### 1.8 Publications

The contents of this thesis and related work have been published in journals and presented at a number of international conferences. These publications are listed below.


Modeling Planar Inductors on Printed Circuit Boards

“The scientists of today think deeply instead of clearly. One must be sane to think clearly, but one can think deeply and be quite insane.” - Nikola Tesla (1856 - 1943)

2.1 Introduction

Planar inductors produced as flat two-dimensional structures are an interesting and often useful alternative to regular inductors. They are especially useful in specialized electronic applications where size constraints are strict or where unique geometries are required, such as radio frequency identification (RFID) systems [28, 29] and CET systems [13, 14].

In all of these applications, planar inductors play a vital role as inductor or for the inductive coupling of two circuits for the purpose of transferring power, information, or both, over their inductive link. Accurately modeling these inductors to predict their behavior, is paramount in developing any application in which they will be used.

For the purpose of this work, the focus will mainly fall on the planar inductors produced as copper tracks on PCBs that is used in CET systems where no magnetic material is present in or around the inductors.

One existing, and quite popular method for modeling and extracting the lumped parameters from an inductor, is through the use of the finite element method (FEM). This numerical technique approximates various inductor parameters by modeling them in a two- or three-dimensional environment and solving the Maxwell’s equations in and around the devices. Using this purely numerical
method, can however be very time consuming especially when complex structures are simulated.

The purpose of this chapter is to introduce the reader to planar PCB inductors and to build up a library of analytical and numerical methods and procedures for modeling them, so that the use of FEM software can be minimized or avoided completely. The purpose of this “library of techniques” is not only to model the inductors used in the variable-phase CET system as presented in later chapters, but to act as a single repository for information on modeling similar planar PCB inductors for various future CET applications.

In this chapter, firstly the reader is introduced to planar PCB inductors and transformers which do not contain any magnetic material. Secondly, the most important parameters which predict the behavior of these devices are identified. The rest of the chapter is then dedicated to modeling and estimating these various parameters. These models include: estimating the DC- and AC-resistances (skin and proximity effect) of rectangular cross-sectioned planar inductors; estimating their mutual- and self-inductances; calculating the inductor magnetic field intensities; predicting inductor parasitic (parallel) capacitances; as well as discussing some various operational limiting factors, such as estimating the maximum inductor excitation frequency in regards to its self-resonating frequency, and the validity of the magneto-static approximation they are based on.

All these models are tested, and the results obtained from them are compared against measurement results, as well as results obtained from FEM simulations. All results show excellent agreement.

### 2.2 Planar Inductors

Planar inductors refer to a group of flat inductors which spans a planar area in two-dimensions but with very limited and small depth in the third dimension. Inductors produced as copper tracks on PCB’s can be classified as planar inductors since their planar dimensions are often one or two orders of magnitude larger then the copper track thickness (usually in the range of 35 µm to 105 µm).

In Fig. 2.1, six planar PCB inductors are depicted. These inductors mostly consist of multiple turns, and are often referred to as windings. Throughout this chapter, these six inductors are used for experimental validation of the presented models, and are labeled \( W_1, W_2, W_3, W_4, W_5, \) and \( W_6 \). The values for their geometrical parameters as shown in this figure, are given in Table 2.1. Here, \( w \) is the inductor track width, \( s \) is the inter-track spacing, \( N \) is the amount of turns, \( l^{EQ} \) is the total inductor track length, and \( d \) is the copper track thickness. In Table 2.1, the abbreviation n/a (meaning: not applicable) indicates that the certain geometrical parameter is not applicable to the specific inductor.
Figure 2.1: The six planar PCB inductors used to verify the different models created in this chapter. They are denoted as (a) $W_1$, (b) $W_2$, (c) $W_3$, (d) $W_4$, (e) $W_5$, and (f) $W_6$. 
Table 2.1: The geometric parameters for the six test windings.

<table>
<thead>
<tr>
<th></th>
<th>W₁</th>
<th>W₂</th>
<th>W₃</th>
<th>W₄</th>
<th>W₅</th>
<th>W₆</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R_O)</td>
<td>30 mm</td>
<td>n/a</td>
<td>n/a</td>
<td>20 mm</td>
<td>25 mm</td>
<td>25 mm</td>
</tr>
<tr>
<td>(R_I)</td>
<td>0 mm</td>
<td>n/a</td>
<td>n/a</td>
<td>0 mm</td>
<td>10 mm</td>
<td>15 mm</td>
</tr>
<tr>
<td>(H_O)</td>
<td>n/a</td>
<td>42 mm</td>
<td>120 mm</td>
<td>n/a</td>
<td>n/a</td>
<td>74 mm</td>
</tr>
<tr>
<td>(H_I)</td>
<td>n/a</td>
<td>16 mm</td>
<td>84 mm</td>
<td>n/a</td>
<td>n/a</td>
<td>56 mm</td>
</tr>
<tr>
<td>(D_O)</td>
<td>n/a</td>
<td>40 mm</td>
<td>100 mm</td>
<td>n/a</td>
<td>n/a</td>
<td>50 mm</td>
</tr>
<tr>
<td>(D_I)</td>
<td>n/a</td>
<td>12 mm</td>
<td>64 mm</td>
<td>n/a</td>
<td>n/a</td>
<td>30 mm</td>
</tr>
<tr>
<td>(N)</td>
<td>17</td>
<td>7.5</td>
<td>10</td>
<td>22</td>
<td>8</td>
<td>7</td>
</tr>
<tr>
<td>(w)</td>
<td>1.0 mm</td>
<td>1.0 mm</td>
<td>1.0 mm</td>
<td>0.5 mm</td>
<td>1.0 mm</td>
<td>1 mm</td>
</tr>
<tr>
<td>(s)</td>
<td>0.5 mm</td>
<td>1.0 mm</td>
<td>1.0 mm</td>
<td>0.25 mm</td>
<td>1.0 mm</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>(d)</td>
<td>35 (\mu)m</td>
<td>35 (\mu)m</td>
<td>35 (\mu)m</td>
<td>105 (\mu)m</td>
<td>35 (\mu)m</td>
<td>35 (\mu)m</td>
</tr>
<tr>
<td>(I_{EQ})</td>
<td>1.64 m</td>
<td>0.86 m</td>
<td>3.68 m</td>
<td>1.43 m</td>
<td>0.88 m</td>
<td>1.24 m</td>
</tr>
</tbody>
</table>

2.3 Lumped Inductor and Transformer Models

In general terms: the inductors in a CET system transfer energy from the primary inductor to a load connected to the secondary inductor, through their shared magnetic field.

An electrical device which closely relates to this phenomenon is the two winding transformer. When a voltage is applied to the primary winding of such an transformer, the resulting primary current creates a magnetic flux in the transformer core. In the case of an ideal transformer, all the flux is confined within the transformer core. According to Faraday’s law of induction, the magnetic flux in the core will induce an EMF in the secondary winding. When the secondary winding is connected to a load, a secondary current will flow, depending on the load impedance. In this way, electrical energy will flow through the secondary coil and load, even though they are not directly connected to the primary winding by conductive wires.

In reality, the flux created by the primary and secondary windings are not all confined to the transformer core, but also in the air surrounding the windings. In the case of these non-ideal transformers, the total flux can be separated into two parts. First, is the magnetizing flux which links the primary and secondary windings of the transformer through their shared magnetic field (magnetizing field). Second, is the leakage flux: the flux generated by one winding, which does not couple with the other. Since the leakage flux created by the primary winding does not couple with the secondary winding, it can not be used to generate an EMF in the secondary winding, and does not aid in the transfer of energy from the primary side to the load. Practical transformers are often modeled as non-ideal transformers.

When the magnetic core is completely removed from the transformer, it is often referred to as an air-cored transformer. Without the presence of the high-permeability material, the magnetic fields permeate though the air around the
windings without any predominant path, resulting in lower magnetizing flux and thus higher leakage flux.

Two magnetically coupled air-cored CET windings essentially form a two winding transformer with large leakage flux. Transformer theory can thus be used to model the behavior of two magnetically coupled CET windings.

Figure 2.2 (a) shows a CET system with two inductors transferring energy. On the left is the primary winding, denoted $W_P$, and on the right is the secondary winding $W_S$. Figure 2.2 (b) shows the equivalent lumped parameter model of the air-cored transformer.

Here, $L_P$ and $R_P$ are the primary winding inductance and resistance, respectively. The primary winding parasitic capacitance (also known as the stray or parallel capacitance) is given as $C_{P\text{par}}$. The current flowing into the primary winding is $I_P$, and $V_P$ is the voltage there over. Similarly, $L_S$ and $R_S$ are the secondary winding inductance and resistance, respectively. $C_{S\text{par}}$ is the secondary winding parasitic capacitance. The current flowing out of the secondary winding is given as $I_S$. The voltage over the secondary winding is $V_S$. The mutual inductance between the primary and secondary winding is $M_{PS}$.

The subscripts $P$ and $S$ are used to label the different inductor parameters in this chapter, and refer to the parameters of the primary and secondary CET inductors, respectively. In some of the models presented in this chapter however, these two parameters are omitted. In these cases, the models can be used for both the primary and secondary inductors.
2.4 Resistance Estimation

Accurately predicting the inductors’ resistances ($R_P$ and $R_S$ in Fig. 2.2 (b)) is important when designing CET applications, since they are some of the main factors that limit the power transfer capability and efficiency of CET systems.

In this section, methods are presented for estimating both the DC- and AC-resistances of planar inductors, with rectangular cross-sections. This method models the inductors in two-dimensions and the AC-resistance calculations takes into account skin- and proximity effect.

The DC-resistance calculations are generally based on simple analytical expressions where the conductor resistivity, its cross-sectional area and length is used to calculate the resistance [30, 31].

In AC-resistance calculations, the current density distribution inside the conductor needs to be determined before the frequency dependent resistance can be calculated. Due to the cross-sectional symmetry of circular conductors, there exist some semi-analytical means of approximating their current density and ergo their resistances [32, 33]. For inductors with non-circular cross-sections, numerical means, like the method of moments, must be used to solve the fields, current distributions and resistances [34, 35, 36, 37].

Here, firstly, analytical expressions for estimating the DC- and AC-resistances of planar inductors with rectangular cross-sections are presented. Secondly, a numerical implementation for predicting the frequency dependent AC-resistances is presented. Finally, the presented numerical implementation is verified through various experiments. The results are compared against measurements, and show excellent agreement.

2.4.1 Analytical Solution

DC-Resistance

The DC-resistance is the electrical resistance offered by an conductor to the flow of a DC-current. At DC and very low frequencies, the magnetic fields inside the conductor are approximately static. Since the time rate of change of the applied voltage source is approximate zero, no additional electric fields are induced inside the conductor. The current density in such a conductor is thus approximately constant and uniformly distributed. The DC-resistance can easily be calculated as:

$$R_{DC} = \frac{\rho l^{EQ}}{A}. \quad (2.1)$$

Here, $\rho$ is the electrical resistivity of the conductor material, $l^{EQ}$ is the length of the conductor and $A$ is the cross-sectional area of the conductor.

In general, the planar PCB inductors used in CET systems are excited with time-varying sinusoidal currents. The AC-resistance calculations are thus required for them. Calculating the DC-resistance however, can be useful, as it can
easily and quickly be calculated and used as a baseline or an initial ball park estimate for the AC-resistances.

AC-Resistance

The AC-resistance is the electrical resistance offered by an conductor to the flow of an AC-current. Due to the time-varying currents flowing inside the conductors and consequent time-varying magnetic fields and magnetic flux, electric fields are induced inside the conductors which oppose the applied source voltage. These induced electric fields give rise to eddy currents which directly oppose the initially applied current. Consequently, a new higher applied voltage is needed to maintain the same current. Essentially, this phenomenon can be seen as an increase in conductor resistance due to current crowding inside the conductor. The current density will no longer be constant, and the distribution of the current density needs to be estimated, in order to obtain the AC-resistance values.

When the equivalent rise in resistance is due to fields inside the same conductor, the phenomenon is also referred to as the skin effect. Moreover, when the rise in resistance is due to fields created by neighboring conductors, it is referred to as the proximity effect. Figure 2.3 (a) shows a graphical representation of the skin effect and shows the primary current, the internal conductor magnetic field, the flux area due to the magnetic field, as well as the flow of eddy currents opposing the primary current. Figure 2.3 (b) shows the proximity effect, where a current in one conductor creates magnetic fields inside neighboring conductors.

Figure 2.3: (a) Time-varying currents, $I_P$, through a conductor with cross-section, $C_1$, create magnetic fields and eddy currents within the same conductor. These eddy currents oppose the primary current and are referred to as the skin effect. (b) Time-varying currents through a conductor with a cross-section $C_3$ create magnetic fields and eddy currents in the neighboring conductors with cross-sections $C_2$ and $C_4$. This is called the proximity effect.
The analytical solution for the AC-resistance of inductors with rectangular cross-sections is developed by modeling the inductor tracks in a two-dimensional plane. Here, the conductor is assumed to be infinite in length. The solution is based on the work in [36], and it is further expanded to include the proximity effect.

The solution is derived by starting from the Maxwell equations, that can be written as [38]:

\[ \nabla \cdot B = 0, \]  
\[ \nabla \cdot D = \rho, \]  
\[ \nabla \times E = -\frac{\partial B}{\partial t}, \]  
\[ \nabla \times H = J + \frac{\partial D}{\partial t}. \]

Here, \( B \) is the magnetic field, \( D \) is the electric displacement field, \( H \) is the magnetic field intensity, \( E \) is the electric field, \( \rho \) is the total charge density, and \( J \) is the current density.

The magnetic field and magnetic field intensity relates to each other as \( B = \mu_0 \mu_r H \), where, \( \mu_0 \) is the permeability of free space, and \( \mu_r \) is the relative permeability of the medium. Similarly, the electric displacement field relates to the electric field as \( D = \varepsilon_0 \varepsilon_r E \). Here \( \varepsilon_0 \) is the permittivity of free space, and \( \varepsilon_r \) is the relative permittivity of the medium.

In order to solve these equations, the electric displacement current \( \partial D/\partial t \) is assumed to be much smaller then the applied current, and can thus be neglected [34, 36].

Furthermore, the magnetic field \( B \), relates to the magnetic vector potential \( A \), through the curl operator as:

\[ B = \nabla \times A. \]

Also, the electric field, \( E \), relates to the current density \( J \), through Ohm’s law as:

\[ J = \sigma E. \]

Here, \( \sigma \) is the conductivity of the conductive material.

By solving the Maxwell equations and noting that due to the presence of time-varying currents and subsequent time-varying magnetic fields inside the conductor, the electric field \( E \) is no longer conservative and can not be described in terms of a scalar potential \( \varphi \) alone. By including the magnetic vector potential however, the electric field becomes:

\[ E = -\nabla \varphi - \frac{\partial A}{\partial t}. \]
By substituting Ohm’s law (2.7) into (2.8), this equations can be rewritten in terms of current densities as:

\[ J = \sigma E = -\sigma \nabla \varphi - \sigma \frac{\partial A}{\partial t}. \]  

(2.9)

The conductor is assumed to be infinite in length and directed along the z-axis. The current is thus also directed in the z-direction only. Subsequently, the magnetic field lies solely in the xy-plane, so that \( A_x = A_y = 0 \). The current density and magnetic vector potential thus only have components in the z-direction, and are independent of \( z \). The vectors \( A \) and \( J \) can thus be written as scalars with the understanding that they are solely z-directed. Equation (2.9) can thus be rewritten as:

\[ J_z(x, y, t) = -\sigma \nabla \varphi - \sigma \frac{\partial A_z}{\partial t}(x, y, t). \]  

(2.10)

Here, the first term on the right-hand side, \(-\sigma \nabla \varphi\), is the current flow due to the applied source voltage, and can be written as \( J_{SRC} \). Finally, the analytical solution for calculating the AC-resistance due to the skin effect in a rectangular cross-section can be written as:

\[ J_z(x, y, t) = J_{SRC}(t) - \sigma \frac{\partial}{\partial t} A_z(x, y, t). \]  

(2.11)

Furthermore, the magnetic vector potential can be written in terms of the current density by noting that the magnetic vector potential relates to the magnetic field and magnetic field intensity through the curl operator as:

\[ \mu_0 \mu_r H = B = \nabla \times A. \]  

(2.12)

With the Coulomb gauge condition \( \nabla \cdot A = 0 \), and the well known operator relation \( \nabla \times \nabla \times A = \nabla (\nabla \cdot A) - \nabla^2 A \), (2.12) becomes:

\[ \nabla^2 A = -\mu_0 \mu_r J. \]  

(2.13)

Equation (2.13) is also referred to as the Vector Poisson’s equation.

Due to the two-dimensional z-directed current distribution in the conductor, (2.13) can be written as [36, 38]:

\[ A_z(x, y, t) = \frac{\mu_0 \mu_r}{2\pi} \int_{C_1} J_z(x', y', t) \ln \left( \sqrt{(x-x')^2 + (y-y')^2} \right) dx' dy'. \]  

(2.14)

Here, \( C_1 \) is the conductor cross-section.

In view of (2.14), (2.11) can be written in terms of current densities alone

\[ J_z(x, y, t) = J_{SRC}(t) - \left( \frac{\sigma \mu_0 \mu_r}{2\pi} \right) \frac{\partial}{\partial t} \int_{C_1} J_z(x', y', t) \ln \left( \sqrt{(x-x')^2 + (y-y')^2} \right) dx' dy'. \]  

(2.15)
For a time-harmonic excitation, that is, \( J_\hat{z}(x,y,t) = J_\hat{z}(x,y)e^{j\omega t} \), (2.15) can be presented in phasor form as:

\[
J_\hat{z}(x,y) = J_{SRC} - \frac{j\omega \sigma \mu_0 \mu_r}{2\pi} \int_{C_1} J_\hat{z}(x',y') \ln \left( \sqrt{(x-x')^2 + (y-y')^2} \right) dx' dy'. \tag{2.16}
\]

Equation (2.16) gives the means for solving the current distribution inside the conductor cross-section. However, due to the definite integral in the equation, it can not be easily solved or implemented.

### 2.4.2 Numerical Implementation

The numerical solution for the AC-resistance of a conductor with a rectangular cross-section is derived by subdividing the conductor cross-section into multiple sub-conductors. By making these sub-bars small enough, their individual current densities are assumed to be constant, and the integral in (2.16) is converted into a finite sum which can be solved numerically.

First, let the conductor cross-section be divided into \( Q \)-amount of sub-bars. Now, let \( S_m \), and \( S_n \) be two individually indexed sub-bars, located at coordinate points, \( x_m, y_m \) and \( x_n, y_n \), with areas \( A_m = (w_m)(d_m) \) and \( A_n = (w_n)(d_n) \), as shown in Fig. 2.4, respectively.

![A conductor with a rectangular cross-section, showing two of its sub-bars.](image)

Equation (2.16) is integrated over sub-bar \( S_m \) so that:

\[
\int_{A_m} J_\hat{z}(x_m, y_m) \, dx \, dy = \int_{A_m} J_{SRC}(y_m, x_m) \, dx \, dy - \frac{j\omega \sigma \mu_0 \mu_r}{2\pi} \times \int_{A_m} \left( \int_{C_1} J_\hat{z}(x', y') \ln \left( \sqrt{(x-x')^2 + (y-y')^2} \right) dx' \, dy' \right) \, dx \, dy. \tag{2.17}
\]

Assuming a constant current density in the sub-bars, the integral over \( C_1 \) in (2.17), is converted into a finite sum over all \( Q \) sub-bars, using the newly indexed sub-bars denoted as \( S_n \). Equation (2.17) can be rewritten as:
2.4. Resistance Estimation

\[ J_z(x, y)A_m = J_{SRC}(x, y)A_m - \frac{j\omega \sigma \mu_0 \mu_r}{2\pi} \times \]

\[ \sum_{n}^{Q} J_z(x', y') \int_{A_m} \int_{A_n} \ln \sqrt{(x-x')^2 + (y-y')^2} dx' dy' dx dy. \] (2.18)

The geometrical mean distance between two different rectangles \( S_m \) and \( S_n \) is defined in as:

\[ \ln D_{mn} = \int_{A_m} \int_{A_n} \ln \sqrt{(x-x')^2 + (y-y')^2} dx' dy' dx dy \] (2.19)

of which the exact expression is given in [36].

Equation (2.18) thus becomes:

\[ J_z(x_m, y_n) = J_{SRC}(x_m, y_n) - \frac{j\omega \sigma \mu_0 \mu_r}{2\pi} \sum_{n}^{Q} J_z(x'_n, y'_n)A_n \ln D_{mn}, \] (2.20)

for \( m = 1 \cdots Q \), and \( n = 1 \cdots Q \).

This set of complex equations can be written in matrix form as:

\[
\begin{bmatrix}
1 + k_{11} & k_{12} & k_{13} & \cdots & k_{1Q} \\
 k_{21} & 1 + k_{22} & k_{23} & \cdots & k_{2Q} \\
 k_{31} & k_{32} & 1 + k_{33} & \cdots & k_{3Q} \\
 \vdots & \vdots & \vdots & \ddots & \vdots \\
 k_{Q1} & k_{Q2} & k_{Q3} & \cdots & 1 + k_{QQ}
\end{bmatrix}
\begin{bmatrix}
J_z(S_1) \\
J_z(S_2) \\
J_z(S_3) \\
\vdots \\
J_z(S_Q)
\end{bmatrix}
= \begin{bmatrix}
J_{SRC}(S_1) \\
J_{SRC}(S_2) \\
J_{SRC}(S_3) \\
\vdots \\
J_{SRC}(S_Q)
\end{bmatrix},
\]

(2.21)

Here \( k_{mn} = j\omega \sigma \mu_0 \mu_r A_n \ln D_{mn}/2\pi \).

This matrix equation can be rewritten in a compact matrix form as:

\[ KJ_z = J_{SRC} \] (2.22)

and solved for \( J_z \).

Finally, the AC-resistance due to the skin effect is calculated as

\[ R_{AC} = \frac{\int_{C_1} |J_z|^2 dC_1}{\sigma |I_P|^2}, \] (2.23)

where \( I_P \) denotes the current through the cross-section \( C_1 \).

Similarly, the proximity effect of neighboring inductor tracks can be included into the AC-resistance estimations by including their magnetic field contributions into (2.20).
2.4.3 Experimental Validation

Firstly, the DC-resistances for the six test windings shown in Fig. 2.1 are measured using a Wayne Kerr Automatic LCR Meter (DC - 1 MHz). The DC-resistances are then calculated using (2.1) and the track lengths as shown in Table 2.1. The results are shown in Table 2.2.

Table 2.2: Calculated and measured DC-resistance values.

<table>
<thead>
<tr>
<th></th>
<th>Calculated Resistance</th>
<th>Measured Resistance</th>
<th>Absolute Error [100%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_1$</td>
<td>810 mΩ</td>
<td>841 mΩ</td>
<td>3.69 %</td>
</tr>
<tr>
<td>$W_2$</td>
<td>424 mΩ</td>
<td>435 mΩ</td>
<td>2.53 %</td>
</tr>
<tr>
<td>$W_3$</td>
<td>1820 mΩ</td>
<td>1900 mΩ</td>
<td>4.40 %</td>
</tr>
<tr>
<td>$W_4$</td>
<td>471 mΩ</td>
<td>486 mΩ</td>
<td>3.09 %</td>
</tr>
<tr>
<td>$W_5$</td>
<td>433 mΩ</td>
<td>441 mΩ</td>
<td>1.81 %</td>
</tr>
<tr>
<td>$W_6$</td>
<td>608 mΩ</td>
<td>633 mΩ</td>
<td>3.95 %</td>
</tr>
</tbody>
</table>

Secondly, the AC-resistances of the test windings, $W_1$, $W_4$, and $W_6$ are calculated using (2.23). First, the AC-resistances due to only the skin effect are calculated. The per-unit-length resistances are multiplied by the total inductor track lengths (from Table 2.1) to obtain the total AC-resistances. Second, the AC-resistances due to the skin and proximity effect together are calculated. The calculations are done for an increasing amount of neighboring conductors, whereafter their average resistances are estimated and multiplied by the inductor track lengths. The AC-resistance results of the three test windings are $W_1$, $W_4$, and $W_6$, are shown in Fig. 2.5, Fig. 2.6, and Fig. 2.7, respectively. Here, the dotted lines indicate the +10 % and -10 % measurement margins, respectively.

2.4.4 Conclusions

In this section, methods for estimating the DC- and AC-resistances of planar PCB inductors are given.

The DC-resistance measurement results show good agreement with the estimated values, with a maximum absolute error of less than 5 %.

The AC-resistance estimation results show that the skin effect alone, is not enough to accurately model the planar PCB inductors at high frequencies. In the frequency range of up to approx. 100 kHz, all the calculated AC-resistance results show similar behavior. In this frequency range, the AC-resistance values are mainly dominated by the skin effect. At frequencies above 100 kHz however, the AC-resistance results start to diverge. At these frequencies, the AC-resistance becomes noticeably more influenced by the proximity effect, which can no longer be ignored at the higher frequencies. The results furthermore indicate, that at least 50 % to 75 % of the neighboring conductor tracks should be modeled to obtain accurate results, for the tested frequency range. It is possible that even
2.5 Magneto-Static Approximation and the Current Filament Model

A magneto-static (or magneto-quasi-static) system refers to a system which is governed by magnetic fields originating from current density elements. In magnetic circuits where the operating frequency is relatively low, the circuit can more tracks should be modeled at higher frequencies. For the frequency range of up to 5 MHz however, the AC-resistance measurements show good agreement with the estimated values.

Figure 2.5: Calculated and measured AC-resistances for winding $W_1$ (Fig. 2.1 (a)).

Figure 2.6: Calculated and measured AC-resistances for winding $W_4$ (Fig. 2.1 (d)).
be described as magneto-static, in which case a simplified subset of the Maxwell equations can be used to solve the magnetic fields outside the conductors. In these equations, the electric displacement current is assumed to be negligible (compared to the applied electric current).

In air-cored CET systems, where no soft-magnetic material is present, the magnetic field permeates through the air without any predominant path. Due to the linearity of the permeability of air, the magneto-static version of the Maxwell equations in integral form can be solved to calculate the magnetic field created by CET windings, if the applied current is known. Furthermore, through superposition, these calculations can be performed numerically and implemented in a computer program.

However, it is important to note that models based on the magneto-static approximation hold true only if these approximation assumptions are met. This will be discussed in more detail in section 2.10.

To further simplify the Maxwell equations, the copper tracks of the planar PCB inductors can be modeled by thin thread-like structures called current filaments, or filamentary wires. These filaments have an infinitely small cross-section, but have a finite length. By using filamentary currents, the current density elements inside the magnetic field integral equations (as shown in section 2.8) can be moved out of the integral, which significantly simplifies these equations, and in some cases allows for analytical solutions to the problems.

Due to the planar nature of the copper tracks and the magnetic fields’ reper-torial dependency to distance, the magnetic fields outside the conductors, created by a uniform-, non-uniform-, and filamentary currents inside the conductor, are approximately identical.

In Fig. 2.8, four different scenarios are depicted, where current is flowing through a two-dimensional rectangular cross-section (representing an inductor track). In Fig. 2.8 (a), the current density is uniformly distributed over the cross-
section. In Fig. 2.8 (b), the current is concentrated in the corners of the rectangle, approximating a high-frequency current distribution. In Fig. 2.8 (c), the current is represented by a current filament placed at the center of the rectangle, and in Fig. 2.8 (d), the current is modeled by multiple current filaments with $\Delta w$ separation between them.

![Figure 2.8: The magnetic field created by a two-dimensional rectangular conductor cross-section with (a) a uniform current density, (b) a high-frequency current density, (c) a single current filament, and (d) multiple current filaments.](image)

Here, the distance $h$, is the minimum height above the conductor, at which the magnetic field will be evaluated in the CET application. The two-dimensional axis origin is in the center of the conductors. The amount current filaments used to model the conductor in Fig. 2.8 (d) is given as $N$.

By solving Ampere’s integral law in two dimensions for these different scenarios, the magnetic field intensity at position $P$ can be estimated. The magnetic field intensity resulting from the magneto-static approximation as shown in Fig. 2.8 (a) is labeled $H_a$. Similarly, the magnetic field intensity resulting from the high-frequency current distribution, as shown in Fig. 2.8 (b) is labeled as $H_b$. The magnetic field intensity from the single filament approximation as shown in Fig. 2.8 (c) is labeled as $H_c$, and the magnetic field intensity resulting from the multiple filament approximation as shown in Fig. 2.8 (d) is labeled as $H_d$. The fields $H_a$ and $H_b$ represents the two limit cases for a static and a high-frequency magnetic field intensity.

The most convenient scenario is where the conductor current is modeled by a single current filament. For high-frequency applications, this means comparing the magnetic fields from Fig. 2.8 (b) and (c), and determining the magnetic field error in relation to the minimum distance $h$, and the track dimensions. The error between the two fields can be expressed as:

$$\varepsilon_{bc} = 1 - \frac{H_b(P)}{H_c(P)}.$$  \hfill (2.24)

If $h$ is a fixed value, then the magnetic field error $\varepsilon_{bc}$ can be estimated. The error should be kept to a minimum to assure accurate filamentary approximations. If the error is too large, multiple filaments should be used to model the conductor. In this way, the minimum amount of current filaments which is needed to accurately model the magnetic field, at the distances required, can be ascertained. This is also referred to as the mesh-matrix method [40, 42, 45].
The planar PCB inductors are modeled by placing the current filament(s) at the center of the tracks. Inductors consisting of straight line segments only, like square-, rectangular-, hexagonal-, and octagonal windings and spirals, can be easily modeled this way. Figure 2.9 (a) and (b) shows how a rectangular PCB spiral inductor with 3 turns is modeled by eleven current filaments.

Planar inductors consisting of curved and arched windings, like the rounded rectangular spiral shown in Fig. 2.10 (a) can not be directly modeled in this way. For analytical solutions to the field equations, analytical expressions for the curved filament paths are required. Later in this chapter it is shown, however, that numerical solutions are more practical for these types of windings. In these cases, replacing the curved windings with multiple smaller straight line filaments is more useful. Choosing the amount of filaments used to approximate curved inductor tracks is not a straight-forward task however, and depends on the curvature [46] of the track in question. In [51] and section 2.6, this issue is discussed in more detail. Figure 2.10 (b) shows how this curved inductor can be modeled with multiple smaller straight line current filaments.

Figure 2.9: (a) A straight rectangular spiral inductor on PCB and (b) its tracks modeled by current filaments.

Figure 2.10: (a) A circular rectangular spiral inductor on PCB and (b) its tracks modeled by current filaments.
2.5. Magneto-Static Approximation and the Current Filament Model

2.5.1 Quantitative Analysis

To quantify the magnetic field errors associated with the filament approximations described in this section, the magnetic fields and the magnetic field errors are calculated for an infinitely long conductor with a rectangular cross-section of \( w = 0.5 \) mm and \( d = 105 \) µm (representing an inductor winding track). The conductor is placed at the \( xz \)-axis origin and is directed in the \( y \)-direction.

Firstly, the distribution of the absolute magnetic field intensity, resulting from a constant current density in the cross-section, is calculated in the \( x \)-direction, at a height of \( z = 1 \) mm, as shown in Fig. 2.8 (a). This represents the magneto-static approximation of the magnetic field. Secondly, the distribution of the magnetic field created by a high-frequency current distribution, as shown in Fig. 2.8 (b), is calculated. Thirdly, the magnetic field created by a single filament placed at the center of the track, as shown in Fig. 2.8 (c), is calculated. Figure 2.11 shows the resultant magnetic fields distributions.

![Figure 2.11](image)

**Figure 2.11:** The magnetic field intensities resulting from the magneto-static, high-frequency, and single filament approximation of the current distribution in the inductor track cross-section, respectively.

From Fig. 2.11 it can be seen that the three magnetic field distributions are not the same, with the biggest differences located above the center of the tracks.

Secondly, the magnetic field errors are calculated using the single filament approximation as reference. The results are shown in Fig. 2.12. Here, \( \varepsilon_{bc} \) is the error between fields \( H_b \) and \( H_c \), and \( \varepsilon_{ac} \) is the error between fields \( H_a \) and \( H_c \).

From Fig. 2.12 it can be seen that the maximum field errors occur above the center of the tracks. In this specific case, the maximum absolute error between the magneto-static and single filament approximation, \( \varepsilon_{ac} \), is approximately 2 \%. The maximum absolute error between the high-frequency and single filament approximation, \( \varepsilon_{bc} \), is approximately 5 \%.
2.6 Mutual Inductance Estimation

The mutual inductance between two inductors is a measure of their magnetic coupling and the amount of shared magnetic field between them. Accurately estimating the mutual inductance between two planar PCB inductors ($M_{PS}$ in Fig. 2.2 (b)) is essential because it significantly impacts the power transfer efficiency and capability of the CET system they are used in.

In [39, 40, 41] various methods are presented for calculating the mutual inductance between coaxial parallel circular filaments, and conductors. In [42, 43, 44] these approaches are extended to include parallel circular filaments with lateral displacement, thus non-coaxial filaments. Moreover, in [45] this theory is extended to include non-coaxial circular filaments with inclined axes.

When working with applications which employ circular windings, these semi-analytical method can be used. However, when working with applications which utilize straight lined windings and spirals, like hexagon or rectangular spiral windings, these methods above can no longer be used. In these cases numerical ap-
proximations of the analytical solutions are more useful.

In this section, methods are presented for estimating the mutual inductance between various types of inductors modeled as filamentary windings. Firstly, a general analytical mutual inductance solution is presented. Secondly, the analytical solution is converted into a numerical approximation, which can be implemented as an algorithm in a computer program to estimate the mutual inductance between various straight-lined multi-filamentary structures. Finally, the presented numerical implementation is verified through various experiments. The results are compared against magneto-static FEM simulations, results found in literature, as well as measurements, and show excellent agreement with a maximum absolute error of less than 5%.

2.6.1 Analytical Solution

The analytical solution for calculating the mutual inductance between two filamentary structures is based on the magneto-static approximation, and is derived using the magnetic vector potential, Stokes theorem [38], and Faraday’s law of induction and is formulated as:

\[
M(l, k) = \mu_0 \frac{1}{4\pi} \oint_l \oint_k \frac{dl \cdot dk}{|r - r'|},
\]  

Equation (2.25) is also known as the Neumann formula [39, 40, 49]. The integral can unfortunately only be completed analytically for relatively few configurations.
2.6.2 Numerical Implementation

Since an analytical solution of (2.25) is in general quite impractical, numerical integration can be used to approximate it by evaluating the integrant in very small increments. The advantage of using numerical integration as proposed in this section is that the definite integral in (2.25) can be implemented with low computational effort to evaluate arbitrary positioned and orientated filamentary structures. The procedure for implementing the numerical integration is shown in the next seven steps [50, 51]:

1. The total amount of straight line segments on the primary filament contour, \( l \), and the secondary filament contour, \( k \), is determined. They are stored as \( a^l \) and \( a^k \), respectively.

2. The total amount of vertices (including start and stop positions) on the primary and secondary structures is determined. They are stored as \( b^l \), where \( b^l = a^l + 1 \), and \( b^k \), where \( b^k = a^k + 1 \), accordingly.

3. The three-dimensional coordinates of the primary- and secondary vertices are determined and stored in vectors \( q^l \) and \( q^k \), respectively.

4. The Euclidean lengths of all the primary and secondary filaments are then calculated and stored in the arrays \( c^l \) and \( c^k \), consequently.

5. The normalized filament vectors are calculated as:

\[
n^l_m = \frac{d^l_{m+1} - d^l_m}{c_m}, \quad m \in [1 \cdots a^l].
\]  

(2.26)

Here, \( n^l \) represents the normalized position vectors for the primary structure filaments. Similarly, the normalized position vectors for the secondary filaments can be calculated and is represented as \( n^k \). The subscript \( m \) denotes the appropriate filament.

6. The flux linkage between the primary and secondary structures is then estimated as:

\[
\lambda'(l, k) = \frac{\mu_0 I_P \Delta_l \Delta_k}{4\pi} \sum_{\alpha=1}^{a^l} \sum_{\beta=1}^{a^k} \sum_{\chi=0}^{X} \sum_{\xi=0}^{E} \frac{n^l_{\alpha} \cdot n^k_{\beta}}{|K_{\chi} - K_{\xi}|},
\]  

(2.27)

where

\[
K_{\chi} = q^l_{\alpha} + (\chi)(\Delta_l) \cdot n^l_{\alpha},
\]  

(2.28)

\[
K_{\xi} = q^k_{\beta} + (\xi)(\Delta_k) \cdot n^k_{\beta},
\]  

(2.29)
$X = c^l_\alpha / \Delta_l, \ E = c^k_\beta / \Delta_k. \ \ \ \ (2.30)$

Here, the lengths $\Delta_l$ and $\Delta_k$ are small numerical integration elements.

7. Finally, the mutual inductance between the two filaments can then be estimated from (2.27) as:

$M'_{l,k} = \lambda'_{l,k} / I_P. \ \ \ \ (2.31)$

Caution must be taken when choosing the values for $\Delta_l$ and $\Delta_k$. On the one hand, choosing these values too small will increase the accuracy of the model but also increase the processing time needed to solve the numerical integral. On the other hand, choosing these values too big reduces the model accuracy. By iterating (2.27) with decreasing values of $\Delta_l$ and $\Delta_k$, the convergence of the resultant mutual inductance can be used as a gauge in determining the accuracy of the model. Generally, these values should be about the same order of magnitude (preferably smaller) than the PCB track thickness and inter-track spacing.

Furthermore, when working with applications which utilize rounded rectangular inductors (Fig. 2.1 (f) and [48]), elliptical inductors, or half- and quarter circle geometry inductors [47], none of the mentioned semi-analytical or numerical methods can be used directly. In these cases, the curved tracks can be modeled by multiple straight-lined filaments placed on the track contours. In this way, the whole inductor can be approximated by straight lines and (2.27) can be used to calculate the mutual inductance. The curvature of a circular track is given as the reciprocal of its radius [46]. Generally for circular filaments with radii up to $R = 50$ mm, it is enough to choose the amount of modeled vertices between $2/R$ to $4/R$, with $R$ given in millimeter.

2.6.3 Experimental Validation

The numerical method for estimating the mutual inductances of planar PCB inductors as proposed in subsection 2.6.2, is demonstrated through two sets of experiments.

Experiment Set 1

In the first set of experiments, the mutual inductance between two hexagon spiral windings, as illustrated in Fig. 2.14 (a), is calculated using (2.27) - (2.31). The primary winding has a radius of 12 mm, and 13 turns. Its copper thickness is 105 $\mu$m, the track width is 0.5 mm, and the inter-track spacing is 0.25 mm. The secondary winding is winding $W_4$ as shown in Fig. 2.1 (d) and with geometrical parameters as described in Table 2.1.

With $\Delta_l = \Delta_k = 0.5$ mm, the primary winding is placed at the axis origin, and the secondary winding is placed in parallel to the primary at different $x$-, $y$-, and $z$-coordinates, as shown in Fig. 2.14 (a). Figure 2.14 (b) shows the setup used to measure the mutual inductance. Here, the secondary hexagon spiral winding
is placed on top of a matrix of hexagon spiral windings. The activated primary winding is hidden underneath the secondary winding, and white plastic spacers are used to separate the windings. The current probe, measuring the primary current, and voltage probe, measuring the secondary induced voltage, can also be seen.

The results of the calculated and measured mutual inductances are shown in Table 2.3. The agreement between calculated and measured values is within 5% in all situations.

![Figure 2.14: (a) The positions of the primary and secondary windings during mutual inductance calculations and measurements. (b) The setup used to measure the mutual inductances between the hexagon spiral windings.](image)

### Experiment Set 2

In [44] a method is presented for calculating the mutual inductance of movable planar coils on parallel surfaces. Here the planar coils are two circular spiral inductors sandwiched between two ferrite layers. The mutual impedance between the primary and secondary windings is estimated by first calculating their mutual inductance in free space and then adding the additional impedance due to the presence of the magnetic material. The formula for calculating the mutual inductance between two non-coaxial circular filaments in free-space, as given in [44] is:

\[ M_{PS} = \mu_0 \Upsilon_P \Upsilon_S \int_0^\infty \int_0^{\pi} \frac{\Upsilon_S - z' \cos \varphi}{r} J_1(k\Upsilon_P)J_1(kr)e^{-k \Upsilon_D}d\varphi dk. \]  

(2.32)

Here, as shown in Fig. 2.15, \( \Upsilon_P \) and \( \Upsilon_S \) are the radii of the primary and secondary windings, respectively, and \( r = \sqrt{\Upsilon_P^2 + z'^2 - 2\Upsilon_S z' \cos \varphi} \). The mutual inductance between the circular filaments is \( M_{PS} \), and \( J_1(\cdot) \) is a Bessel function of the first kind. The axial distance between the centers of the two windings is given...
Table 2.3: Results of the mutual inductance calculations and measurements for the first set of experiments. Measurements are performed at a frequency of \( f = 500 \text{ kHz} \).

<table>
<thead>
<tr>
<th>Secondary winding position ((x',y',z')) [mm]</th>
<th>Calculated mutual inductance (M_{PS}^{\text{calc}}) [(\mu\text{H})]</th>
<th>Measured mutual inductance (M_{PS}^{\text{meas}}) [(\mu\text{H})]</th>
<th>Absolute error [(100%)]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 0, 2</td>
<td>3.20 (\mu\text{H})</td>
<td>3.16 (\mu\text{H})</td>
<td>1.27%</td>
</tr>
<tr>
<td>0, 0, 3</td>
<td>2.75 (\mu\text{H})</td>
<td>2.64 (\mu\text{H})</td>
<td>4.17%</td>
</tr>
<tr>
<td>0, 0, 4</td>
<td>2.37 (\mu\text{H})</td>
<td>2.35 (\mu\text{H})</td>
<td>0.85%</td>
</tr>
<tr>
<td>0, 0, 5</td>
<td>2.05 (\mu\text{H})</td>
<td>2.01 (\mu\text{H})</td>
<td>1.99%</td>
</tr>
<tr>
<td>10.4, 0, 2</td>
<td>1.40 (\mu\text{H})</td>
<td>1.43 (\mu\text{H})</td>
<td>2.10%</td>
</tr>
<tr>
<td>10.4, 0, 3</td>
<td>1.25 (\mu\text{H})</td>
<td>1.29 (\mu\text{H})</td>
<td>3.10%</td>
</tr>
<tr>
<td>10.4, 0, 4</td>
<td>1.11 (\mu\text{H})</td>
<td>1.15 (\mu\text{H})</td>
<td>3.48%</td>
</tr>
<tr>
<td>10.4, 0, 5</td>
<td>1.00 (\mu\text{H})</td>
<td>1.03 (\mu\text{H})</td>
<td>2.91%</td>
</tr>
<tr>
<td>12, 0, 2</td>
<td>0.90 (\mu\text{H})</td>
<td>0.92 (\mu\text{H})</td>
<td>1.75%</td>
</tr>
<tr>
<td>12, 0, 3</td>
<td>0.83 (\mu\text{H})</td>
<td>0.80 (\mu\text{H})</td>
<td>3.00%</td>
</tr>
<tr>
<td>12, 0, 4</td>
<td>0.76 (\mu\text{H})</td>
<td>0.75 (\mu\text{H})</td>
<td>1.20%</td>
</tr>
<tr>
<td>12, 0, 5</td>
<td>0.70 (\mu\text{H})</td>
<td>0.72 (\mu\text{H})</td>
<td>2.23%</td>
</tr>
</tbody>
</table>

as \(p\), and their vertical displacement as \(z'\). In one of the examples, the mutual inductance is calculated between two circular spiral windings. The spirals are modeled as multiple coaxial circular filaments of which the mutual inductances are estimated using \((2.32)\). The primary and secondary spirals are identical with outer radii \(\Upsilon_P = \Upsilon_S = 15 \text{ mm}\), track widths of 0.5 mm, inter-track spacing of 0.3 mm, and 6 turns each.

In the first test, with the air gap fixed at \(z' = 2 \text{ mm}\), the mutual inductances are calculated while the axial distance is increased from \(p = 0\) and 30 mm. The results are shown in Fig. 2.16 (a). In the second test, the axial distance is fixed at \(p = 9 \text{ mm}\) and the mutual inductances are calculated while the air gap is increased from \(z' = 2 \text{ mm}\) to 6 mm. The results are shown in Fig. 2.16 (b). In Fig. 2.16, the straight lines indicate the semi-analytical results calculated from \((2.32)\), while the circles indicate the results obtained using the numerical method presented in this section. It is clear that the proposed numerical approach leads to good results.

In Fig. 2.16 (a) negative mutual inductance results are predicted. Negative mutual inductance values, however, are not physically possible. Here, the negative values indicate sign changes in the induced voltages over the terminals of the windings. In Fig. 2.23 (subsection 2.8.3) two two-dimensional views of the magnetic field distributions of test windings \(W_3\) and \(W_5\) are shown, respectively. Here it can be seen that the magnetic fields have large positive values at the centers of the windings, but negative values at the edges. A secondary winding placed around the edges of these primary windings thus have a negative net flux linkage, resulting in a negative induced voltage and subsequent negative mutual
2.6.4 Conclusions

In this section, methods for estimating the mutual inductance between two planar PCB inductors are given. These methods are all based on the magneto-static assumption and uses current filaments to model the inductor tracks. For circular windings, there exist various semi-analytical methods for estimating their mutual inductances. For inductors with straight tracks however, these methods can not be used. Here, a numerical implementation of the Neumann formula can be used to approximate the mutual inductance. Planar PCB inductors containing circular or curved tracks can also be modeled using this numerical method, by modeling their curved tracks as multiple shorter straight filaments. This numerical implementation is verified through various experimental measurements and magneto-static FEM simulations, and show excellent agreement, for straight-lined as well as curved planar PCB inductors.
2.7 Inductance Estimation

Inductance can be described by the electromotive force induced in a circuit by the change in current in the same circuit, due to its own flux linkage. Accurately estimating the self-inductances of planar PCB inductors ($L_P$ and $L_S$ in Fig. 2.2 (b)) are important because they are used in the CET power transfer calculations. It is also a vital parameter in tuning the resonant frequency of resonant CET systems.

In [49, 53], a few analytical approximations for circular and rectangular filamentary wire loops are given. Furthermore, in [52] analytical expressions for circular-, hexagonal-, square, and octagonal spiral windings are presented.

As in the case with the mutual inductance estimations, when working with applications which utilize rounded rectangular windings or spirals (Fig. 2.1 (f) and [48]), or half- and quarter circle geometries [47] however, these methods can no longer be used. In these cases a more generalized numerical implementation is required.

In this section methods are presented for estimating the self-inductance of various types of inductors modeled as filamentary windings. Firstly, a few existing analytical expressions are presented for circular-, hexagonal-, square, and octagonal spiral windings are presented. Secondly, a more general analytical self-inductance expression, for modeling more complex inductor geometries is presented. Thirdly, this analytical expression is converted into a numerical expression which can be implemented as an algorithm in a computer program to estimate the self-inductance of these various complex planar inductor shapes. Finally, the presented numerical implementation is verified through various experiments. The results are compared against magneto-static FEM simulations, results found in literature, as well as measurements, and show excellent agreement with a maximum absolute error of less than 10 %.

2.7.1 Analytical Solution

The analytical solution for calculating the self-inductance of an inductor modeled as a filamentary structure is derived from the mutual inductance models presented in section 2.6. The self-inductance of a winding can be estimated by evaluating (2.27) along the contour of the primary winding, instead of the secondary winding. Equation (2.27) can be thus rewritten to estimate the self-inductance as:

$$L = M(l, l) = \frac{\mu_0}{4\pi} \oint_{l} \oint_{l} \frac{dl \cdot dl}{|r - r'|}.$$  \hspace{1cm} (2.33)

Here, as shown in Fig. 2.17, $l$ is the contour, and $dl$ an infinitesimally small integration element on the primary filamentary structure. $r$ and $r'$ are two vectors pointing from the axis origin to $l$.

Inherent to (2.33) are singularities at the points where $|r - r'| = 0$. When evaluating this expression with the singularities included, the inductance results become infinite and are unusable. In [54] and [55] methods are presented for
solving this problem. The solutions involve separating the integral in (2.33) into two integrals over the same region. The first integral evaluates (2.33) but excludes points within a certain distance $\Delta_X$, from the singularities. The second integral, evaluates (2.33) within the distance $\Delta_X$, and due to the very small values of $|r - r'|$, is usually approximated by another solvable function.

Equation (2.33) can be rewritten to manage the singularities as [54]:

$$L = \frac{\mu_0}{4\pi} \oint_1 \oint_1 \frac{dl \cdot dl}{|r - r'|} + \left( \frac{\mu_0}{2} \Delta_X^2 \right), \quad (2.34)$$

for $|r - r'| > \Delta_X$.

Furthermore, this analytical expression can only be solved for relatively few, simple inductor geometries. For more complex geometries, the numerical implementation of (2.34) should be used.

### 2.7.2 Numerical Implementation

The numerical solution for estimating the self-inductance of planar PCB inductors is based on the solution for the mutual inductance as presented in subsection 2.6.2. The self-inductance of a inductor modeled as a filamentary structure can be estimated as:

$$L' = M'(l, l) = \frac{\mu_0 \Delta^2}{2\pi} \sum_{\alpha=1}^{\alpha^l} \sum_{\beta=1}^{\beta^l} \sum_{\chi=0}^{X} \sum_{\xi=0}^{E} \frac{n_\alpha^l \cdot n_\beta^l}{|K_\chi - K_\xi|}, \quad (2.35)$$

where
\[ K_\chi = q_\alpha^l + (\chi)(\Delta l) \cdot n_\alpha^l, \]
\[ K_\zeta = q_\beta^l + (\zeta)(\Delta l) \cdot n_\beta^l, \]
\[ X = c_\alpha^l/\Delta l, \quad E = c_\beta^l/\Delta l. \]

Here, all the variables are the same as in subsection 2.6.2, with \( L' \) as the inductor self-inductance.

The numerical results for the self-inductance calculations, also contains singularities which can be managed in the same manner as the analytical expression.

### 2.7.3 Experimental Validation

The numerical method for estimating the self-inductances of the planar PCB inductors, as presented in the previous subsection, is validated through a few experiments performed on the six test inductors \( W_1 \) through \( W_6 \) (Fig. 2.1 in section 2.2). Firstly, their individual inductances are measured using an Agilent 4294A Precision Impedance Analyzer (40 Hz - 110 MHz). The inductances are then estimated using (2.35) as well as magneto-static FEM simulations. Finally, the self-inductances are calculated using the analytical methods presented in [52].

Table 2.4 shows the measured, numerically calculated, FEM simulated, as well as the analytically estimated inductance values, and the absolute errors. The measured inductance values are the average values measured in the frequency range of 1 kHz to 500 kHz. Here, the abbreviation “n/a” (meaning: not applicable) indicates that the inductance method in [52] could not be used for the specific inductor.

<table>
<thead>
<tr>
<th>Winding</th>
<th>Measured inductance ([\mu H])</th>
<th>Calculated inductance (this work) ([\mu H])</th>
<th>Abs. error ([100%])</th>
<th>(\Delta l) ([\text{mm}])</th>
<th>Estimated inductance (FEM) ([\mu H])</th>
<th>Calculated inductance (using [52]) ([\mu H])</th>
</tr>
</thead>
<tbody>
<tr>
<td>( W_1 )</td>
<td>5.98</td>
<td>5.47</td>
<td>8.53 %</td>
<td>0.25, 0.25</td>
<td>5.82</td>
<td>5.03</td>
</tr>
<tr>
<td>( W_2 )</td>
<td>2.17</td>
<td>2.06</td>
<td>5.07 %</td>
<td>0.50, 0.50</td>
<td>1.98</td>
<td>2.07</td>
</tr>
<tr>
<td>( W_3 )</td>
<td>17.3</td>
<td>16.3</td>
<td>5.78 %</td>
<td>0.50, 0.50</td>
<td>17.6</td>
<td>n/a</td>
</tr>
<tr>
<td>( W_4 )</td>
<td>7.16</td>
<td>6.62</td>
<td>7.54 %</td>
<td>0.25, 0.25</td>
<td>6.68</td>
<td>6.00</td>
</tr>
<tr>
<td>( W_5 )</td>
<td>2.58</td>
<td>2.42</td>
<td>6.20 %</td>
<td>0.25, 0.25</td>
<td>2.49</td>
<td>2.60</td>
</tr>
<tr>
<td>( W_6 )</td>
<td>4.39</td>
<td>4.19</td>
<td>4.56 %</td>
<td>0.25, 0.25</td>
<td>4.32</td>
<td>n/a</td>
</tr>
</tbody>
</table>
2.7.4 Conclusions

In this section, methods for estimating the self-inductances of planar PCB inductors are given. The inductances of the six test windings in Fig. 2.1 are measured, and the results are compared against results obtained from the numerical method presented in this work. The results show relatively good agreement with absolute errors smaller than 10%. The results also show good agreement with the FEM simulations. The accuracy of the results from [52] are disputable, which a few accurate and a few inaccurate results (especially regarding the hexagon spiral windings).

2.8 Magnetic Field Intensity Estimation

A magnetic field is a vector field which is created by permanent magnets and electric currents. In the case of planar PCB inductors used in CET applications, the currents flowing through the inductor windings create magnetic fields, which spread through the air around them.

The term magnetic field is often used for two different vector fields, the $B$-field, also known as the magnetic flux density, and the $H$-field, also called the magnetic field intensity. The $B$-field and the $H$-field relates to each other through the permeability of the magnetic medium $\mu$, as shown in section 2.4. If the magnetic medium is free-space, then $\mu$ is linear, and the two fields can be used interchangeably. The $B$-field is measured in tesla [T], and the $H$-field is measured in ampere-per-meter [A/m].

Estimating the magnetic field, and more specifically, the distribution of the magnetic field, is important when designing planar PCB inductors. By studying the distribution of the magnetic field intensity around the inductors, areas with high field intensities can be determined. It is in these areas where a secondary inductor will experience an increased flux linkage, and hence, a higher mutual inductance, which is beneficial for the power transfer between the two inductors.

Magnetic field calculations can also be used when evaluating the risk posed to humans, when they are in close proximity to the magnetic fields generated by these inductors. Due to the watery (and conductive) content of the human body, exposure to alternating magnetic fields can induce eddy currents in body tissue, resulting in local heating, and possible damage of cells inside the body. Certain regulations exist which limit the amount of magnetic (and electromagnetic) radiation that humans can safely be exposed to. By estimating the magnetic field intensity in the areas around inductors, these dangers can thus be assessed.

Magnetic field calculations are based either on Ampere’s law (2.5) or the Biot-Savart law [38]. Both methods have their advantages and disadvantages, and the choice depends mostly on the application at hand.

Ampere’s law on the one hand, is useful when calculating magnetic fields in structures with obvious symmetry. When calculating the field inside a solenoid, or the magnetic field in and around circular conductors of infinite length, this method is best suited. This method can be used as part of a magneto-static
environment or as part of Maxwell’s complete dynamic equations.

The Biot-Savart law, on the other hand, is useful as a “brute force” method for calculating the magnetic field from wires of finite length and shape. This method is best suited for calculating the magnetic field intensity from planar PCB inductors, but can only be used in a magneto-static system.

In this section methods are presented for estimating the magnetic field and magnetic field intensity produced by planar PCB inductors modeled as filamentary structures. Firstly, a analytical expression for estimating the magnetic field intensity from current filaments are derived. Secondly, a numerical implementation for estimating the fields created by planar PCB inductors are presented. Finally, the proposed numerical implementation is verified through two sets of experiments. The results are compared against FEM simulations as well as results obtained through measurements. All results show excellent agreement.

2.8.1 Analytical Solution

The analytical solution for estimating the magnetic field intensity from planar PCB inductors starts from the Biot-Savart law, which is given as [38]:

\[ H = \frac{1}{4\pi} \int_{V'} \frac{J(r') \times u_{r'r}}{|r - r'|^2} \, dv'. \]  

(2.36)

Here, as shown in Fig. 2.18 (a), \( r \) is the vector pointing from the axis origin to the observation point \( P \), where the magnetic field intensity is evaluated; \( r' \) is the vector pointing from the axis origin to the current density source element; \( u_{r'r} \) is the unit vector pointing from the source to the observation point, and \( dv' \) is the current density volume integration element located at \( r' \).

The magnetic field intensity produced by the current flowing through a current filament is derived by solving (2.36) for a line current. The Biot-Savart integral is analytically completed and written in terms of the three vectors as shown in Fig. 2.18 (b). Here the line current is represented by a vector \( e \), with vector \( f \) pointing from the observer point \( P \), to the start of the line current, and vector \( g \) pointing from \( P \) to the end of the line current. The analytical expression in terms of the three vectors is given as

\[ H = \left( \frac{I_P}{4\pi} \right) \left( \frac{g \times e}{|g \times e|^2} \right) \left( \frac{e \cdot g}{|g|} - \frac{e \cdot f}{|f|} \right). \]  

(2.37)

Equation (2.37) can be used to estimate the magnetic field intensity around a single current filament.

2.8.2 Numerical Implementation

The total magnetic field intensity produced by a multi-filamentary structure is calculated by using the superposition of (2.37), and calculating the magnetic field intensity contribution from each current stick at the observation position \( P \) as:
Figure 2.18: (a) The Biot-Savart vectors connecting the current density element with the observer position. (b) The current filament with its three vectors connecting it to the observation point.

\[
H_T = \left( \frac{I_P}{4\pi} \right) \sum_{n=0}^{a_l-1} \left( \frac{f_{n+1} \times e_n}{|f_{n+1} \times e_n|^2} \left( \frac{e_n \cdot f_{n+1}}{|f_{n+1}|} - \frac{e_n \cdot f_n}{|f_n|} \right) \right). \tag{2.38}
\]

Here, as shown in Fig. 2.19, \(l\) is the planar inductor, modeled as a multi-filamentary structure, with \(a_l\) as the total amount of filaments in the structure. Vectors \(e\), \(f\), and \(g\) are rewritten in terms of the indexed filament \(n\), as \(e_n\), \(f_n\), and \(f_{n+1}\), respectively.

![Figure 2.19: A multi-filamentary structure consisting of three filaments, together with their vectors.](image)

Equation (2.38) can thus be used to calculate the magnetic field intensity created by planar PCB inductors modeled as filamentary structures.

### 2.8.3 Experimental Validation

The method for estimating the magnetic field intensity for planar inductors, as presented in this section, is validated by two sets of experiments.
Experiment Set 1

In the first set of experiments, the distribution of the magnetic field is estimated using (2.38) for test inductor $W_3$ as shown in Fig. 2.1 (e). With the inductor placed at the axis origin in the $xy$-plane, the distribution of the magnetic field $z$-component is estimated in a $xy$-plane parallel to the inductor, at a height of $z = 8$ mm, with a primary current of $I_P = 1$ A. The resultant magnetic field distribution is shown in Fig. 2.20 (a).

The magnetic field distribution is compared against results obtained from FEM simulations using Maxwell 3D version 11 while simulating it in a magneto-static environment. The results are shown in Fig. 2.21 (a).

Finally, the distribution of the magnetic field is estimated through measurements performed on the actual test winding produced on a PCB. The measurements are taken using a Brockhaus 460 gauss meter (Lake Shore Cryotronics, Inc), and the results are shown in Fig. 2.22 (a).

Additionally, a two-dimensional view of the $z$-component of the magnetic field distribution, at the center of the winding, where $x = 0$ mm, is calculated at four different heights. The resultant fields are shown in Fig. 2.23 (a).

Experiment Set 2

In the second set of experiments, the distribution of the magnetic field $z$-component is estimated using (2.38) for the test inductor $W_5$ as depicted in Fig. 2.1 (e). With the inductor placed at the axis origin in the $xy$-plane, the distribution of the magnetic field is estimated in a $xy$-plane parallel to the inductor, at a height of $z = 5$ mm, with a primary current of $I_P = 1$ A. The resultant magnetic field distribution is shown in Fig. 2.20 (b).

![Figure 2.20: Magnetic field distribution for (a) $W_3$ and (b) $W_5$, estimated using the numerical method in (2.38).](image-url)

As in the previous set of experiments, the magnetic field distribution is compared against results obtained from FEM simulations and measurements per-
formed on the actual test winding. The FEM simulation results are depicted in Fig. 2.21 (b), and the measurement results are shown in Fig. 2.22 (b).

![Image of magnetic field distributions](image)

**Figure 2.21:** Magnetic field distribution for (a) $W_3$ and (b) $W_5$ estimated using magneto-static FEM simulations.

![Image of magnetic field measurements](image)

**Figure 2.22:** Magnetic field distribution measurements for (a) $W_3$ and (b) $W_5$.

A two-dimensional view of the $z$-component of the magnetic field distribution, at the center of the winding, where $x = 0$ mm, is calculated at four different heights. The resultant fields are shown in Fig. 2.23 (b).

### 2.8.4 Conclusions

In this section, a method for estimating the magnetic fields from planar PCB inductors is presented. This method is based on the Biot-Savart law, and estimates the magnetic fields created by windings modeled as filamentary structures.

The results obtained from the two sets of experiments show mean absolute differences between the calculated and simulated magnetic field distributions of
2.9 Inductor Parasitic Capacitance

The capacitance of a device is a measure of the amount of electric charge it stores for a given electric potential.

Planar PCB inductors have multiple parallel neighboring copper tracks, which gives rise to inter-winding capacitances and results in an equivalent parasitic capacitance over the inductor terminals ($C_{par}^P$ and $C_{par}^S$ in Fig. 2.2 (b)).

The inductance and parasitic capacitance of the winding forms a parallel resonance circuit which oscillates at a frequency of $f^{SR} = 1/2\pi \sqrt{L_p C_{par}^{P,S}}$. At operating frequencies approaching this self-resonant frequency, the parasitic capacitance causes the input impedance of the planar PCB to rise sharply.

It is thus important to accurately estimate the inductor parasitic capacitance and self-resonant frequency, and make sure that the operating frequency of the CET system is far below this resonance frequency, so that the inductor behaves in the manner it is designed to.

In literature, there exists various different works on calculating the parasitic capacitances of different inductors for power electronics and transformer
purposes, as well as capacitance calculations for planar conductive tracks for high-speed integrated circuit interconnects and high-frequency planar strip waveguides.

In most of the literature, the capacitances are estimated by taking a cross-section of the inductor and modeling it in a two-dimensional plane. Here, the per-unit-length inter-track and track-to-ground capacitances are estimated. Depending on the geometric shape of the inductor, different methods are used to accomplish this. For simple symmetrical geometries, analytical methods can be used [59]. For more complex geometrical, Schwarz-Christoffel transformations are used [60, 61]. Here the complex geometries are mapped through mathematical transformations to more simpler geometries, where they can more easily be solved. In other cases a combination of Coulomb’s law and Gauss’s law is used to estimate the charge densities and total charge on the inductors and in this way determine their capacitances [62, 63, 64, 65].

In this section, a semi-analytical method is presented for calculating the parasitic capacitances of planar PCB inductors. The proposed method is verified through measurements performed on the six test windings and the results show good agreement.

2.9.1 Semi-Analytical Solution

The semi-analytical solution for calculating the parasitic capacitance of planar PCB inductors is derived in two parts.

Firstly, a method for estimating the inter-track capacitance between two inductor tracks is presented. This method models the inductor tracks as two-dimensional rectangular cross-sections of infinite length, and estimates the per-unit-length capacitances using approximations based on Schwarz-Christoffel transformations.

Secondly, the inductor parasitic capacitances are estimated using a distributed capacitance model. Based on the inter-track capacitances, the energy stored in the electrical fields between all neighboring inductor tracks are estimated. By calculating the total amount of energy stored in this way, the equivalent inductor parasitic capacitance is determined.

Estimating Inter-Winding Capacitances

The two-dimensional per-unit-length capacitance between two rectangular cross-sections is estimated through mapping techniques using a Schwartz-Christoffel transformation [56].

Firstly, the partial capacitance between the two tracks, in the presence of a substrate is calculated as:

\[
C_{sub} = \varepsilon_0 \varepsilon_r \frac{K(k')}{2K(k)}.
\]

(2.39)

where,
2.9. Inductor Parasitic Capacitance

\[
k = \frac{\tanh\left(\frac{\pi s}{4h}\right)}{\tanh\left(\frac{\pi s}{4h} + \frac{\pi w}{2h}\right)} \quad \text{and} \quad k' = \sqrt{1 - k^2}.
\]

Here, \(w\) and \(s\) are the track width and the inter-track spacing, respectively, as explained in previous sections, \(\varepsilon_0\) is the primitivity of free-space and \(\varepsilon_r\) is the relative permittivity of the substrate. \(K(\cdot)\) is the complete elliptic integral of the first kind, and \(h\) is the thickness of the PCB substrate, usually between 1 - 2 mm.

Secondly, the partial capacitance between the two tracks, in the absence of the substrate is given as:

\[
C_{\text{air}} = \varepsilon_0 \frac{K(k_o')}{2K(k_o)} \quad \text{(2.40)}
\]

where,

\[
k_o = \frac{s}{2w + s} \quad \text{and} \quad k_o' = \sqrt{1 - k_o^2}.
\]

Finally, the total line capacitance per-unit-length can be calculated as \(C_T = C_{\text{sub}} + C_{\text{air}}\).

**Estimating the Total Inductor Parasitic Capacitance**

Once the value of \(C_T\) is known, the calculation of the parasitic capacitance of a planar PCB inductor can be determined based on the distributed capacitance model (DCM) presented in [57]. This method calculates the energy stored in the electrical fields between the parallel inductor tracks, and then estimates the equivalent inductor parasitic capacitance from the total energy.

Firstly, referring to the planar PCB inductor example in Fig. 2.24, let the lengths of each inductor track be defined as \(l_1, l_2, \cdots, l_n\), where \(n\) is the total amount of tracks. Similarly, let the tracks be labeled \(t_1, t_2, \cdots, t_n\). Now, let \(h_k\) be the ratio of the currently indexed track \(k\), to the total inductor length, so that \(h_k \equiv l_k/\ell_{EQ}\) (\(\ell_{EQ}\) is the total inductor length).

Secondly, apply a voltage source to the inductor terminals, so that \(V_{P+} = V_o\) and \(V_{P-} = 0\). The voltage at the beginning of each track \(V_{k+\text{beg}}\) and the voltage at the end of each track \(V_{k+\text{end}}\) can then be calculated as:

\[
V_{k+\text{beg}} = V_o (1 - (h_1 + h_2 + \cdots + h_{k-1})), \quad (2.41)
\]

\[
V_{k+\text{end}} = V_o (1 - (h_1 + h_2 + \cdots + h_{k-1} + h_k)). \quad (2.42)
\]

The model assumes a constant voltage over each straight track, which can then be calculated as \(V_k = \frac{1}{2}V_{k+\text{beg}} + \frac{1}{2}V_{k+\text{end}}\).

Thirdly, let \(t_k\) and \(t_m\) be two adjacent parallel inductor tracks. Their capacitance can then be estimated using the per-unit-length capacitance \(C_T\) as calculated in the previous subsection, multiplied by the length of the shortest track, so that
\[ C_{k-m} = C_T \cdot \min[l_k, l_m]. \] (2.43)

Here, \( \min[\cdot] \) is a function that selects the smallest value inside the brackets. In this way, the capacitances between all the parallel neighboring tracks in the inductor is estimated.

Fourthly, the electrical energy stored in the electric field between the neighboring tracks is calculated as:

\[ E_{k-m} = \frac{1}{2} C_{k-m} (V_k - V_m)^2. \] (2.44)

Fifthly, the total electrical energy stored in the electrical fields between all the neighboring tracks is calculated and labeled \( E_T \).

Finally, the parasitic capacitance of the planar PCB inductor can be expressed as:

\[ C_{par} = \frac{2E_T}{V_o}. \] (2.45)

Although this method is only demonstrated on a planar PCB inductor with straight tracks, it can also be used for inductors with curved and circular tracks.

### 2.9.2 Experimental Validation

The methods for estimating the inter-winding-, and inductor parasitic capacitances, as presented in this section, is validated through various experiments.

In the first set of experiments, the inter-winding capacitances between two inductor tracks are estimated. Here, (2.40) is used to calculate the capaci-
2.9. Inductor Parasitic Capacitance

stance, $C_{\text{air}}$, between two inductor tracks surrounded by air (when no substrate is present). Both inductor tracks have widths of $w = 0.5$ mm, and the capacitances are calculated while increasing the inter-track spacing $s$, from 1 mm to 15 mm, in 1 mm increments. The results are compared against electro-static FEM simulations performed on inductor tracks with the same widths, but with thicknesses of $d = 5 \, \mu\text{m}$, $d = 35 \, \mu\text{m}$, and $d = 105 \, \mu\text{m}$, respectively. The results are shown in Fig. 2.25. Here it can be seen that the simulation results from FEM closely follow the calculated capacitance values. The maximum error between the calculated, and the thin 5 $\mu$m track, is approx. 1 %. The maximum error between the calculated, and the 35 $\mu$m track, is approx. 5 %, and the maximum error between the calculated, and the 105 $\mu$m track, is about 10 %. The capacitance calculations assume a infinitely thin capacitor track, which is the reason why simulation results on the thin 5 $\mu$m track show the smallest error.

![Figure 2.25: The inter-winding capacitance simulation and calculation results for the first set of experiments.](image)

In the second set of experiments, the inter-winding capacitances between the same two inductor tracks are calculated. Here, (2.39) and (2.40) are used to calculate the total capacitance, $C_T$, between two tracks, in the presence of a substrate. The substrate consists of FR4 material, commonly used in PCB’s, with a relative permittivity of $\varepsilon_r = 4.7$ [58]. The capacitances are calculated while increasing the inter-track spacing $s$, from 1 to 15 mm, in 1 mm increments. The results are compared against FEM simulations performed on inductor tracks with thicknesses of $d = 5 \, \mu\text{m}$, $d = 35 \, \mu\text{m}$, and $d = 105 \, \mu\text{m}$, respectively. The results are shown in Fig. 2.26.

Here the calculated results and FEM simulation results show good agreement. The maximum error between the calculated, and the thin 5 $\mu$m track, is approx. 4.29 %. The maximum error between the calculated, and the 35 $\mu$m track, is
Figure 2.26: The inter-winding capacitance simulation and calculation results for the second set of experiments.

about 5.83 %, and the maximum error between the calculated, and the 105 µm track, is approx. 8.62 %.

In the third set of experiments, the semi-analytical method for estimating the total inductor parasitic capacitances of planar PCB inductors, as presented in this section, is validated through measurements performed on the six test inductors $W_1$ through $W_6$.

Firstly, the individual capacitances are measured using an Agilent 4294A Precision Impedance Analyzer (40 Hz - 110 MHz). However, since the planar PCB inductors are in fact designed to behave as inductors and not capacitors, their capacitances can not be directly measured by the impedance analyzer. Instead, the capacitances are estimated by measuring the inductors’ self-resonating frequencies. This is the frequency at which the inductor and its parasitic capacitance are in resonance, at which point the inductors’ input impedance increases significantly. This subject will be discussed in more detail in the next section.

Unfortunately, since the expected capacitance values are very small, in the range of around 1 pF, it is possible that the inductor’s self-resonating frequency might fall outside of the impedance analyzers’ operational frequency range. To solve this problem, an additional high-frequency capacitor with a known value is placed in parallel with the planar PCB inductor during measurement. The increased capacitance will lower the inductors’ self-resonating frequency, well within the possible measurement range. With the extra added capacitor, with a value of $C_X = 1.083$ pF, the new self-resonant frequency, $f^{SR}$, will thus occur at:

$$f^{SR} = \frac{1}{2\pi \sqrt{(L_P)(C_P^{\text{par}} + C_X)}}.$$  \hspace{1cm} (2.46)
2.9. Inductor Parasitic Capacitance

By measuring the new resonant frequency $f_{SR}^*$, and using the inductance values as measured in section 2.7, the parasitic capacitance $C_{par}$, can be determined. Here $L_P$ is the measured inductance values from Table 2.4.

Table 2.5 show the results of the measured and calculated capacitance values, as well as the new resonant frequency $f_{SR}^*$, and the absolute error.

Table 2.5: Measured and calculated parasitic capacitance values of the six test windings shown in Fig. 2.1.

<table>
<thead>
<tr>
<th>Winding</th>
<th>Self-resonating frequency with added parallel capacitor, $f_{SR}^*$</th>
<th>Measured capacitance</th>
<th>Calculated capacitance</th>
<th>Absolute error [100%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_1$</td>
<td>51.40 MHz</td>
<td>0.52 pF</td>
<td>0.45 pF</td>
<td>13.46 %</td>
</tr>
<tr>
<td>$W_2$</td>
<td>85.60 MHz</td>
<td>0.51 pF</td>
<td>0.58 pF</td>
<td>13.73 %</td>
</tr>
<tr>
<td>$W_3$</td>
<td>25.44 MHz</td>
<td>1.18 pF</td>
<td>1.17 pF</td>
<td>0.85 %</td>
</tr>
<tr>
<td>$W_4$</td>
<td>50.41 MHz</td>
<td>0.31 pF</td>
<td>0.28 pF</td>
<td>9.67 %</td>
</tr>
<tr>
<td>$W_5$</td>
<td>80.00 MHz</td>
<td>0.45 pF</td>
<td>0.49 pF</td>
<td>8.88 %</td>
</tr>
<tr>
<td>$W_6$</td>
<td>50.05 MHz</td>
<td>1.22 pF</td>
<td>1.10 pF</td>
<td>9.94 %</td>
</tr>
</tbody>
</table>

2.9.3 Conclusions

In this section, a method for estimating the parasitic capacitances of planar PCB inductors is presented. This method first calculates the inductor inter-track capacitances using semi-analytical methods, by modeling the inductor tracks as two-dimensional infinite length rectangular cross-sections. Using these results, the inductor parasitic capacitances are estimated from the total energy stored in the electric fields between the inductor tracks.

The calculated and simulated inter-winding capacitances in the first and second sets of experiments show reasonable agreement, with maximum absolute errors of 10 % and 8.62 %, respectively. The calculated and measured parasitic capacitance values of the six test inductors, $W_1$ though to $W_6$, show good agreement, with a maximum absolute error of 14 %.

The method for calculating the inter-track capacitances, models the inductor tracks as infinitely thin conductors. Although this method shows reasonable results for inductor tracks with thicknesses up to 105 µm, it might not be accurate enough for thicker inductors, in which case another method should be used.

The error values in Table 2.5 are not correlated to the specific winding geometric parameters, and are mostly contributed to the inter-winding capacitance modeling errors and measurement errors.
2.10 Inductor Excitation Frequency Limits

In section 2.3 the lumped parameter model of planar PCB inductors is presented, and it is shown that these inductors contain parasitic capacitances in parallel to the main inductances. In section 2.9 a method is presented for estimating these capacitances. Here, the term self-resonating frequency is briefly mentioned, and it is stated that the parasitic capacitances have a frequency dependent effect on the inductors' input impedances.

Furthermore, in section 2.5, the issue of the magneto-static approximation is discussed, and it is shown that a majority of the planar PCB inductor models are based on these approximations. It is also mentioned that these approximations depend on the excitation frequency of the applied current or voltage sources.

From both these sections, it become apparent that the planar PCB inductors can not be excited with an unlimited frequency range, but that there exists some upper frequency limit to which the models can be accurately used. In this section, the two main mechanisms that limit the excitation frequency of planar PCB inductors are discussed. They are, the excitation frequency limit due to the inductor self-resonating frequency, and the frequency limit due to the magneto-static approximation validity.

2.10.1 Inductor Self-Resonating Frequency

The inductance and parasitic capacitance of a planar PCB inductor forms a parallel resonance circuit which resonates at a frequency of \( f_{SR} = \frac{1}{2\pi} \sqrt{\frac{L_P}{C_{par}^P}} \). This is also called the self-resonant frequency of the inductor. At operating frequencies much lower than the self-resonant frequency, the majority of the applied current flows through the inductor tracks, and the effects of the parasitic capacitance may be neglected. At frequencies approaching the self-resonant frequency, however, more current will flow through the capacitive paths and less current through the inductor tracks. At these frequencies, the input impedance of the inductor will increase and not be equal to the impedance calculated from the inductance and resistance. At resonance, the voltage-current transfer function reaches a maximum as energy is circulated between the capacitor and the inductor. At frequencies above the self-resonant frequency, the winding will no longer behave as an inductor, but as a capacitor.

The effects of the parasitic capacitor and the operating frequency on the inductor input impedance can be studied by looking at the inductor input impedance equation which is given as:

\[
Z^{-1} = j\omega C_{par}^P + \frac{1}{R_P + j\omega L_P}
\]  

(2.47)

With further expansion, (2.47) can be written as

\[
Z = k_1 R_P + j\omega k_2 L_P,
\]  

(2.48)
where

\[ k_1 = \frac{1}{(1 - \omega^2 L_P C_P^{\text{par}})^2 + (\omega C_P^{\text{par}} R_P)^2}, \]

\[ k_2 = \frac{1 - \omega^2 L_P C_P^{\text{par}} + \frac{C_P^{\text{par}}}{L_P} R_P^2}{(1 - \omega^2 L_P C_P^{\text{par}})^2 + (\omega C_P^{\text{par}} R_P)^2}. \]

Here, \( k_1 \) and \( k_2 \) are two frequency dependent impedance scaling coefficients.

If \( C_P^{\text{par}} = 0 \), or the parasitic capacitance is neglected in the model, \( k_1 \) and \( k_2 \) will remain one, and will be independent of the signal frequency. The inductor input impedance then becomes \( Z = R_P + j\omega L_P \), as expected.

The effects of the parasitic capacitance on \( k_1 \) and \( k_2 \) can be investigated by calculating their values at different frequencies. For these calculations, the equations for \( k_1 \) and \( k_2 \) is simplified by making \( (\omega C_P^{\text{par}} R_P) = 0 \) and \( \left( \frac{C_P^{\text{par}}}{L_P} R_P^2 \right) = 0 \). These are reasonable approximations, when looking at the values estimated for the planar PCB inductors used in this chapter. From Table 2.6, it can be seen through calculations that \( \omega C_P^{\text{par}} \approx 0 \), for frequencies even as high as ten times the self-resonating frequency. Furthermore, \( R_P C_P^{\text{par}} \ll L_P / R_P \) for all test inductors used in this chapter, which qualifies the second approximation. Fig. 2.27 shows the values for \( k_1 \) and \( k_2 \) as a function of the normalized radial frequency for \( L_P = 1 \) H and \( C_P = 1 \) F.

![Figure 2.27](image)

**Figure 2.27:** The normalized frequency dependent scaling factors, (a) \( k_1 \) and (b) \( k_2 \), as function of the normalized radial frequency.

From Fig. 2.27 it can be seen that at low frequencies, \( k_1 \) and \( k_2 \) are both approximately 1. As the frequencies increase, both \( k_1 \) and \( k_2 \) will increase asymptotically until they becomes infinite at the resonance frequency. Above the resonance frequency, \( k_1 \) will decrease down to zero; \( k_2 \), however, will be negative and increase until it reaches zero. In this frequency range, the inductor acts as a capacitor, as the negative complex impedance suggests.
Exciting the windings with an excitation frequency of approximately 10 % of \( f_{SR} \) will ensure that the parasitic capacitive effects will not influence the inductor impedance values, and that the parasitic capacitances, \( C_{par}^P \) and \( C_{par}^S \), can be removed from the inductor model shown in Fig. 2.2.

As an example, the self-resonance frequencies of the six test windings in Fig. 2.1 are measured and shown in Table 2.6, together with the 10 % self-resonance frequency.

Table 2.6: Inductance, capacitance, and self-resonating frequency of the six test windings in Fig. 2.1.

<table>
<thead>
<tr>
<th>Windings</th>
<th>Inductance ( L_P ) (µH)</th>
<th>Capacitance ( C_{par}^P ) (pF)</th>
<th>Self-resonating frequency ( f_{SR}^P ) (MHz)</th>
<th>( f_{SR}^P \div 10 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>W₁</td>
<td>5.98</td>
<td>0.52</td>
<td>90.25</td>
<td>9.03</td>
</tr>
<tr>
<td>W₂</td>
<td>2.17</td>
<td>0.51</td>
<td>151.29</td>
<td>15.13</td>
</tr>
<tr>
<td>W₃</td>
<td>17.3</td>
<td>1.18</td>
<td>35.23</td>
<td>3.52</td>
</tr>
<tr>
<td>W₄</td>
<td>7.16</td>
<td>0.31</td>
<td>106.83</td>
<td>10.68</td>
</tr>
<tr>
<td>W₅</td>
<td>2.58</td>
<td>0.45</td>
<td>147.71</td>
<td>14.77</td>
</tr>
<tr>
<td>W₆</td>
<td>4.39</td>
<td>1.22</td>
<td>68.77</td>
<td>6.88</td>
</tr>
</tbody>
</table>

2.10.2 The Magneto-Static Approximation Validity

As mentioned before, a magneto-static (or magneto-quasi-static) system refers to a system which is governed by magnetic fields originating from current density elements. In these systems, the electric-and displacement fields and their derivatives are assumed to be negligible.

Similarly, a electro-static (or electro-quasi-static) system refers to a system which is governed by electric fields originating from electric charges. In these systems, the time rate-of-change of the magnetic field is assumed to be zero.

Under these two assumptions, a magneto-static and electro-static system can be viewed as separate identities, which can be solved separately using Maxwell’s electro- or magneto-static approximated laws. Technically, it is only true for truly static fields. In practice however, static fields are still good approximations for systems operating with slow changing fields (quasi-static).

As the frequency of such a system increases and the time rate-of-change of both the electric displacement field and the magnetic field increase, the static field approximations will become more and more inaccurate. As the frequency is further increased the fields can no longer be classified as static but become dynamic. The electric- and magnetic fields can no longer be seen as separate identities, and the full Maxwell’s equations are necessary for solving these electromagnetic wave equations. This is also where the phenomenon of wave propagation comes into effect. An important question rises from these facts: how to determine whether a system can be modeled as a static system? In the context of CET systems,
the models presented in this work, assume a magneto-static regime. If this is no
longer the case, the models will no longer be accurate, and the CET systems
will not work the way they are designed to. In [38], this problem is addressed by
looking at the time rate-of-change of the assumed static fields, and comparing
it against the spatial scale of the system in question. From this, the so called
“error” fields are estimated. Thus, for the static and quasi-static approximations
to be justified, the error fields must be small compared to the static fields.

From [38] the magnetic error field is given as:

$$\frac{H_{\text{error}}}{H} = \frac{\epsilon_0 \mu_o L^2}{\tau^2}. \quad (2.49)$$

Here, $L$ is the spatial scale of the system and $\tau$ is the characteristic time of the
sinusoidal excitation (the reciprocal of the angular frequency).

Consequently, a system can be described as magneto-static only when

$$\frac{L}{c} << \tau, \quad (2.50)$$

where, $c$ is the speed of light.

From (2.49) and (2.50) an expression can be derived for calculating the char-
acteristic frequency, $f^M$, at which the magnetic “error” field will be equal to the
magnetic field,

$$f^M = \frac{c}{2\pi l^{EQ}}. \quad (2.51)$$

In the context of CET systems and their planar PCB inductors, $f^M$ can be
calculated by using the inductor track length $l^{EQ}$ as its spatial length $L$. When
designing planar PCB inductors for CET applications, it is important to make
sure that the excitation frequency is far below the characteristic frequency of the
inductor, to ensure accurate model behavior.

As an example, the characteristic frequencies of the six test windings (Fig. 2.1)
are calculated, together with the frequency at which their magnetic error fields
are approximately 10 %. The results are shown in Table 2.7.

**Table 2.7:** The calculated characteristic frequencies and 10 % error field frequencies
of the six test inductors in Fig. 2.1.

<table>
<thead>
<tr>
<th>Inductor</th>
<th>$f^M$</th>
<th>Frequency for</th>
</tr>
</thead>
<tbody>
<tr>
<td>length $l^{EQ}$</td>
<td></td>
<td>a 10 % “Error” field</td>
</tr>
<tr>
<td>$W_1$</td>
<td>1.64 m</td>
<td>29.11 MHz</td>
</tr>
<tr>
<td>$W_2$</td>
<td>0.86 m</td>
<td>55.52 MHz</td>
</tr>
<tr>
<td>$W_3$</td>
<td>3.68 m</td>
<td>12.97 MHz</td>
</tr>
<tr>
<td>$W_4$</td>
<td>1.43 m</td>
<td>33.39 MHz</td>
</tr>
<tr>
<td>$W_5$</td>
<td>0.88 m</td>
<td>54.25 MHz</td>
</tr>
<tr>
<td>$W_6$</td>
<td>1.24 m</td>
<td>38.51 MHz</td>
</tr>
</tbody>
</table>
The operating frequencies of CET planar inductors modeled by the techniques described in this work, are thus essentially limited by the length of their inductors tracks.

2.11 Discussions and Conclusions

Planar inductors constructed as copper tracks on printed circuit boards are an interesting and often useful alternative to regular inductors. In the context of CET systems, these flat low profile inductors can be used to transfer energy between two devices, through their shared magnetic field, without adding much weight or volume to the devices. It is important to understand how these inductors operate, and how to accurately model them, if they are to be reliably used in CET systems.

This chapter focuses on planar PCB inductors in the context of CET systems, and presents methods for modeling their various lumped parameters. First, the lumped inductor and transformer models are presented. Afterwards, various semi-analytical and numerical methods and procedures are presented for modeling these various important parameters. These include: the DC- and AC-resistances, the self- and mutual inductances, magnetic field estimation and parasitic capacitance modeling. Finally, the two main inductor excitation frequency limiting factors are presented. They are: the frequency excitation limit due to inductor self-resonance frequency and the frequency limit due to the magneto-static approximations.

All the presented methods are validated through measurements, FEM simulations, and also through some existing methods found in literature. All results show good agreement.

The methods presented in this chapter will be used in the rest of the thesis for modeling the planar PCB inductors used in the variable-phase CET platform.
Synthesis of the Variable-Phase CET Platform

“The most exciting phrase to hear in science, the one that heralds the most discoveries, is not ‘Eureka!’ (I found it!) but ‘That’s funny...’ ” - Isaac Asimov (1920 - 1992)

3.1 Introduction

The variable-phase CET platform is a specialized implementation of the CET concept, and is used for transferring power to mobile electronic devices placed on its surface. Planar PCB inductors are used to transfer power between the primary circuits embedded in the CET platform, to secondary circuits inside the mobile devices, through their shared magnetic field. In this way, the various power adaptors and plugs needed to power and recharge these devices can be removed, and the mobile devices can be effortlessly powered by simply placing them on the CET platform. A high degree of freedom regarding the placement and orientation of these devices on the CET platform is obtained through the use of multiple primary inductors arranged in a matrix (or lattice) configuration. Figure 3.1 shows an example of a CET platform, powering various electronic devices without using any cables, adaptors or plugs. Figure 3.2 shows the same CET platform, but in addition, it shows the matrix of primary inductors embedded in the CET platform which is used to transfer power to the mobile devices.

From a consumer point of view, the use of CET for powering various mobile devices will not only simplify the charging process, but will also be more aesthetically pleasing, since no bothersome adaptors, cables, or plugs are used. In this
In literature, various CET charging platforms are designed for exactly this purpose [16, 17, 18]. The variable-phase CET platform however, is different, and contains some unique features that distinguishes it from the existing systems.

Firstly, the variable-phase CET platform has the ability to locate possible CET-enabled mobile devices placed on its surface. In this way, only a few primary inductors, located closest to the mobile devices, need to be activated for power transfer. This increases the efficiency of the system, and also limits the magnetic field to active power transfer locations only. The CET platform can also locate and distinguish between conductive materials (e.g. keys, pens, and coins) and magnetic materials (magnets) placed on its surface. Since these objects could interfere with the normal operation of the system, the CET platform can thus avoid activating the inductors close to these objects.

Secondly, the variable-phase CET platform investigates the efficiency of a
novel method of magnetic field shielding. Existing CET systems use either or both of the two well known magnetic field shielding methods available today. These are: magnetic field shielding through “flux-guiding” [66, 67], where high permeability magnetic materials are used to “guide” the magnetic field around and away from sensitive electronic circuits and components, as well as eddy current cancelation [38, 67], where the currents inside conductive materials (caused by the applied fields) generate magnetic fields which tend to oppose the applied magnetic field. These methods, however, require the use of additional materials to obtain the required shielding effects. The variable-phase CET platform on the other hand, uses the method of destructive wave interference to reduce the stray magnetic fields. This method does not require any additional materials, and uses the existing CET platform inductors to create magnetic fields which counteract the fields created during power transfer.

The purpose of this chapter is to synthesize the variable-phase CET platform. Here the CET platform user requirements are combined with the planar PCB inductor modeling methods presented in Chapter 2, and together with the inductive power transfer details and CET inductor- and platform particulars presented in this chapter, are used to calculate the electrical parameters and inductor geometric- and lumped parameters which will allow for the successful transfer of power to the mobile devices. Here, the load detection process as well as the effects of destructive wave interference of the magnetic fields are calculated and discussed.

### 3.2 Inductive Power Transfer

This section describes the process of inductive power transfer as used in the variable-phase CET platform. Firstly, the basic concept and notation used to describe the power transfer between two inductively coupled circuits are discussed. Secondly, the concept of resonance is introduced. It is shown that a primary and secondary series compensated resonant circuit can be used to increase the power transfer capability and efficiency of such a system. Finally, it is shown that a current controlled primary circuit will remove some of the inherent problems associated with the power transfer circuit, and will also allow for better control of the magnetic fields created by the primary windings.

#### 3.2.1 Power Transfer Basics

The synthesis of the variable-phase CET platform starts by assuming that the CET system will operate in a magneto-static mode (see section 2.5), and that the excitation frequency is much lower than the primary and secondary inductors’ self-resonating frequencies (see section 2.10). In this case, the primary and secondary parasitic capacitances may be neglected, and the circuit in Fig. 2.2 (b) can be completed, so that power is transferred from a voltage source to a load, as shown in Fig. 3.3.
Here, $Z_L$ represents the load, and $V_P$ is the primary voltage source. The voltage over the load is $V_L$ and the induced voltage over the secondary inductor is $V_S$.

Assuming a sinusoidal voltage excitation and steady-state operation, the equations that govern the transfer of power can be written in phasor notation as

$$V_P = j\omega L_P I_P + R_P I_P - j\omega M_{PS} I_S, \quad (3.1)$$

$$V_S = j\omega M_{PS} I_P = j\omega L_S I_S + R_S I_S + Z_L I_S, \quad (3.2)$$

which can be combined to give

$$V_P = \left( j\omega L_P + R_P + \frac{\omega^2 M_{PS}^2}{j\omega L_S + R_S + Z_L} \right) I_P. \quad (3.3)$$

In this research work, another useful method based on equivalent impedances is often used for writing the power transfer equations in (3.1), (3.2), and (3.3). Fig 3.3 can be expressed in terms of impedances as shown in Fig 3.4.

$$Z_S = j\omega L_S + R_S + Z_L. \quad (3.4)$$
3.2. Inductive Power Transfer

The reflected impedance $Z_R$ relates the secondary impedance to the primary circuit as

$$Z_R = \frac{\omega^2 M^2_{PS}}{Z_S},$$  \hspace{1cm} (3.5)

and finally, the complete primary circuit impedance becomes

$$Z_P = R_P + j\omega L_P + Z_R.$$  \hspace{1cm} (3.6)

Both these circuit representations can be used to calculate the power transfer capability and efficiency between two magnetically coupled circuits.

3.2.2 Resonant Circuits

Ideal transformer models are often used as a first approximation to describe the operation principle of the devices. In practice, however, transformers applied in power electronic systems are not ideal because e.g. the magnetic cores are not infinitely permeable, magnetization losses, winding losses, and leakage fluxes are present [68, 69].

Usually three important device characteristics are defined to quantify non-ideal effects in transformers. They are: the voltage regulation, the power efficiency, and the power factor.

Voltage regulation is a measure of the change in the terminal voltage of the transformer (due to voltage dividing action of the winding resistances, leakage inductances, and the load), while power efficiency is a measure of the amount of power loss inside the transformer (due to losses in the windings, and hysteresis and eddy-current losses in the core).

The power factor is defined as the ratio of active power flowing into a device, to the apparent power it draws. Therefore, the power factor related to a transformer is thus highly dependent on the type of load it is connected to. Due to the transformer’s leakage and magnetizing inductances, the device has already intrinsically an inductive behavior. With the addition of extra inductive loads, the power factor can be dramatically decreased. The higher the reactive power required in an application, the higher will be the system losses. As a consequence, unity (or as close as possible) power factor should thus be sought.

Transformers created by air-cored planar PCB inductors, in general, have higher flux leakages and higher winding resistances than their magnetically-cored counterparts, and the non-ideal effects mentioned above strongly influence their behavior.

The voltage regulation and efficiency mainly depend on the design characteristics of the transformers, and can not be compensated for by adding extra passive devices. The power factor of the application, on the other hand, can be increased by adding extra compensation capacitors to the primary and secondary windings. Essentially these capacitors are used to store and supply reactive power to and from the inductors, reducing the amount of reactive power drawn from the power supply.

However, full reactive power compensation - that is, unity power factor – only
occurs in a very narrow frequency band around the natural resonance frequency of a given combination of inductors and capacitors, restricting the operational frequency range of the transformer in an application.

Resonant capacitors are used widely in CET circuits [14, 16], and are used to increase the efficiency and capability of the systems they are used in. Generally, capacitor compensation in the primary circuit is used in order to minimize the VA rating of the power supply, while compensation on the secondary circuit is used to enhance the power transfer capability and efficiency of the system [70, 71].

There exists four basic topologies for capacitive compensation. Figure 3.5 shows the four different topologies, which can be described as follows:

- Firstly is the series compensated primary- and series compensated secondary circuit, which is labeled as “sr,sr”, and shown in Fig. 3.5 (a).
- Secondly is the series compensated primary- and parallel compensated secondary circuit, which is labeled as “sr,pr”, and shown in Fig. 3.5 (b).
- Thirdly is the parallel compensated primary- and series compensated secondary circuit, which is labeled as “pr,sr”, and shown in Fig. 3.5 (c).
- Fourthly is the parallel compensated primary- and parallel compensated secondary circuit, which is labeled as “pr,pr”, and shown in Fig. 3.5 (d).

![Figure 3.5](image)

**Figure 3.5:** (a) A series primary and series secondary compensated circuit. (b) A series primary and parallel secondary compensated circuit. (c) A parallel primary and series secondary compensated circuit. (d) A parallel primary and parallel secondary compensated circuit.

Here, the superscripts, sr and pr, are used to indicate a series or parallel (resonant) compensation capacitor, respectively.
All four topologies have different advantages and disadvantages, and their choice mainly depends on the type of application it is used in. In [71], these advantages and disadvantages are described in more detail.

In general, a series compensated secondary capacitance is used to supply a stable secondary load voltage, where a parallel compensated secondary capacitance is used to supply a stable secondary load current. On the primary side, a series capacitor is used to reduce the primary voltage, while a parallel capacitor is used to give large primary currents.

The series compensated primary and series compensated secondary circuit has two advantages that make it exceptionally useful for use in the variable-phase CET platform. Firstly, the a series compensated secondary circuit has only a reflected real impedance, with no reactive components. This means that the secondary circuit will draw only active power, giving the secondary circuit a unity power factor. Secondly, the primary and secondary resonant frequencies are independent of the mutual inductance and the load resistance, and only depend on the primary and secondary winding inductances and their respective resonant capacitors. For an application such as the CET platform, where the secondary winding position, orientation as well as the load current are variable, it is thus advantageous to have a resonant frequency independent of these parameters. The three remaining topologies all have resonant frequencies that change with the mutual inductance and load currents.

After considering the four different capacitive compensation topologies, together with their advantages and disadvantages, the choice that is best suited for the variable-phase CET platform is the primary series- and secondary series capacitor resonant circuit, as shown in Fig. 3.5 (a).

By choosing the primary resonant frequency and the secondary resonant frequency equal to the nominal circuit operational frequency so that

$$f_{CET} = \frac{1}{2\pi\sqrt{L_P C_P^s}} = \frac{1}{2\pi\sqrt{L_S C_S^s}},$$

(3.7)

the new equations that govern the power transfer between primary and secondary series resonant circuits operating at their resonant frequencies become

$$V_P = R_P I_P - j\omega M_{PS} I_S,$$

(3.8)

$$V_S = j\omega M_{PS} I_P = R_S I_S + Z_L I_S,$$

(3.9)

which is combined to give

$$V_P = \left( R_P + \frac{\omega^2 M_{PS}^2}{R_S + Z_L} \right) I_P.$$

(3.10)

At the resonant frequency, the primary and secondary circuits both act as resistive circuits, giving the complete CET circuit a unity power factor.
3.2.3 Power Transfer Using a Current Source

When the circuit in Fig. 3.3, together with its primary and secondary series resonant capacitors is supplied with a constant voltage source, two inherent disadvantages of the circuit, as described by (3.8), (3.9), and (3.10), comes to light.

Firstly, is the fact that the primary circuit will draw a maximum current when it is not loaded by the secondary circuit. The cause of the primary current increase in (3.10) is due to the inversely proportional relationship between the secondary circuit impedance, $Z_S$, and the reflected impedance, $Z_R$, in (3.5). Practically, this means that a CET platform implemented in this way, will have the highest primary currents, and thus the highest primary winding losses when the platform is not transferring any power, or when no load device is placed on the platform. This is clearly an unacceptable scenario, since it will unnecessarily waste power and dramatically reduce the efficiency of the system.

Secondly, as the secondary load resistance $Z_L$ decreases and the secondary load current $I_S$ increases, the primary current will decrease. This in turn will cause the secondary induced voltage $V_S$, which is partly dependent on the primary current, also to decrease. The secondary induced voltage is thus completely unregulated, and not only dependent on the current frequency $f_{CET}$, and mutual inductance $M_{PS}$, but also on the amount of power drawn by the load at that instant. This is an undesirable characteristic that can make it difficult to guarantee a minimum secondary induced voltage, which mobile devices, finally connected to the secondary circuit, may require.

In order to solve these problems, the primary voltage source must be dynamically adjusted to keep the secondary induced voltage $V_S$, approximately constant, and independent of the secondary current. A dynamic primary voltage source will also be able to limit the primary winding currents when the load device is not drawing any power, or when the load is absent. One elegant solution to this problem is to drive the primary circuit with a dynamically controlled voltage source acting as a current source.

The advantages of the current controlled power transfer can be described shortly in a few steps:

1. When the load device is not drawing any power and no secondary current is flowing ($Z_L \rightarrow \infty$), the primary current can be held constant at $I_P = V_P/R_P$, with the minimum amount of current needed to maintain a specific minimum secondary induced voltage.

2. When the load device starts drawing current, the reflected voltage, $V_R = j\omega M_{PS}I_S$ in (3.8), will also increase. To maintain the constant primary current and constant secondary induced voltage, however, the primary voltage source $V_P$ will be increased.

3. If the CET platform is in the process of transferring power, and the power transfer is abruptly interrupted (the load device is suddenly removed, or switches off) the reflected voltage will drop, causing the primary current to
rise. By controlling the primary voltage (and thus the current), this rise in primary current can be managed and kept relatively constant.

4. In section 2.8 it is shown that a magnetic field is created when a current flows through the windings of an inductor. Furthermore, the purpose of the variable-phase CET platform is to control and minimize the stray magnetic field generated by the windings in the CET platform. In order to control the individual windings’ magnetic fields however, the amplitude and phase of the currents in the primary windings need to be precisely controlled. By dynamically controlling the voltage source, the current amplitude can be controlled, and the attenuation of the stray magnetic field can be achieved.

Using a dynamically controlled voltage source acting as a current source, adds stability and controllability to the variable-phase CET platform. The new current controlled power transfer circuit with its series resonant capacitors, which will be used to model the power transfer between two magnetically coupled circuits in the variable-phase CET platform is shown in Fig. 3.6.

![Figure 3.6: A circuit diagram of inductive power transfer to a load. The circuit incorporates two series resonant capacitors, and is driven by a current source.](image)

In order to dynamically control the voltage source to operate as a current source, however, the voltage amplitude needs to be controllable. To accomplish this, a PI compensator is used. Figure 3.7 shows a block diagram of the PI controller.

Here, $I_{P \text{ref}}$ is the primary reference current, which the PI controller will attempt to follow. The estimated value of the primary current is $I_{P \text{est}}$. The error between the reference- and estimated currents is given as $e$. The PI controller’s proportional gain is given as $K_P$, and the integral gain is $K_I$. The “process” describes the whole primary circuit, where the manipulated variable, $u$, will be used to dynamically adjust the supply voltage amplitude. The disturbance, $\delta$, represents the disturbances in the primary current due to any of the above mentioned reasons.

Physically, the current controller is implemented using various analog and digital components as well as a digital controller. The values of $K_I$ and $K_P$ depend highly on the different components and systems which are used to estimate the primary current, as well as the mechanism it uses to manipulate the primary voltage source. For this reason, the rest of the details regarding the
design of the PI controller will not be discussed here, but in Chapter 4, where the implementation of the variable-phase CET platform is discussed.

### 3.3 CET Platform Windings

At the heart of the variable-phase CET platform lay the primary and secondary inductors which transfer power from the CET platform to the mobile devices through their shared magnetic field. The inductor geometries play a vital role in determining the power transfer capability and efficiency of the CET system.

Generally, these mobile electronic devices are relatively small and compact, and not much space is available for embedding extra inductors inside them, especially not bulky cuboidal inductors.

Inductors produced as copper tracks on printed circuit boards offer a good alternative, since they can be produced on very thin and even flexible substrates, which makes them easier to embed and incorporate into new or existing mobile devices. These windings can also be constructed easily and cheaply using existing PCB manufacturing techniques and facilities.

This section discusses planar PCB inductors, and their limitations due to their planar nature. It is shown how spiral inductors, compared to single turn windings, have unique magnetic field distributions which concentrate the magnetic field above the windings, increasing the mutual inductance between primary and secondary windings. Moreover, it is shown that a matrix of hexagon spiral windings can be used to increase the area of the CET platform, increasing mobile device placement and orientation freedom. Finally, it is shown that for variable-phase operation, that each winding geometry can only support a certain limited set of phases, and that the hexagon spiral winding matrix is best suited for three-phase operations.
3.3.1 Spiral Windings

Due to the low profile nature of PCB inductors, only their planar dimensions can be used to create inductor tracks. Optimal use of the limited surface area is thus important.

Spiral inductors constructed from multiple concentric turns are a popular choice in literature [13, 14, 15, 16, 72]. These inductors concentrate their magnetic fields above the center of the windings, forming triangular-shaped magnetic field distributions above the windings, which have strong z-components (perpendicular to the inductor area). Spiral inductors placed above and in parallel to each other, have an increased flux linkage and mutual inductance, compared to their single turn inductor counterparts. Fig. 3.8 shows a few examples of circular- and circular-spiral windings. Figures 3.8 (a), (b), (c) and (d) show circular windings with, one, five, eleven, and fifteen turns, respectively. Fig. 3.8 (e) shows a three-dimensional representation of a circular spiral winding (without indicating the amount of turns) as used in this thesis.

![Figure 3.8: (a) A top view of a circular winding with 1 turn. Top views of (b) a circular spiral winding (non-filled) with 5 turns, (c) 11 turns, and (d) a completely filled circular spiral winding with 15 turns. (d) A three-dimensional view of a circular spiral winding.](image)

Figure 3.9 shows two-dimensional (xz-plane) cross-sections of the absolute-and z-component of the magnetic field distribution created by circular spirals placed in the xy-plane. The circular windings all have outer radii of 12.5 mm, track widths of 0.5 mm, inter-track spacings of 0.25 mm, and are all excited with 1 A (peak) currents. The magnetic fields are calculated at 2.5 mm above the surface of the windings. Here, N indicates the amount of turns.

From Fig. 3.9 it can be seen that the magnetic field above the winding increases with the amount of winding turns. A higher magnetic field, in turn, will create a larger flux linkage to a secondary winding inside the field, resulting in a higher mutual inductance. The triangular-shape of the magnetic field distributions of the fully turned circular spiral inductors are also apparent in these figures.

For the purpose of this work, only spiral windings with full turns, as shown in Fig. 2.1 (a) and (d), and Fig. 3.8 (d), will be used. This will allow maximum magnetic field coupling between primary and secondary windings.
Figure 3.9: A $xz$-section of the distribution of (a) the magnetic field $z$-component, and (b) the absolute magnetic field, created by a circular spiral winding with different amount of turns, calculated at 2.5 mm above the windings surface. The amount of winding turns is indicated as $N$.

3.3.2 Winding Matrices

The variable-phase CET platform consists of multiple primary windings which gives it a high degree of freedom regarding the placement of CET-enabled mobile devices on the platform surface. Here, multiple primary windings are placed next to each other to form a matrix of windings. Depending on the size of the winding matrix, and the specific implementation of the CET platform, single or multiple mobile devices can be powered, independent of their position or orientation.

Furthermore, in the previous subsection it is shown that spiral windings are good options for the CET platform, and that these windings concentrate their magnetic fields above the center of the windings. These fields however, decrease substantially towards to perimeter of the windings. A winding matrix consisting of spiral windings will thus not have a uniform magnetic field distribution, but instead, a distribution of peaks and valleys, corresponding to the winding centers, and edges, respectively.

To ensure that the CET winding matrix will be able to transfer power to load devices placed anywhere on its surface, the sizes of these valleys of low magnetic fields should be kept as small as possible. This is done by minimizing the distances, and removing any gaps between neighboring windings, by selecting appropriate spiral winding geometries.

Figure 3.10 shows two winding matrices created by pentagon-shaped spiral windings and circular-shaped spiral windings, respectively. Here, neighboring windings do not fit well together, creating gaps in between them.

Figure 3.11 shows winding matrices consisting of triangular-shaped spiral windings, square spiral windings, and hexagon spiral windings, respectively. Here, neighboring windings fit well next to each other, minimizing the gaps between them. These windings are all possible candidates for use in the variable-phase
3.3. CET Platform Windings

**Figure 3.10:** A matrix of (a) pentagon-shaped spiral windings, and (b) circular spiral windings.

**Figure 3.11:** A matrix of (a) triangular spiral windings, (b) square spiral windings, and (c) hexagon spiral windings.

As an example, Fig. 3.12 shows the valleys and peaks in the absolute magnetic field created 5 mm above a matrix of hexagon spiral windings. The primary windings have the same parameters as the fully turned 15-turn winding example shown in Fig. 3.9.

Furthermore, one of the novel aspects of the variable-phase CET platform, is its ability to attenuate the stray magnetic field through the destructive wave interference of its individual winding fields. In section 3.6 it will be shown that field attenuation is more efficient for windings with similar magnetic field distributions, such as the fields created by closely positioned identical windings. To achieve field attenuation, a cluster of primary windings are simultaneously excited, each winding with a specific current phase shift and amplitude. The amount of windings in a winding cluster depends on the specific winding matrix, but in general, a smaller amount of windings inside a cluster is more efficient in attenuating the stray magnetic field (as discussed in section 3.6).

Square and hexagon spiral windings contain \(xy\)-symmetry which make the magnetic field distributions of neighboring windings more similar. Triangular spiral windings on the other hand, contain only \(x\)-symmetry in which case the field
distributions of neighboring windings are less similar, and thus less appropriate for magnetic field attenuation. The triangular spiral winding will thus not be considered for use in the variable-phase CET platform.

Figure 3.13 shows how a matrix of square spiral windings can be used make two-phase and four-phase (as well as nine-phase) CET platforms. In Fig. 3.13 (a), A and B indicate the two current phases, namely $0^\circ$ and $180^\circ$. In Fig. 3.13 (b), A, B, C and D indicate the four current phases, $0^\circ$, $90^\circ$, $180^\circ$, and $270^\circ$, respectively.

Furthermore, Fig. 3.14 shows that a matrix of hexagon spiral windings can be used make three-phase, and seven-phase CET platforms. In Fig. 3.14 (b), A, B, and C indicate the three current phases, $0^\circ$, $120^\circ$, and $240^\circ$, respectively. In
Fig. 3.14 (b), A, B, C, D, E, F, and G indicate the seven current phases, namely 0°, 51.4°, 102.9°, 154.3°, 205.7°, 257.1°, and 308.6°.

Figure 3.14: A (a) three- and (b) seven-phase CET hexagon winding platform, showing the individual winding phases.

For the purpose of this thesis, only the three-phase hexagon spiral winding matrix will be analyzed further.

3.3.3 Selected Primary Winding Matrix Geometry

After considering the facts on inductive power transfer, resonant topologies, and the advantages of a current controlled power supply in section 3.2, together with the particulars on the structure of the power transfer windings and the winding matrix, the following summary about the implementation of the variable-phase CET platform can be made. The CET platform will be implemented using:

- series primary and series secondary resonant capacitors,
- a current controlled voltage source using a PI controller,
- a matrix of fully turned hexagon spiral windings, and
- a three-phase stray magnetic field attenuation methodology.

3.4 CET Platform Power Transfer Synthesis

The first priority of the variable-phase CET platform is to transfer power to CET-enabled mobile devices placed the platform surface.

In this section, the geometrical-, lumped inductor-, as well as the electrical parameters which will allow power to be transferred from a single primary hexagon spiral winding to a load connected to a secondary hexagon spiral winding, within the limits of the user requirements and conditions, are calculated.

Firstly, the various user requirements and conditions regarding the geometry of the windings, as well as the electrical parameters are discussed.
Secondly, the vast combinations of possible winding geometric- and electrical parameters are iterated. Here, the different winding geometries are first converted into lumped inductor parameters using the models presented in Chapter 2.1. Afterwards, power transfer calculations are preformed. Through this process, the combinations of winding and electrical parameters which satisfies all the requirements are determined.

Finally, one combination of primary and secondary hexagon spiral winding geometries, together with their lumped inductor parameters, and electrical parameters are chosen for the variable-phase CET platform.

### 3.4.1 Specifications and Constraints

In Chapter 1, a list of requirements for the variable-phase CET platform is briefly presented. These requirements are given to ensure that the CET platform can reliably perform the task of powering the small mobile electronic devices placed on its surface. These requirements are summed up as:

- $\Upsilon_S \leq 20 \text{ mm}$,
- $V_S \geq 10 \text{ V (RMS)}$,
- $V_L \geq 10 \text{ V (RMS)}$,
- $P_L \in [0 \cdots 8 \text{ W}]$,
- $P_{RP} \leq 1 \text{ W}$,
- $P_{RS} \leq 1 \text{ W}$,
- $h_{PS} \in [1 \text{ mm} \cdots 5 \text{ mm}]$,
- $f_{CET} \leq 5 \text{ MHz}$.

Here, $\Upsilon_S$ is the radius of the secondary hexagon spiral winding. It is limited to 20 mm to ensure that it can fit inside, or be embedded into the casing of most standard mobile phones, and other small mobile devices. The secondary windings’ no-load induced voltage is given as $V_S$, and the voltage over the load is $V_L$, as shown in Fig. 3.6. Under normal operating conditions, these values should always be above 10 V (RMS). This is to ensure that at least 5 - 6 V (DC) will be available to the mobile device connected to the secondary circuit after rectification and voltage conversion is performed. The power delivered to the load is given as $P_L$, and the loss limits for the primary and secondary windings are given as $P_{RP}$ and $P_{RS}$, respectively. The CET platform should be able to transfer up to 8 W to the secondary circuit. This is to ensure that at least 5 - 6 W is available to the connected mobile device, after rectification and voltage conversion losses are taken into account. The power loss in the primary and secondary windings should be kept to a minimum, to avoid excessive heating. The power loss per winding should no exceed 1 W. The air gap size between the primary winding matrix and the secondary winding, $h_{PS}$, should be between 1 mm and 5 mm.
This is to allow for tolerances in the plastic and casing thicknesses, of the devices where the secondary windings is installed in. Here, the subscript denotes the two physically separated windings. Generally, the subscript “PS” is used for the distance between a primary and a secondary winding. For labeled windings, like \( W_A \) and \( W_D \) for example, the subscript “AD” is also commonly used. The maximum operating frequency of the system, \( f_{CET} \), should not exceed 5 MHz, due to possible difficulty in the hardware implementation.

In addition to these requirements, there are two more sets of conditions that need to be taken into account. First is the manufacturing conditions for the hexagon spiral windings. Here, the limits of the PCB milling machine and the available copper sheets used to create the planar PCB inductors are taken into account. These conditions are:

- \( w \geq 0.50 \text{ mm}, \)
- \( s \geq 0.25 \text{ mm}, \)
- \( d \in [35 \mu\text{m}, 105 \mu\text{m}], \)
- \( \lambda_P \in [1, 2], \)
- \( \lambda_S = 1. \)

Here, as mentioned earlier, \( w \) is the winding track width, and \( s \) is the inter-track spacing. At the time of production, only two types of copper laminate sheeting were available. Their thicknesses are \( d = 35 \mu\text{m} \) and \( d = 105 \mu\text{m} \). The primary winding matrix can be produced as single or double sided PCB windings. Here, \( \lambda_P \) is used to indicate the amount of layers. The secondary winding is limited to one layer, so that \( \lambda_S = 1. \)

The second set of additional conditions are related to the maximum frequency at which the inductors can be excited. In section 2.10 it is stated that the planar PCB inductor models are based on approximations which only hold true for a limited excitation frequency range. The limits on the excitation frequency are:

- \( f_{CET} \leq 0.15 \times f_{SR}^P, \)
- \( f_{CET} \leq 0.15 \times f_{SR}^S, \)
- \( f_{CET} \leq 0.15 \times f_{M}^P, \)
- \( f_{CET} \leq 0.15 \times f_{M}^S. \)

The frequency limit due to the self-resonant frequency is given as \( f_{SR} \), and the frequency limit due to the inductor parasitic capacitance is given as \( f_M \).

At this point, it is possible that there exist various different combinations of hexagon spiral winding geometries and electrical parameters which could satisfy the mentioned specifications and conditions, and allow for the successful transfer of power from the primary winding matrix to the secondary winding.
The combinations of variables which allow for the successful transfer of power are obtained by iterating through all the allowed winding parameters. First, the lumped parameters for the windings are calculated, whereafter the power transfer calculations are solved to determine whether power transfer is successful, and whether all the power transfer conditions are met.

### 3.4.2 Winding Radii and Mutual Inductance Investigation

The mutual inductance between the primary and secondary windings is position dependent and changes as the secondary winding is moved around above the primary winding cluster. It is also dependent on the winding radii, as well as the amount of turns in each winding.

From the previous subsection it is shown that a large amount of geometric and electrical variables need to be iterated and calculated, in order to find the combinations of variables which will provide a successful power transfer, within the given constraints. The primary and secondary winding radii are two of these geometric variables.

So far, the only limit on the winding radii is the secondary winding radius, $\Upsilon_S$, which should be smaller or equal to 20 mm. In order to reduce the range of radii to iterate, an investigation is conducted to determine the effects of varying primary and secondary winding radii on the mutual inductance at different winding positions. In this way, combinations of primary and secondary winding radii which give maximum mutual inductance, can be determined.

The influence of the primary and secondary winding radii on the mutual inductance is investigated by calculating the mutual inductance for different winding placements and various winding radii. Fig. 3.15 shows three two-dimensional (top view) illustrations of the three most important secondary winding positions, in regards to the primary winding cluster. The three main secondary winding positions are labeled: Position $P_1$, position $P_3$, and position $P_2$. Position $P_1$ corresponds to a perfect overlap of the secondary winding above the first primary winding (Fig. 3.15 (a)). This is the best winding placement with a minimum winding separation distance. Position $P_3$ corresponds to a secondary winding placement exactly between two neighboring primary windings (Fig. 3.15 (b)). Position $P_2$ corresponds to the worse-case secondary winding placement, exactly at the edge of the primary winding (Fig. 3.15 (c)).

Firstly, for a worse case air gap size of $h_{AD} = 5$ mm, the mutual inductances between the primary winding $W_A$ and secondary winding $W_D$ are calculated by varying the primary winding radii between 5 mm and 30 mm, and varying the secondary winding radius between 5 mm and 20 mm. The calculation are performed at all three positions and the results are shown in Fig. 3.16. The primary and secondary windings have the same track widths of $w = 0.5$ mm and track spacings of $s = 0.25$ mm. Mutual inductance calculation are based on the methods presented in section 2.6.

Here, the mutual inductance results are normalized against the highest result obtained from the worse case secondary winding placement at position $P_2$.

Secondly, for a best case air gap size of $h_{AD} = 1$ mm, the mutual inductances
between the primary winding $W_A$ and secondary winding $W_D$ are estimated by varying the primary winding radii between 5 mm and 30 mm, and varying the secondary winding radius between 5 mm and 20 mm. The calculation are performed at all three positions and the results are shown in Fig. 3.17. Here, the mutual inductance results are again normalized against the highest result obtained from the worse case secondary winding placement at position $P_2$ and air gap size of $h_{AD} = 5$ mm. The circle sectors and segments in Fig. 3.16 and Fig. 3.17 show the regions with the highest mutual inductance values. Regions $A$, $B$, and $C$, show the ranges of highest mutual inductances for positions $P_2$, $P_3$, and $P_1$, for an air gap size of $h_{AD} = 5$ mm, respectively. Similarly, regions $D$, $E$, and $F$, show the ranges of highest mutual inductances for positions $P_2$, $P_3$, and $P_1$, for an air gap size of $h_{AD} = 1$ mm, respectively.

Studying Fig. 3.16 and Fig. 3.17, it becomes clear that the maximum mutual inductances occur at different winding radii for the different positions and air gap sizes. There is thus no single combination of primary and secondary winding
radii which will give a maximum mutual inductance for all the positions. Furthermore, it can be seen that although positions $P_3$ and $P_2$ have relatively low mutual inductance values, that the ranges with the maximum values are relatively closely grouped, compared to position $P_1$. For position $P_1$ however, the maximum mutual inductance values are relatively high, but the maxima occur at higher primary winding radii.

To ensure successful power transfer to all the secondary winding positions, the worse case winding placement, position $P_2$, at an air gap of $h_{AD} = 5$ mm, is first considered. Region $A$ gives the highest mutual inductance values for the worse case winding placement. Region $A$ also gives acceptable mutual inductance values at positions $P_1$ and $P_3$, even though this ranges does not include their respective maximum values, the values are still higher than the maximum value found at position $P_2$. The winding radii corresponding approximately with region $A$, will thus be used further in the synthesis of the variable-phase CET platform. The winding radii constraints thus become:

- $\Upsilon_P \in [10$ mm $\cdots 15$ mm$],$
- $\Upsilon_S \in [17$ mm $\cdots 20$ mm$].$

### 3.4.3 Algorithm for Solving the Power Transfer Calculations

The primary and secondary windings’ geometrical parameters as well as the electrical parameters which allow for the successful transfer of power within the given constraints are determined through a parametric search process. Set$_1$ and Set$_2$ give a list of primary and secondary hexagon spiral winding geometrical parameters which are iterated.
3.4. CET Platform Power Transfer Synthesis

\[
\begin{align*}
\text{Set}_1 = & \left\{ \begin{array}{l}
\Upsilon_P \in [10 \text{ mm} \cdots 15 \text{ mm}], \\
w_P \in [0.5 \text{ mm} \cdots 3 \text{ mm}], \\
s_P \in [0.25 \text{ mm} \cdots 3 \text{ mm}], \\
\lambda_P \in [1, 2], \\
d_P \in [35 \mu\text{m}, 105 \mu\text{m}],
\end{array} \right. \\
\text{Set}_2 = & \left\{ \begin{array}{l}
\Upsilon_S \in [17 \text{ mm} \cdots 20 \text{ mm}], \\
w_S \in [0.5 \text{ mm} \cdots 3 \text{ mm}], \\
s_S \in [0.25 \text{ mm} \cdots 3 \text{ mm}], \\
\lambda_S \in [1], \\
d_S \in [35 \mu\text{m}, 105 \mu\text{m}].
\end{array} \right.
\end{align*}
\]

The parametric search process used to determine the combinations of winding geometric- and electrical parameters which will allow successful power transfer from a single primary winding to a single secondary windings, within the given requirements and constraints is shown in Fig. 3.18. The process can be explained in more detail in the following steps:

1. Iterate through all the geometric variables in Set₁ and in each case create the primary winding \( W_P \).

2. Iterate through all the geometric variables in Set₂ and in each case create the secondary winding \( W_S \).

3. Calculate the length of \( W_P \) and store it in \( l_{EQ}^P \). Calculate the winding inductance \( L_P \) using (2.35), the parasitic capacitance \( C_{par}^P \) using (2.45), the DC-resistance \( R_{DC}^P \) using (2.1), the self-resonating frequency \( f_{SR}^P \), as well as the inductor characteristic frequency \( f_M^P \) using (2.51).

4. Calculate \( l_{EQ}^S \), \( L_S \), \( C_{par}^S \), \( R_{DC}^S \), \( f_{SR}^S \), and \( f_M^S \), for the secondary winding \( W_S \).

5. Calculate the maximum excitation frequency \( f_{max}^CET \) as the minimum value in the series \( \{0.15 \times f_{SR}^P, 0.15 \times f_M^P, 0.15 \times f_{SR}^S, 0.15 \times f_M^S, 5 \text{ MHz}\} \).

6. Place \( W_P \) in the \( xy \)-plane at the axis origin.

7. Place \( W_S \) at the worse case secondary winding position \( P_2 \), at a height of \( h_{PS} = 5 \text{ mm} \) (as described in subsection 3.4.2).

8. Calculate the mutual inductance \( M_{PS} \) using (2.31).

9. Iterate though the possible operating frequencies, so that \( f_{CET} \in [1 \cdots f_{max}^CET] \). If \( f_{max}^CET \) is reached, return to step 1.

10. Calculate the frequency dependent AC-resistances \( R_{AC}^P \) and \( R_{AC}^S \), using (2.23).

11. Calculate the maximum primary and secondary currents from the loss limits, as \( I_{max}^P = \sqrt{P_{R_P}/R_{AC}^P} \) and \( I_{max}^S = \sqrt{P_{R_S}/R_{AC}^S} \), respectively.
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Figure 3.18: A flow diagram illustrating the parametric search process used to find the winding geometric- and electrical parameter which will allow successful power transfer from a single primary to a single secondary winding.

12. Iterate through the possible primary winding currents, so that 
   \[ I_P \in [1 \cdots I_P^{\text{max}}] \]. If \( I_P^{\text{max}} \) is reached, return to step 9.

13. Calculate the absolute secondary induced voltage as 
   \[ |V_S| = 2\pi f_{CET} M_{PS} I_P \]. If |\( V_S \)| ≥ 10 (RMS) then the secondary winding’s no-load voltage is adequate. If this voltage is too small, the primary current should be increased (return to step 12).

14. Attempt to transfer 8 W of power to the load resistance \( Z_L \). By finding the roots of the polynomial \( V_L^2 + V_S V_L - P_L R_S = 0 \), the load voltage can be
determined. The secondary current can then be calculated as $I_S = \frac{P_L}{V_L}$.

15. If $|V_L| \geq 10$ V (RMS) then the power transfer was successful and an adequate load voltage is maintained. If the load voltage is too small, the primary current should be increased (return to step 12).

16. If $I_S \leq I_S^{max}$ then the secondary winding loss is within acceptable limits. If the secondary current is too big, the primary current should be increased (return to step 12).

17. If this point is reached, the present combination of geometric- and electrical parameters are sufficient for successfully transferring 8 W of power from the primary winding to the secondary winding placed at the worse case winding position. All conditions are met and the parameters are stored. Successful power transfer of the maximum 8 W to the worse case winding positions automatically assures successful power transfer for all lower powers and all other winding positions and heights.

The results from the parametric search process is obtained, and show that no solutions exist for the single layer primary and single layer secondary winding combinations ($\lambda_P = 1$, $\lambda_S = 1$). For the double layer primary winding combinations however ($\lambda_P = 2$, $\lambda_S = 1$), solutions are found, of which a summary of the general range of solutions are shown in Table 3.1.

### 3.4.4 The Final CET Platform Topology

After iterating through all the possible winding geometric- and electrical variables, and solving the power transfer equations, the following parameters are chosen for the variable-phase CET platform.

Firstly, primary and secondary windings printed on the thicker 105 µm copper laminates are chosen ($d_P = 105$ µm, $d_S = 105$ µm). In general they have lower resistances then their 35 µm counterparts. Secondly, a nominal frequency of $f_{CET} = 2.777$ MHz is chosen. All though not directly related, this frequency falls within the range of 2.5 MHz to 3.0 MHz of which the FCC and CISPR 15 regulations allow for slightly more “relaxed” limitations for conducted emissions for radio-frequency lighting systems, such as electrode-less fluorescent lamps [73]. The implemented CET platform will contain switching elements and also produce magnetic fields in close vicinity to people and other electronic devices. Compliance with existing safety regulations is thus important for such a device. Finally, primary windings with 12.5 mm radii and secondary windings with 20 mm radii are chosen, both with a 0.5 mm track thickness and 0.25 mm inter-track spacing.

The values for the geometrical parameters of the primary and secondary windings, as well as their lumped parameter are shown in Table 3.2, and the electrical parameters are given in Table 3.3. Primary and secondary hexagon spiral windings with the values shown in these tables, will thus be able to transfer up to 8 W of power from the primary winding to a load connected to the secondary winding, within the set of conditions.
Table 3.1: A summary of the general range of solutions to the parametric search process for different track thicknesses.

<table>
<thead>
<tr>
<th>Copper thicknesses</th>
<th>Winding and electrical parameters</th>
</tr>
</thead>
</table>
| $d_P = 35 \, \mu m$, $d_S = 35 \, \mu m$ | $\Upsilon_P \in [10 \, \text{mm} \cdots 15 \, \text{mm}]$  
$\Upsilon_S \in [18 \, \text{mm} \cdots 20 \, \text{mm}]$  
$f_{CET} \in [3.55 \, \text{MHz} \cdots 5.00 \, \text{MHz}]$  
$I_P \in [0.59 \, \text{A} \cdots 1.60 \, \text{A}]$  
$I_S \in [0.52 \, \text{A} \cdots 0.8 \, \text{A}]$  
$R_{AC}^P \in [0.77 \, \Omega \cdots 5.65 \, \Omega]$  
$R_{AC}^S \in [1.81 \, \Omega \cdots 2.27 \, \Omega]$  
$\eta \sim 70.3\% \text{ to } 75.30\%$ |
| $d_P = 105 \, \mu m$, $d_S = 35 \, \mu m$ | $\Upsilon_P \in [10 \, \text{mm} \cdots 14 \, \text{mm}]$  
$\Upsilon_S \in [17 \, \text{mm} \cdots 20 \, \text{mm}]$  
$f_{CET} \in [2.49 \, \text{MHz} \cdots 5.00 \, \text{MHz}]$  
$I_P \in [0.76 \, \text{A} \cdots 2.32 \, \text{A}]$  
$I_S \in [0.44 \, \text{A} \cdots 0.80 \, \text{A}]$  
$R_{AC}^P \in [0.37 \, \Omega \cdots 3.43 \, \Omega]$  
$R_{AC}^S \in [1.58 \, \Omega \cdots 2.27 \, \Omega]$  
$\eta \sim 72.4\% \text{ to } 76.6\%$ |
| $d_P = 35 \, \mu m$, $d_S = 105 \, \mu m$ | $\Upsilon_P \in [10 \, \text{mm} \cdots 15 \, \text{mm}]$  
$\Upsilon_S \in [17 \, \text{mm} \cdots 20 \, \text{mm}]$  
$f_{CET} \in [3.25 \, \text{MHz} \cdots 5.00 \, \text{MHz}]$  
$I_P \in [0.59 \, \text{A} \cdots 1.67 \, \text{A}]$  
$I_S \in [0.51 \, \text{A} \cdots 0.79 \, \text{A}]$  
$R_{AC}^P \in [0.77 \, \Omega \cdots 5.65 \, \Omega]$  
$R_{AC}^S \in [1.13 \, \Omega \cdots 1.46 \, \Omega]$  
$\eta \sim 74.1\% \text{ to } 77.1\%$ |
| $d_P = 105 \, \mu m$, $d_S = 105 \, \mu m$ | $\Upsilon_P \in [10 \, \text{mm} \cdots 15 \, \text{mm}]$  
$\Upsilon_S \in [17 \, \text{mm} \cdots 20 \, \text{mm}]$  
$f_{CET} \in [2.25 \, \text{MHz} \cdots 5.00 \, \text{MHz}]$  
$I_P \in [0.76 \, \text{A} \cdots 2.37 \, \text{A}]$  
$I_S \in [0.43 \, \text{A} \cdots 0.79 \, \text{A}]$  
$R_{AC}^P \in [0.36 \, \Omega \cdots 3.44 \, \Omega]$  
$R_{AC}^S \in [0.90 \, \Omega \cdots 1.45 \, \Omega]$  
$\eta \sim 74.6\% \text{ to } 77.9\%$ |

The mutual inductance values between a primary and the secondary hexagon spiral winding placed above and in parallel to each other, are calculated at the minimum and maximum height separations of $h_{AD} = 1 \, \text{mm}$ and $h_{AD} = 5 \, \text{mm}$, respectively, and is shown in Fig. 3.19.
Table 3.2: The primary and secondary hexagon spiral winding geometrical parameters, as well as their lumped parameters.

<table>
<thead>
<tr>
<th>Primary winding</th>
<th>Value</th>
<th>Secondary winding</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Upsilon_P$</td>
<td>12.5 mm</td>
<td>$\Upsilon_S$</td>
<td>20 mm</td>
</tr>
<tr>
<td>$w_P$</td>
<td>0.5 mm</td>
<td>$w_S$</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>$s_P$</td>
<td>0.25 mm</td>
<td>$s_S$</td>
<td>0.25 mm</td>
</tr>
<tr>
<td>$\lambda_P$</td>
<td>2</td>
<td>$\lambda_P$</td>
<td>1</td>
</tr>
<tr>
<td>$d_P$</td>
<td>105 $\mu$m</td>
<td>$d_S$</td>
<td>105 $\mu$m</td>
</tr>
<tr>
<td>$l_{EQP}$</td>
<td>1.16 m</td>
<td>$l_{EQS}$</td>
<td>1.43 m</td>
</tr>
<tr>
<td>$L_P$</td>
<td>5.72 $\mu$H</td>
<td>$L_S$</td>
<td>7.16 $\mu$H</td>
</tr>
<tr>
<td>$R_{PC}^D$</td>
<td>190 m$\Omega$</td>
<td>$R_{SC}^D$</td>
<td>505 m$\Omega$</td>
</tr>
<tr>
<td>$R_{PC}^A$</td>
<td>2.30 $\Omega$</td>
<td>$R_{SC}^A$</td>
<td>1.95 $\Omega$</td>
</tr>
<tr>
<td>$C_{P_{par}}$</td>
<td>6.02 pF</td>
<td>$C_{S_{par}}$</td>
<td>0.31 pF</td>
</tr>
<tr>
<td>$C_{P_{sr}}$</td>
<td>574 pF</td>
<td>$C_{S_{sr}}$</td>
<td>458 pF</td>
</tr>
<tr>
<td>$f_{M_P}^C$</td>
<td>41.21 MHz</td>
<td>$f_{M_S}^C$</td>
<td>33.39 MHz</td>
</tr>
<tr>
<td>$f_{SR_P}^C$</td>
<td>27.13 MHz</td>
<td>$f_{SR_S}^C$</td>
<td>106.83 MHz</td>
</tr>
</tbody>
</table>

Table 3.3: The variable-phase CET platform electrical parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{CET}^C$</td>
<td>2.777 MHz</td>
</tr>
<tr>
<td>$I_{max_P}$</td>
<td>933 mA (RMS)</td>
</tr>
<tr>
<td>$I_{max_S}$</td>
<td>716 mA (RMS)</td>
</tr>
</tbody>
</table>

The mutual inductance values between the windings in the primary winding cluster, as shown in Fig. 3.14 (a) are calculated and shown in Table 3.4.

Table 3.4: The mutual inductance values between the windings of a primary winding cluster.

<table>
<thead>
<tr>
<th>Mutual inductance</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$M_{AB}$</td>
<td>-0.278 $\mu$H</td>
</tr>
<tr>
<td>$M_{AC}$</td>
<td>-0.278 $\mu$H</td>
</tr>
<tr>
<td>$M_{BC}$</td>
<td>-0.278 $\mu$H</td>
</tr>
</tbody>
</table>

The mutual inductance values are calculated for the three secondary winding positions, $P_1$, $P_2$, and $P_3$, at the minimum and maximum winding separation heights of $h_{AD} = 1$ mm and $h_{AD} = 5$ mm, as shown in Fig. 3.15. The results are displayed in Table 3.5.
Chapter 3. Synthesis of the Variable-Phase CET Platform

Figure 3.19: The mutual inductance distribution between a primary and secondary hexagon spiral winding at height separations of (a) $h_{AD} = 1$ mm and (b) $h_{AD} = 5$ mm, respectively.

Table 3.5: The mutual inductance values between a primary winding, and a secondary winding placed at different heights and positions above it.

<table>
<thead>
<tr>
<th>Height $h_{AD} = 1$ mm</th>
<th>Height $h_{AD} = 5$ mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mutual Inductance</td>
<td>Value</td>
</tr>
<tr>
<td>(Position)</td>
<td></td>
</tr>
<tr>
<td>$M_{AD} (P_1)$</td>
<td>3.94 $\mu$H</td>
</tr>
<tr>
<td>$M_{AD} (P_2)$</td>
<td>1.10 $\mu$H</td>
</tr>
<tr>
<td>$M_{AD} (P_3)$</td>
<td>1.62 $\mu$H</td>
</tr>
</tbody>
</table>

3.5 Load Detection

Load detection is the process in which the variable-phase CET platform attempts to find the positions of load devices placed on the winding matrix surface, and is important for several reasons.

Firstly, as stated in Chapter 1, the variable-phase CET platform consists of a matrix of primary windings and operates by energizing a cluster of three primary windings closest to the secondary device, as to transfer power to it. This is one of the features that makes the variable-phase CET platform different from existing CET platforms. In order for the CET platform to energize the primary winding cluster closest to the load device however, the system must first detect the position of the load device, and determine which primary windings it is closest to.

Secondly, from a practical point of view, a CET platform used in an office environment might also contain objects which are not CET-enabled and should not
be charged. Some of these objects can be metallic in nature, like a bunch of keys, soft-drink cans, pens, coins, etc. Exciting the primary windings close to these conductive objects could create eddy currents and result in undesired heating of the objects. Other objects, like magnets and ferrites, with high permeability, can also interfere with the normal operation of a CET desktop or platform. Finding and disabling the primary windings closest to these objects is thus also important.

In [18], a CET desk with a cord-free power supply is presented. This system employs an unique method for locating load devices placed on the desk surface, and makes use of the existing inductance path between the primary and secondary windings, to send signals that indicate the presence of a secondary load device. This method, however, requires extra circuits and intelligence on the secondary device to work. In CET applications where the size of the secondary winding and circuitry is limited, this might not be a viable solution.

This research work presents a new and novel method of load detection which does not require adding any extra signalling circuitry to the secondary devices. The load detection method proposed here, is done by “scanning” the individual windings while measuring and evaluating their equivalent impedances. In this way the positions of the load devices, as well as the primary windings they are closest to, can be determined. In addition to the load devices, this method can also distinguish between conductive materials and soft-magnetic materials, making it practically quite useful.

The load detection process takes place in three stages, defined as: calibration, load-position detection, and load-type estimation.

### 3.5.1 Calibration

The purpose of calibration is to determine the exact impedances of the individual primary windings when they are not loaded. Due to the specific implementation of the variable-phase CET platform, as well as possible etching mistakes in the winding tracks, the total circuit impedance might not be equal to the calculated or simulated winding impedances. For this reason the variable-phase CET platform is first calibrated. The calibration process estimates the equivalent winding impedances by measuring the primary winding current amplitude, relative current phase, voltage amplitude, as well as voltage phase, for every primary winding, when the CET platform is void of any load devices. In this way, a baseline for comparing future load devices’ influence on the winding impedances can be obtained. For instance, for the case shown in Fig. 3.20 (a), where an arrangement of primary hexagon spiral windings is in a matrix form. The nineteen winding impedances are stored in array $Z_{P}^{CAL}$ as:

$$Z_{P}^{CAL} = (|X_1^C| \angle \phi_1^C, |X_2^C| \angle \phi_2^C, \ldots, |X_{19}^C| \angle \phi_{19}^C).$$

### 3.5.2 Load-Position Detection through “Scanning”

Before the CET platform can power a load device placed on its surface, it must first determine the position of the load device on the surface of the platform, and
establish which primary windings it is nearest to. This is done by “scanning” the CET platform, and involves exciting each primary winding individually for a short duration (typically 1 - 2 seconds each) while measuring and storing its current phase, current amplitude, voltage amplitude, and voltage phase. As in the calibration process, the measurements are converted to impedance values which are stored in array $Z_{PS}^{SCAN}$ as:

$$Z_{PS}^{SCAN} = (|X_1^S|\angle\phi_1^S, |X_2^S|\angle\phi_2^S, \ldots, |X_{19}^S|\angle\phi_{19}^S). \tag{3.12}$$

After the windings are all “scanned”, the measurements are compared against the previously calibrated values. The changes in impedance amplitudes and phases are calculated as $\Delta \phi_n = (\phi_n^S/\phi_n^C - 1)(100)$ and $\Delta X_n = (|X_n^S|/|X_n^C| - 1)(100)$, where $n \in \{1, 2, 3, \ldots, 19\}$. Any deviations outside $\pm 10\%$ of the impedance amplitudes and $\pm 5\%$ of the impedance phases are potential load devices (called the threshold limits). Deviation in impedances in neighboring windings could suggest that a load device is placed between the two windings. Fig. 3.20 (b) shows the positions of possible load devices found after scanning.

A deviation in impedance measurement at winding $W_1$ could suggest a load device placed at position $A$. Deviations in impedance measurements at windings $W_6$ and $W_7$, as well as $W_{16}$ and $W_{19}$ could suggest a load device placed between the windings at positions $B$ and $C$, respectively. Deviations in three neighboring windings, like $W_{13}$, $W_{14}$, and $W_{17}$, could suggest a load device placed in the center of the three windings, at position $D$. In this way, possible load device positions can be determined over the surface of the CET platform.

### 3.5.3 Load-Type Estimation

When the location of the load device has been determined, the primary winding closest to the load device is activated and the process of load-type estimation is started. The current in this winding is systematically increased while the wind-
3.5. Load Detection

The current amplitude, current phase, voltage amplitude, and voltage phase are measured and stored. The measurements are again converted into impedance values. By analyzing the changes in the impedance amplitudes and phases of the loaded primary windings, three different load devices can be distinguished.

**A Valid Resonant Load as a Load Device**

When the load device placed on the CET platform surface is an actual valid resonant load, the circuit equations that govern the transfer of power from the primary winding to the secondary load is given as (3.8), (3.9), and (3.10). A simplified schematic diagram of the circuit is shown in Fig. 3.6.

If the primary and secondary series capacitors are chosen to operate in resonance with their respective windings, (3.8) and (3.9) can be rewritten in terms of the primary impedance as:

\[
Z_P = \frac{V_P}{I_P} = \left( R_P + \frac{\omega^2 M_{PS}^2}{R_S + Z_L} \right).
\]

(3.13)

Here, the variables are the same as described in Fig. 3.6.

Although the mutual inductance and secondary load values will most likely be unknown, it can be seen from (3.13) that \( Z_P \) is real and thus do not induce a extra phase shift between the voltage and current, only the impedance amplitude changes.

**Conductive Material as a Load Device**

When a piece of conductive material (with low permeability), like a piece of copper or aluminum, is placed on the CET platform, the magnetic field produced by the primary windings will create eddy currents in the conductive material. The eddy currents in turn, will create magnetic fields which will oppose the primary field. Although the eddy currents’ sizes are unknown, the conductive material may be modeled as an equivalent inductor with a resistance in series. The circuit equations that govern the transfer of power from the primary winding to the conductive material can be written as:

\[
V_P = j\omega L_P I_P + I_P / j\omega C_P^{sr} + R_P I_P - j\omega M_{PE} I_E,
\]

(3.14)

\[
j\omega M_{PE} I_P = j\omega L_E I_E + R_E I_E.
\]

(3.15)

As shown in Fig. 3.21, \( L_P \) and \( R_P \) are the primary winding inductance and resistance, respectively. The primary resonance capacitor is given as \( C_P^{sr} \), while \( I_P \) is the primary current and \( V_P \) is the fundamental component of the primary switching voltage. The mutual inductance between the primary winding and the conductive material is \( M_{PE} \). The equivalent inductance and resistance of the conductive material, is \( L_E \) and \( R_E \), respectively. The eddy currents are represented as \( I_E \).
With the primary winding and capacitor in series resonance, (3.14) and (3.15) can be rewritten in terms of the primary impedance as:

\[
Z_P = \frac{V_P}{I_P} = \left( R_P + \frac{\omega^2 M_{PE}^2 R_E}{R_E + \omega^2 L_E^2} \right) - j \left( \frac{\omega^3 L_E M_{PE}^2}{R_E + \omega^2 L_E^2} \right) 
\]  
(3.16)

Although the variables \( M_{PE} \), \( L_E \) and \( R_E \) will most likely be unknown, it can be seen in (3.16) that due to the series resonance between \( L_P \) and \( C_P \), the resulting impedance has a negative complex term which forces the current to lead the voltage. In this case the impedance phase will become more negative, and the impedance amplitude might increase.

Soft-Magnetic Material (Ferrite) as a Load Device

When a piece of soft-magnetic material is placed on the CET platform, a new low reluctance path is created for the magnetic field generated by the winding. This will cause an increase in the winding’s own flux linkage and effectively increase its self-inductance. Since the primary resonant capacitor is specifically chosen to compensate the original winding inductance, the capacitive reactance will no longer be sufficient to counteract the new increase inductive reactance, thus leaving the primary circuit more inductive. The current will thus lag behind the primary voltage. The new circuit equation for the primary winding can thus be written as:

\[
V_P = j\omega(L_P + \Delta L_P)I_P + I_P/j\omega C_P^{sr} + R_PI_P. 
\]  
(3.17)

With primary resonance, (3.17) can be rewritten as:

\[
Z_P = \frac{V_P}{I_P} = j\omega(\Delta L_P) + R_P. 
\]  
(3.18)

Here, the variables are still the same as before and \( \Delta L_A \) represents the extra increase in primary inductance. The circuit is now more inductive, and will have a new resonant frequency (different from the CET system frequency). An increase in both impedance amplitude and impedance phase is expected.
3.5.4 Intended Implementation

From the previous subsection it can be deduced that the three different load device types all influence the winding impedances in different ways. These relative changes in winding impedances can thus be used to find the locations of the loads placed on the platform and distinguish between the 3 object types.

The load detection process is implemented in a digital controller, which activates and measures individual windings by using various analog and digital circuits and components. These components measure the winding’s current and voltage amplitudes and phases, and convert the values into impedances, whereafter they are used to located and distinguish the objects types. The implementation of the load detection system, however, will not be further discussed here, but in Chapter 4.

3.6 Stray Field Attenuation through Wave Interference

One of the novel aspects of the variable-phase CET platform is its ability to attenuate the stray magnetic field through the destructive wave interference of its individual winding fields.

Currently, two well known methods of magnetic field shielding exist. First, is “flux-guiding”, where magnetic fields are guided around and away from sensitive electronic parts through low reluctance paths created by the high permeability material placed in the magnetic field [66]. Second, is through eddy current cancellation [38], where the eddy currents (inside conductive material) produced by an alternating magnetic field, create opposing fields which cancel the primary field.

Magnetic field attenuation through destructive wave interference can also be viewed as a form of magnetic field shielding since it actively attenuates the magnetic field. It does not, however, require the presence of any special materials to achieve the shielding effects, but uses exciting windings to actively counteract the fields created when power is transferred from the variable-phase CET platform.

In this section, the concept for wave interference is first discussed, whereafter its applicability on magnetic fields are shown through two examples. This concept is then further extended to include magnetic fields created by hexagon spiral windings on the variable-phase CET platform. Magnetic field values are calculated for typical secondary windings positions and power transfer levels for the three modes of operation: the single-phase mode, the three-phase mode, and the variable-phase mode. Finally magnetic field values are compared and it is shown that three-phase and variable-phase modes are able to attenuate the stray magnetic fields.

3.6.1 Wave Interference and Magnetic Fields

Wave interference occurs when two or more waves superimpose on each other to create a new wave pattern. When two waves of the same frequency are in-phase, they add together constructively to produce a new wave with a larger amplitude. Conversely, when two waves are out-of-phase, they add together
destructively to produce a new wave with a smaller amplitude. Figure 3.22 shows how the interference of two sinusoidal waves, $S_1$ and $S_2$, with the same amplitude, $A$, can be used to increase or decrease the resultant wave ($S_3$) amplitude.

![Figure 3.22: (a) Constructive wave interference of waves $S_1$ and $S_2$ creates a new wave, $S_3$, with a larger amplitude, (b) while the destructive wave interference of the two waves, create a new wave with a smaller amplitude.]

Here, the two sinusoidal waves, $S_1$ and $S_2$, can both be described by

$$y(t) = Ae^{j\omega t + j\theta},$$

where $\omega$ is the radial frequency, and $\theta$ is the phase angle of the waves, respectively. In this case, the amplitude of the waves, $A$, is constant.

Non-static magnetic fields created by alternating currents are also waves, and thus also prone to wave interference. In the case of magnetic fields however, the wave amplitudes are not constant, but position dependent. The Biot-Savart law (2.36) is used to calculate the magneto-static field of a conductor in a three-dimensional space, which gives the field its spatial dependency. The time-varying sinusoidal current flowing through the conductor gives the field its temporal dependency. A magnetic field generated by a steady-state sinusoidal current can thus be written as

$$B(x, y, z, t) = B(x, y, z)e^{j\omega t + j\theta}.$$  \hspace{1cm} (3.20)

Due to the non-uniform distribution of the magnetic field created by hexagon spiral windings, the wave amplitude (or more correctly in this case: the magnetic field distribution) is not constant but position dependent, unlike the two waves, $S_1$ and $S_2$ in the example shown in Fig. 3.22.

The efficiency of destructive wave interference on magnetic fields depends on the similarity of the magnetic field distributions used. For two magnetic fields to cancel out completely, their magnetic field distributions need to be exactly the same. Since the primary windings of the variable-phase CET platform are physically separated by certain physical distances, their magnetic field distributions are not equal (as seen from a fixed observation point). Complete cancelation of the magnetic field through destructive wave interferences is thus not possible.
The amplitude of a magnetic field, however, has a reciprocal dependency to the
distance of its current source. Magnetic field distributions from similarly
shaped and closely positioned windings, when observed from relatively far away,
become more homogeneous, and thus more alike. Conversely, the magnetic field
distributions from the windings, when observed relatively close to one winding,
will be dominated by that individual winding’s distribution.

From these facts it can be concluded that destructive wave interference in
regards to the matrix of hexagon spiral windings will have the following charac-
teristics:

- The magnetic field attenuation efficiency through destructive wave interfer-
ence on magnetic fields will increase with the distance from the windings.

- Due to existing symmetry in the winding matrix, certain areas exist where
two or more portions of the individual magnetic field distributions are equal
in which case complete destructive wave interference in these areas are
possible. These areas are called “dead spots” or “dead areas”.

From these characteristics, it becomes clear that magnetic field shielding
through destructive wave interference operates distinctively different than the
two other methods of magnetic field shielding mentioned earlier. The traditional
magnetic field shielding methodologies make use of physical materials to attenu-
te the fields, so that the magnetic field in the shielded side is smaller compared
to the non-shielded side. Magnetic field shielding through wave interference on
the other hand, only gradually attenuates the magnetic field as the distance
from source increases. Attenuation of the fields close to the windings is thus
not effective, and only stray magnetic fields, far from the sources are effectively
attenuated.

The concept of destructive wave interference regarding magnetic fields is
demonstrated through two examples. The first example is shown in Fig. 3.23 (a).
Here, A, B, and C are three long thin parallel conductors each carrying a current
$I_A$, $I_B$, and $I_C$, respectively. The conductors are placed on the $xy$-plane, with a
separation distance of $r_1$ between them.

Ampere’s law is used to calculate the individual magnetic fields produced by
the three conductors. The resultant absolute magnetic field is calculated at dif-
ferent heights above conductor B in the $z$-direction, for four different conductor
separations of $r_1 = 0.1$ mm, $r_1 = 3$ mm, $r_1 = 6$ mm, and $r_1 = 9$ mm. Firstly,
the resultant fields are calculated for equivalent conductor currents of 1 A (RMS)
with in-phase currents. Secondly, the fields are calculated while each conductor
is excited with a 1 A (RMS) current amplitude, but with 120° phase-shift be-
tween them. The magnetic field results are shown in Fig. 3.23 (b). From these
results it can be seen that the in-phase field results all show similar behavior
at increased heights, which supports the fact that the fields become more ho-
mogeneous far away from the sources. The results of the three-phase fields show
definite attenuation in the magnetic field values. Here, field attenuation increases
for smaller conductor separations, and also for increased distances from the con-
ductors (in both cases where the individual magnetic field distributions become more similar).

The second example is shown in Fig. 3.24 (a). Here the magnetic field attenuation effects from a two-wire two-phase system, is compared against a three-wire three-phase, and a four-wire four-phase system. In all three tests, the long thin parallel conductors are placed in the $xy$-plane, centered around the $y$-axis, with a separation distance of $r_1 = 1$ mm between them. In the two-wire two-phase test, the conductors are excited with $0^\circ$ and $180^\circ$ phase-shifts, respectively. In the three-wire three-phase test, the conductors are excited with $0^\circ$, $120^\circ$, and $240^\circ$ phase-shifts, respectively. In the four-wire four-phase test, the conductors are excited with $0^\circ$, $90^\circ$, $180^\circ$, and $270^\circ$ phase-shifts, respectively. The conductors are all excited with current amplitudes of $1$ A (RMS).

The magnetic field results are shown in Fig. 3.24 (b). Here it can be seen that the two-phase, the three-phase and the four-phase simulations all attenuate the magnetic field, and they all become more effective at increased heights. The two-wire two-phase system show the best field attenuation, followed by the three-wire three-phase system, and finally the four-wire four-phase system. From these results it can be concluded that the amount of conductors and phases used in a system employing destructive wave interference should be kept minimum, for increased magnetic field reduction, at specific heights.

Similar to the two previous examples, the variable-phase CET platform attempts to minimize the stray magnetic field through the process of destructive wave interference. As mentioned earlier, the CET platform operates by exciting a cluster of three primary windings closest to the secondary winding. By conveniently exciting the individual windings with currents of different amplitudes and phase-shifts, the stray magnetic field is attenuated, while simultaneously transferring power to the load device.
3.6. Stray Field Attenuation through Wave Interference

![Figure 3.24](image)

**Figure 3.24:** (a) Long thin parallel conductors spaced $r_1$ from one another in the $xy$-plane. (b) Absolute magnetic field values for the conductors excited with two-phase, three-phase and four-phase currents.

Due to the non-uniform distribution of the magnetic field created by the hexagon spiral windings, the mutual inductances between primary and secondary windings change with their relative positions. The primary and secondary winding currents will thus differ with different power transfer levels and secondary winding placements. Fig. 3.25 (a) shows a three-dimensional image of the primary winding cluster, and the secondary winding during power transfer. To quantify the effectiveness of the destructive wave interference on the magnetic field, five important secondary winding positions are used during the calculations. These positions are showed in Fig. 3.25 (b) and are labeled as $P_1$, $P_2$, $P_3$, $P_4$, and $P_5$.

Furthermore, the magnetic fields are calculated at three different heights above the primary winding cluster. Firstly, the magnetic field is calculated at a height exactly halfway between the top primary winding and the secondary winding. This value is labeled as $B^{close}$, and quantifies the field inside the air gap between the primary and secondary windings. Secondly, the magnetic field is calculated at a height of 10 mm above the secondary winding. This value is labeled as $B^{mid}$, and quantifies the field well inside the mobile device. Thirdly, the stray magnetic field is calculated at a height of 100 mm above the secondary winding. The value is labeled as $B^{stray}$, and quantifies the stray magnetic field. These fields are calculated for the best and worse case secondary winding heights of $h_{AD} = 1$ mm and $h_{AD} = 5$ mm, respectively. Figure 3.26 shows a $yz$-cross-section of the variable-phase CET platform with the secondary winding embedded in a mobile device, showing the primary and secondary windings, and the magnetic field calculation heights.

The variable-phase CET platform is designed to operate in one of three different modes, labeled: single-phase mode, three-phase mode, and variable-phase mode.

In single-phase mode, the currents through the three primary windings are all
in-phase. All magnetic fields undergo constructive interference and no additional reduction in the magnetic field is experienced.

In three-phase mode, the three primary windings in the activated primary winding cluster are all excited with 120° phase-shifted currents of equal amplitude. As discussed previously, areas close to the windings are subjected to minimal destructive wave interference, and power is thus still transferred. At areas further away from the windings, the individual fields destructively interfere
3.6. Stray Field Attenuation through Wave Interference

with each other and a reduction in the resultant magnetic field is experienced.

The variable-phase mode expands on the three-phase concept. The three-phase mode reduces the stray magnetic field through the destructive wave interference of the fields created by the three primary windings, but do not take into account the field produced by the secondary winding during power transfer. The variable-phase mode attempts to further reduce the stray magnetic field by taking into account the field created by the secondary winding, during various secondary winding positions and power transfer levels. It is shown that unique sets of primary current amplitudes and phases exist that can transfer power to the secondary load device, while simultaneously reducing the stray magnetic field produced by the primary and the secondary windings.

3.6.2 “Dead Spot” Detection and Removal

Inherent to CET platforms operating with multi-phased primary winding clusters as, for instance, the variable-phase CET platform operating in three-phase mode, are “dead spots” or positions on the platform where no power transfer occurs [74].

The origin of the “dead spots” and the proposed solution can best be explained by viewing Fig. 3.25 (b). Here, a two-dimensional (top view) of a hexagon spiral winding matrix with the primary cluster windings, \( W_A, W_B, W_C \), together with primary winding \( W_E \), as well as secondary winding \( W_D \), is shown.

During three-phase operational mode, the three primary windings are excited with currents of the same amplitude, but with 120° phase shifts, so that:

\[
I_A(t) = I_o e^{j\omega t + j0}, \quad (3.21)
\]
\[
I_B(t) = I_o e^{j\omega t + j2/3\pi}, \quad (3.22)
\]
\[
I_C(t) = I_o e^{j\omega t - j2/3\pi}. \quad (3.23)
\]

Here, \( I_A(t), I_B(t), \) and \( I_C(t) \), are the currents flowing in the three primary windings, \( W_A, W_B, \) and \( W_C \), respectively.

The voltage induced in the secondary winding, \( W_D \), can be written as

\[
V_D(t) = j\omega I_o e^{j\omega t} \left( M_{AD} + e^{j\pi \frac{2}{3}} M_{BD} + e^{-j\pi \frac{2}{3}} M_{CD} \right), \quad (3.24)
\]

where \( M_{AD}, M_{BD}, \) and \( M_{CD} \) are the mutual inductances between the three primary windings, \( W_A, W_B, \) and \( W_C \), and the secondary winding, \( W_D \), respectively.

From (3.24), the cause of the “dead spots” becomes clear. When the secondary winding is placed so that the individual primary-to-secondary mutual inductances are approximately equal, the induced voltage \( V_D \) will become zero, resulting in no power transfer. Areas around the “dead spots” where the secondary induced voltage is too small to properly power the load devices, are also referred to as “dead areas”. Position \( P_3 \) in Fig. 3.25 (b) is an example of such a “dead spot”.

It is important for the CET system controller to know whether a load device is placed within a “dead area” since a mobile device placed in side such an area can not be powered. Through the use of the load detection process (as described in section 3.5) prior to activating the primary winding cluster, the CET platform controller can determine whether the load device is placed in a “dead area”. A quick method for verifying the presence of a “dead spot” is given in [74], and involves temporarily reducing the current in one of the primary windings to zero. If the load device is indeed inside a “dead area”, the momentary unbalancing of the individually induced voltages will produce a non-zero net secondary voltage. As power is drawn by the secondary load, the flow of secondary current will cause a change in the activated primary windings’ reflected voltages, which can be detected by the CET system controller.

Since the original primary winding cluster will not longer suffice in powering the receiving device, a rearrangement of the primary winding cluster is done by disabling one of the primary windings and exciting one of the adjacent windings with the same phase.

For a secondary winding placed at the “dead spot” at position $P_3$ in Fig. 3.25 (b), the individual mutual inductances will be equal, so that $M_{AD} = M_{BD} = M_{CD}$, and thus $V_D$ is approximately zero. To verify the presence of the “dead spot”, the current in winding $W_C$ is reduced to zero, in which case the secondary induced voltage becomes $V_D = j\omega M_{AD} e^{j\omega t - j\pi/3}$. The “dead spot” is removed by simply disabling winding $W_C$ and energizing winding $W_E$. Since $W_E$ couples very weakly with $W_D$, it does not contribute to the secondary induced voltage or transfer of power to the load device; it does however contribute to the reduction of the stray magnetic field due to the destructive wave interference.

Figure 3.27 shows how the “dead spot” on the variable-phase CET platform is detected and removed. Here the winding matrix is placed in the $xy$-plane, and the secondary winding is moved around above the primary winding cluster at a height of $h_{AD} = 5$ mm. The primary current amplitudes are 0.933 A (RMS). Figure 3.27 (a) shows the secondary induced voltage $V_D$ in the presence of the “dead spot”, and Fig. 3.27 (b) shows $V_D$ when the “dead spot” is removed.

Although the “dead spot” detection and removal process is demonstrated for the variable-phase CET platform operating in three-phase mode, it can also be used for CET platform systems with differently phased primary winding clusters.

### 3.6.3 Single-Phase Operational Mode

The first mode of operation of the variable-phase CET platform, is the single-phase operational mode. During single-phase operation, all the windings inside an activated winding cluster are excited with in-phase currents. Here, no additional reduction of the magnetic field is expected. The power transfer and magnetic field results obtained from the single-phase operation are used as a baseline for comparing the magnetic field attenuation, capability and efficiency of the other operational modes.

The peak magnetic field values are calculated at the five different secondary winding positions, $P_1$ to $P_5$, at the heights of $h_{AD} = 1$ mm, and $h_{AD} = 5$ mm.
Firstly, the single-phase operation with full primary currents is simulated. Here, the three primary winding currents are set to $I_A = I_B = I_C = 0.933 \text{ A (RMS)}$. During each calculation, the magnetic field distributions at the three separate heights above the primary winding cluster are calculated, as described in subsection 3.6.1 and shown in Fig. 3.26. The magnetic field distributions (RMS) are calculated at the “close”, “mid”, and “stray” heights. From these values, the peak values are estimated and shown in Fig. 3.28, Fig. 3.29, and Fig. 3.30, respectively. In these figures, the results are indicated by asterisks.

Secondly, the primary currents which produce the lowest possible magnetic field peaks for the same power transfer, are calculated. These currents differ for each position and power transfer level, and are used to indicate the absolute minimum peak field value that can be obtained for the specific situation. These results are shown in the same figures, and are indicated with circles.

The primary currents, secondary induced voltage and current, as well as the power transfer efficiency of the various calculations will not be presented here, but will instead be shown in Chapter 5, where it will be compared against measured results.

From these results, a few conclusions can be drawn. Firstly, is the fact that the peak magnetic field values from the full primary current calculations, are all relatively flat. This indicates that the magnetic fields, in these cases, are mainly dominated by the currents in the primary windings, and that the currents in the secondary windings do not significantly add to the fields. Secondly, it can be see that from the minimum possible peak magnetic field values, that the fields decrease as less power is transferred to the load. This is to be expected, since less current is drawn by the load, and resultantly less current flows through the secondary winding.

Figure 3.27: The absolute induced voltage in $W_D$ (a) in the presence of a “dead spot” and (b) when the “dead spot” is removed.
3.6.4 Three-Phase Operational Mode

The second mode of operation of the variable-phase CET platform, is the three-phase operational mode. During three-phase operation, the windings inside an activated winding cluster are excited with currents of the same amplitude, but with 120° phase-shifted currents. Here, the phase-shifted currents will destructively add together, leading to attenuation of the stray magnetic field.

As in the previous subsection, the peak RMS magnetic field values are calcu-
3.6. Stray Field Attenuation through Wave Interference

Figure 3.30: Single-phase operation - The peak absolute (“stray”) magnetic field values $B_{stray}$, at 100 mm above the secondary winding, for the different measurement positions and power transfer levels.

lated at the five different secondary winding positions at the heights of $h_{AD} = 1$ mm, and $h_{AD} = 5$ mm, respectively.

At position $P_3$ however, a “dead spot” exists. Here, power can not be transferred from the primary winding cluster to the secondary winding. The “dead spot” removal process, as described in subsection 3.6.2, is used to remove the “dead spot”. When the secondary winding is placed at $P_3$, the primary winding cluster is rearranged, and winding $W_C$ is disabled while winding $W_E$ is enabled.

Firstly, the three-phase operation with full primary currents is simulated. Here the three primary winding currents are set to $I_A = I_B = I_C = 0.933$ A (RMS) with 120° phase shifts between them. During each calculation, the magnetic field distribution at three separate heights above the primary winding cluster are calculated, as described in subsection 3.6.1, and shown in Fig. 3.26. The magnetic field distributions are calculated at the “close”, “mid”, and “stray” heights and the corresponding peak values are shown in Fig. 3.31, Fig. 3.32, and Fig. 3.33, respectively. Here, the results are indicated by asterisks.

Secondly, the primary currents which produce the lowest possible magnetic field peaks for the same power transfer levels are calculated. Again, these currents differ for each position and power transfer level, and are used to indicate the absolute minimum peak field value that can be obtained for the specific situation. These results are shown in the same figures, and are indicated with circles.

From these results it can be seen that the magnetic fields generally decrease as less power is transferred to the load. Also, the field results are not as “smooth” as the results obtained from the single-phase mode. The peaks in the results generally occur at positions $P_3$ and $P_5$, when the secondary winding is either in or close to a “dead spot”.
3.6.5 Variable-Phase Operational Mode

During three-phase operation, the three windings in the activated winding cluster produce 120° phase shifted magnetic fields which attenuate the stray magnetic field. When power is transferred to the secondary load device however, the current in the secondary winding also produce a magnetic field, which adds to the primary fields. This additional secondary magnetic field is not taken into account by the three-phase mode. This can be seen in Fig. 3.33 where the stray
3.6. Stray Field Attenuation through Wave Interference

Figure 3.33: Three-phase operation - The peak absolute (“stray”) magnetic field values $B_{A}^{\text{stray}}$, at 100 mm above the secondary winding, for the different measurement positions and power transfer levels.

magnetic field values during power transfer are generally higher than the fields during low (or no) power transfer.

The third mode of operation of the variable-phase CET platform is the variable-phase operational mode. The variable-phase mode is an extension of the three-phase mode, and attempts to further reduce the stray magnetic field, by taking into account the secondary winding’s magnetic field.

The variable-phase mode, however, operates differently than the previous two modes. Here, depending on the secondary winding placement, either one or two primary windings are used to transfer power to the secondary winding, while the remaining windings are used to actively reduce the stray magnetic fields. When the secondary winding $W_{S}$ is placed at positions $P_{1}$, $P_{2}$, or $P_{4}$ of Fig. 3.25 (b), for example, only a single primary winding, in this case, $W_{A}$ is used to transfer power, while the other two primary windings $W_{B}$ and $W_{C}$ are used to actively reduce the stray magnetic field. When $W_{S}$ is placed at positions $P_{3}$ or $P_{5}$ however, both $W_{A}$ and $W_{B}$ are used to transfer power, while $W_{C}$ is used to reduce the stray field.

For every secondary winding placement and power transfer level, unique current amplitudes and non-120° current phases are calculated which give the minimum peak magnetic fields. These values are calculated by iterating through all possible primary current amplitudes and phases, while calculating the peak stray field values. The values for the primary currents are shown in Table 3.6 and Table 3.7.

As in the previous subsections, the peak RMS magnetic field values are calculated at the five different secondary winding positions at the heights of $h_{AD} = 1$ mm, and $h_{AD} = 5$ mm, respectively.

The “dead spot” at position $P_{3}$ is again removed, as described in subsec-
tion 3.6.2, so that when the secondary winding is placed at $P_3$, the primary winding cluster is rearranged, and winding $W_C$ is disabled while winding $W_E$ is enabled.

The currents are set to the precalculated values and the variable-phase operation is simulated. Again during each calculation, the magnetic field distribution at the three separate heights above the primary winding cluster are calculated, as described in subsection 3.6.1, and shown in Fig. 3.26. The magnetic field distributions are calculated at the “close”, “mid”, and “stray” heights and the corresponding peak values are shown in Fig. 3.34, Fig. 3.35, and Fig. 3.36, respectively. Here, the results are indicated by asterisks. For comparison, the results of the three-phase minimum peak field operation are superimposed on these figures as dashed lines.

Table 3.6: Primary winding cluster current amplitudes and phases for variable-phase operation with $h_{AD} = 1$ mm.

<table>
<thead>
<tr>
<th>$P_L$</th>
<th>$I_A$</th>
<th>$I_B$</th>
<th>$I_C$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_L = 8$ W</td>
<td>425 mA &lt; 0°</td>
<td>325 mA &lt; 213°</td>
<td>325 mA &lt; 213°</td>
</tr>
<tr>
<td></td>
<td>550 mA &lt; 0°</td>
<td>50 mA &lt; 248°</td>
<td>50 mA &lt; 248°</td>
</tr>
<tr>
<td>$P_L = 4$ W</td>
<td>425 mA &lt; 0°</td>
<td>425 mA &lt; 0°</td>
<td>933 mA &lt; 238°</td>
</tr>
<tr>
<td></td>
<td>525 mA &lt; 0°</td>
<td>375 mA &lt; 228°</td>
<td>375 mA &lt; 228°</td>
</tr>
<tr>
<td>$P_L = 1$ W</td>
<td>375 mA &lt; 0°</td>
<td>375 mA &lt; 228°</td>
<td>933 mA &lt; 230°</td>
</tr>
<tr>
<td></td>
<td>150 mA &lt; 0°</td>
<td>225 mA &lt; 233°</td>
<td>225 mA &lt; 234°</td>
</tr>
<tr>
<td></td>
<td>200 mA &lt; 0°</td>
<td>50 mA &lt; 270°</td>
<td>50 mA &lt; 270°</td>
</tr>
<tr>
<td></td>
<td>300 mA &lt; 0°</td>
<td>300 mA &lt; 0°</td>
<td>750 mA &lt; 243°</td>
</tr>
<tr>
<td></td>
<td>375 mA &lt; 0°</td>
<td>275 mA &lt; 228°</td>
<td>275 mA &lt; 228°</td>
</tr>
<tr>
<td></td>
<td>250 mA &lt; 0°</td>
<td>250 mA &lt; 0°</td>
<td>700 mA &lt; 228°</td>
</tr>
<tr>
<td>$P_L = 0$ W</td>
<td>150 mA &lt; 0°</td>
<td>100 mA &lt; 225°</td>
<td>100 mA &lt; 225°</td>
</tr>
<tr>
<td></td>
<td>200 mA &lt; 0°</td>
<td>25 mA &lt; 270°</td>
<td>25 mA &lt; 270°</td>
</tr>
<tr>
<td></td>
<td>200 mA &lt; 0°</td>
<td>200 mA &lt; 0°</td>
<td>450 mA &lt; 190°</td>
</tr>
<tr>
<td></td>
<td>175 mA &lt; 0°</td>
<td>125 mA &lt; 223°</td>
<td>125 mA &lt; 223°</td>
</tr>
<tr>
<td></td>
<td>175 mA &lt; 0°</td>
<td>175 mA &lt; 0°</td>
<td>375 mA &lt; 245°</td>
</tr>
</tbody>
</table>

The results for the “close” magnetic fields show that the variable-phase fields are generally larger than the three-phase fields. For the “mid” magnetic fields, the fields created by the two methods are relatively similar, with a few three-phase field peaks at the $h_{AD} = 5$ mm simulations. The stray magnetic field results however, show a clear reduction in the magnetic fields values especially
3.6. Stray Field Attenuation through Wave Interference

Table 3.7: Primary winding cluster current amplitudes and phases for variable-phase operation with $h_{AD} = 5$ mm.

<table>
<thead>
<tr>
<th>$P_L$</th>
<th>( I_A )</th>
<th>( I_B )</th>
<th>( I_C )</th>
</tr>
</thead>
<tbody>
<tr>
<td>8 W</td>
<td>675 mA &lt; 0°</td>
<td>475 mA &lt; 228°</td>
<td>475 mA &lt; 228°</td>
</tr>
<tr>
<td></td>
<td>800 mA &lt; 0°</td>
<td>75 mA &lt; 255°</td>
<td>75 mA &lt; 255°</td>
</tr>
<tr>
<td></td>
<td>575 mA &lt; 0°</td>
<td>575 mA &lt; 0°</td>
<td>933 mA &lt; 228°</td>
</tr>
<tr>
<td></td>
<td>775 mA &lt; 0°</td>
<td>75 mA &lt; 260°</td>
<td>75 mA &lt; 260°</td>
</tr>
<tr>
<td></td>
<td>550 mA &lt; 0°</td>
<td>550 mA &lt; 0°</td>
<td>933 mA &lt; 225°</td>
</tr>
<tr>
<td>4 W</td>
<td>475 mA &lt; 0°</td>
<td>325 mA &lt; 228°</td>
<td>325 mA &lt; 228°</td>
</tr>
<tr>
<td></td>
<td>550 mA &lt; 0°</td>
<td>50 mA &lt; 250°</td>
<td>50 mA &lt; 250°</td>
</tr>
<tr>
<td></td>
<td>400 mA &lt; 0°</td>
<td>400 mA &lt; 0°</td>
<td>933 mA &lt; 233°</td>
</tr>
<tr>
<td></td>
<td>550 mA &lt; 0°</td>
<td>50 mA &lt; 250°</td>
<td>50 mA &lt; 250°</td>
</tr>
<tr>
<td></td>
<td>375 mA &lt; 0°</td>
<td>375 mA &lt; 0°</td>
<td>933 mA &lt; 215°</td>
</tr>
<tr>
<td>1 W</td>
<td>250 mA &lt; 0°</td>
<td>175 mA &lt; 185°</td>
<td>175 mA &lt; 185°</td>
</tr>
<tr>
<td></td>
<td>350 mA &lt; 0°</td>
<td>25 mA &lt; 183°</td>
<td>25 mA &lt; 183°</td>
</tr>
<tr>
<td></td>
<td>325 mA &lt; 0°</td>
<td>325 mA &lt; 0°</td>
<td>675 mA &lt; 223°</td>
</tr>
<tr>
<td></td>
<td>325 mA &lt; 0°</td>
<td>25 mA &lt; 180°</td>
<td>25 mA &lt; 180°</td>
</tr>
<tr>
<td></td>
<td>300 mA &lt; 0°</td>
<td>300 mA &lt; 0°</td>
<td>575 mA &lt; 180°</td>
</tr>
<tr>
<td>0 W</td>
<td>250 mA &lt; 0°</td>
<td>100 mA &lt; 180°</td>
<td>100 mA &lt; 180°</td>
</tr>
<tr>
<td></td>
<td>350 mA &lt; 0°</td>
<td>25 mA &lt; 180°</td>
<td>25 mA &lt; 180°</td>
</tr>
<tr>
<td></td>
<td>325 mA &lt; 0°</td>
<td>325 mA &lt; 0°</td>
<td>450 mA &lt; 180°</td>
</tr>
<tr>
<td></td>
<td>325 mA &lt; 0°</td>
<td>25 mA &lt; 180°</td>
<td>25 mA &lt; 180°</td>
</tr>
<tr>
<td></td>
<td>275 mA &lt; 0°</td>
<td>275 mA &lt; 0°</td>
<td>420 mA &lt; 180°</td>
</tr>
</tbody>
</table>

at the higher power transfer levels. This means that the variable-phase method is successful in attenuating the stray magnetic fields not only from the primary windings but from the combination of the primary and secondary windings during power transfer as well, as it is intended.

3.6.6 Stray Field Attenuation Results

The effectiveness of the stray magnetic field shielding through three-phase and variable-phase operations are shown in Fig. 3.37. Here, the variable-phase stray magnetic field results and the minimum possible three-phase stray magnetic field results are compared against the minimum possible single-phase stray magnetic field results. The three-phase stray magnetic field attenuation results are shown with circled lines and the variable-phase attenuation results with asterisks.

The results from Fig. 3.37 show definite reduction in the peak absolute stray magnetic fields in both the three-phase as well as the variable-phase modes.

In general, the thee-phase mode show higher attenuation levels for low power transfer levels compared to the higher power transfer levels. This comes from the fact that during the three-phase mode, only the fields from the primary
windings contribute to the destructive wave interference. During power transfer, the current in the secondary winding also contributes to the field. At a few positions, mostly at position $P_3$ at higher power transfer levels, the three-phase field results are actually larger than the single-phase field results, which produces an negative attenuation, and thus an increase in the peak fields.

The stray magnetic field attenuation results from the variable-phase mode show overall improvement from the three-phase results, especially at higher power.
3.6. Stray Field Attenuation through Wave Interference

Figure 3.36: Variable-phase operation - The peak absolute ("stray") magnetic field values $B_{A}^{\text{stray}}$, at 100 mm above the secondary winding, for the different measurement positions and power transfer levels.

Figure 3.37: Stray magnetic field attenuation results for the three-phase and variable-phase operational modes.

transfer levels. This means that the variable-phase mode is successful in incorporating the field produced by the secondary winding during power transfer, in to the destructive wave interference process.

Attenuation results vary widely for the different winding positions and power transfer levels, especially around position $P_3$, due to the “dead spot” removal process. Stray magnetic field attenuation results of between 1.4 dB and 12 dB are obtained for the variable-phase operational mode.
3.7 Discussions and Conclusions

The variable-phase CET platform is a specialized implementation of the CET concept, and is used for transferring power to CET-enabled mobile devices placed on its surface. In this way, bothersome cables, plugs and adaptors, which are normally used to power and recharge these devices can be removed, possibly increasing user comfort. The platform consists of multiple primary inductors arranged to form a matrix of inductors which increases the overall usable platform area. This gives a high degree of freedom regarding the placement and orientation of the CET-enabled devices on the CET platform. The platform operates by activating a cluster of three primary inductors closest to the mobile device. In this way the magnetic fields are limited to areas where power transfer occurs only. The variable-phase CET platform also investigates the efficiency of a novel method of magnetic field shielding through destructive wave interference. This method uses the existing inductors on the CET platform to actively attenuate the stray magnetic field, so that no additional magnetic or conductive materials are needed for shielding.

This chapter focuses on the synthesis of the variable-phase CET platform. Firstly, it is shown that capacitive compensated series resonant primary and secondary circuits and a current controlled primary voltage source can be used to removed inherent disadvantaged of the power transfer process between weakly coupled inductors, increasing the power transfer capability and efficiency. The controllability of the power transfer process is also increased in this way.

Secondly, the planar PCB inductors used on the CET platform are discussed. It is shown that fully turned planar spiral windings shaped as hexagons are best suited for the platform. Spiral windings use their available surface area very effectively and create unique triangular-shaped magnetic field distributions, with strong z-components perpendicular to the windings surfaces. When mobile devices are fitted with similar windings, and placed close to, and in parallel to the platform, the mutual inductance between the windings are relatively large compared to single turned windings. It is also shown that hexagon windings fit very well into a matrix (or lattice) structure, minimizing spaces between the windings. The unique arrangement of windings in the “honey-comb”-like winding lattice, strongly supports a three-phase operational methodology.

Thirdly, a synthesis of the variable-phase CET platform is performed. Here, the user requirements and constraints, together with the planar winding models as presented in Chapter 2, and the power transfer equations are used to simulate and create the CET platform. Through a parametric search process, the set of winding- geometric and lumped parameters as well as the electrical parameters which will allow successful power transfer between a primary and secondary winding are calculated. The CET platform will consist of primary windings with 12.5 mm radii, and secondary windings of 20 mm radii, with an operating frequency of 2.777 MHz.

Fourthly, an unique load position detection scheme is presented. In order for the CET platform to activate the cluster of three primary windings closest to the
CET-enabled load devices placed on the platform surface, the system must first know the relative positions of the devices on top of the platform surface. The load detection process works by “scanning” the CET platform and systematically activating each primary winding while estimating its equivalent impedance. This process can detect and distinguish between valid resonant load devices, conductive materials (keys, pens, soft-drink cans), as well as magnetic materials (ferrites) placed on the CET platform surface. In this way, the windings closest to the mobile devices can be activated, while activation of the windings close to conductive and magnetic materials, which could interfere with the normal operations of the platform, can be avoided.

Fifthly, the magnetic field attenuation through destructive wave interference is discussed. Wave interference occurs when two or more waves superimpose on each other to create a new wave pattern. When the amplitude of the resulting wave pattern is smaller than the primary waves, it is known as destructive wave interference. Non-static magnetic fields are waves, and thus also prone to wave interference. In a quasi-magneto-static environment, the phases of the magnetic fields are the same than the phases of the currents which produce them. By controlling the winding current phases, their individual magnetic fields can thus undergo destructive wave interference, resulting in a reduction in the resultant magnetic field. It is shown that destructive wave interference is more effective for inductors with similar shaped magnetic field distributions, such as closely located and similarly shaped windings. Magnetic field shielding through destructive wave interference however, operates distinctly different than existing shielding techniques, using magnetic- and conductive materials. Magnetic field attenuation through destructive wave interference as implemented in the variable-phase CET platform has the ability to reduced the stray magnetic fields, while the amplitude of the magnetic fields closer to the windings stay relatively unchanged (or in some cases increase).

Sixthly, the stray magnetic field attenuation through destructive wave interference is simulated for the variable-phase CET platform. The platform can operate in one of three different modes, i.e. single-phase mode, three-phase mode, and variable-phase mode. During single-phase mode, all the currents in the primary winding cluster are in-phase, and no additional reduction of the stray magnetic field is expected. During the three-phase mode, the windings in the activated cluster are excited by currents with equal amplitude but with 120° phased shifts. Here the stray magnetic field is attenuated, but mainly for low power transfer levels, since the fields produced by the current in the secondary winding, during power transfer, is not actively attenuated by this process. The variable-phase mode is an extension of the three-phase mode, and attempts to further reduced the stray magnetic field by taking into account the magnetic field created by the secondary winding during power transfer. Unique non-120° phase shifted currents are calculated for various secondary winding positions and power transfer levels. In this way, the variable-phase method further improves the attenuation results, especially at the higher power transfer levels.

Finally, it is shown that both the three-phase and variable-phase modes are
able to attenuate the stray magnetic field. For the higher power transfer levels, the variable-phase mode shows better attenuation results, and for lower power transfer levels, the variable-phase and three-phase mode show similar results. Attenuation values of up to 12 dB are achieved.
Implementation of the Variable-Phase CET Platform

“If we knew what it was we were doing, it would not be called research, would it?” - Albert Einstein (1879 - 1955)

4.1 Introduction

In the previous chapter, a synthesis of the variable-phase CET platform is performed. Firstly, through various calculations and simulations, the geometry of the variable-phase CET platform’s primary winding matrix and secondary windings are estimated. Through an iterative process, the lumped winding parameters and electrical parameters are obtained, which will allow up to 8 W to be transferred from a single primary winding to a load connected to a secondary winding laying on the platform surface. Secondly, the process of load detection is addressed. It is shown that through the process of load detection, the CET platform can locate and distinguish between valid resonant load devices, magnetic materials, and conductive materials placed on the CET winding matrix. Thirdly, a novel stray magnetic field shielding methodology is proposed. It is shown that by exciting the primary windings on the CET platform with currents of various amplitudes and phases, that the peak stray magnetic field can be attenuated, by up to as much as 12 dB.

The results obtained from these various calculations and simulations, however, are still purely theoretical. In order to verify the results through experimental measurements, the variable-phase CET platform needs to be physically implemented in the form of a hardware prototype system.
The purpose of this chapter is to implement the variable-phase CET platform prototype and to discuss, in detail, the various design considerations taken in developing it.

In this chapter, firstly, a system overview of the complete variable-phase CET platform is presented. The CET platform is designed in a modular fashion, with various specialized subsystems, each with a specific function. Here the CET prototype design is broken down into its individual subsystems, and together with their interactions, discussed.

Secondly, the primary-side subsystems and circuits are discussed. These include the primary winding matrix, the primary controller, quadratic buck converters, high-frequency MOSFET half-bridge drivers, half-bridge inverters, half-bridge output voltage amplitude and phase estimators, winding current amplitude and phase estimators, as well as the winding commutation circuits.

Thirdly, the secondary-side subsystems and circuits are discussed. These include the secondary winding, the full-bridge rectifier, the DC-to-DC converter, and the load.

Fourthly, the primary current amplitude and phase controllers are discussed. These compensators are used to control the primary windings’ current amplitudes and phases, which is necessary when the CET platform operates in three-phase and variable-phase mode.

Finally, the implementation of the variable-phase CET platform is concluded. Schematic diagrams and figures of all the implemented subsystems are presented. Also, although this chapter does not deal with measurements and experimental verification (this is addressed in Chapter 5), various measurements are performed on the individual modular subsystems presented in this chapter. In this way the individual parts of the prototype are tested to ensure that they will work as expected within the larger prototype system.

4.2 System Overview

The implementation of the variable-phase CET platform consists of various modular parts referred to as subsystems. Each subsystem has a specific function to perform within the CET platform, and together all the subsystems allow the variable-phase CET platform to operate in the way it is designed to. These subsystems include: the primary hexagon spiral winding matrix, a FPGA controller, quadratic buck converters, high-frequency MOSFET half-bridge drivers, half-bridge inverters, current amplitude-, voltage amplitude-, current phase-, and voltage-phase estimator circuits, winding commutation circuits, as well as the secondary winding, together with its rectifier, DC-to-DC converter and load. The controller houses PI compensators for controlling the primary winding currents, and it also controls the current phases. The CET platform also communicates with a computer using a RS-232 serial connection. The primary controller together with the computer controls all the processes of the variable-phase CET platform, and allows it to transfer power to CET enabled mobile devices placed on the CET platform surface.
4.2. System Overview

The prototype system consists of three CET hardware channels, which allows it to simultaneously excite and control three primary windings at a time. The three hardware channels are labeled: Channel I, channel II, and channel III.

Figure 4.1 shows a block diagram overview of one of the hardware channels on the prototype system. Here, the dark arrows indicate the flow of power, the dashed gray arrows show the interaction of the magnetic fields, and the gray arrows show the control lines.

![Figure 4.1: Block diagram of the CET variable-phase CET platform prototype.](image)

Figure 4.2 shows a simplified schematic diagram of the variable-phase CET platform prototype. Here, the gray arrows show the control lines, and the subscripts I, II, and III are used to indicate the components’ and subsystems’ associated hardware channels. The different subsystems as shown in Fig. 4.2 are briefly described in Table 4.1. The MOSFETs used in the half-bridge inverters are all labeled with a $T$, and subscripted with the associated channel number. Additionally, the low-side MOSFETs are superscripted with asterisks.

A photo of the implemented CET platform prototype is shown in Fig. 4.3.

The CET platform is implemented with a primary hexagon spiral winding matrix consisting of 19 windings labeled $W_A$ through $W_S$, and are connected to the three hardware channels through the winding commutation circuits. The
primary windings are connected to the hardware channels in such a way, that neighboring primary windings are always connected to different channels. The primary windings are connected to the three separate hardware channels, similarly to the three-phase winding placements depicted in Fig. 3.14 (a). The three hardware channels are used to generate the three-phase and variable-phase currents and phases as discussed in Chapter 3. Figure 4.4 shows an illustration of the implemented primary winding matrix with the individually labeled hexagon spi-
Figure 4.3: Photo of the implemented variable-phase CET platform prototype, indicating some of the individual subsystems.
Table 4.1: Descriptions of the subsystems as shown in Fig. 4.2.

<table>
<thead>
<tr>
<th>Subsystem</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>The primary winding commutation circuits.</td>
</tr>
<tr>
<td>B</td>
<td>The primary winding current amplitude and phase estimators.</td>
</tr>
<tr>
<td>D</td>
<td>The half-bridge output voltage phase estimators.</td>
</tr>
<tr>
<td>E</td>
<td>The high-frequency MOSFET drivers.</td>
</tr>
<tr>
<td>F</td>
<td>The quadratic buck converters, and DC-voltage estimators.</td>
</tr>
<tr>
<td>G</td>
<td>The Xilinx FPGA controller.</td>
</tr>
</tbody>
</table>

The operation of one of the channels of the variable-phase CET platform prototype is explained in the following steps:

1. A DC-voltage in the range of 150 V - 200 V, supplied by the power grid, is fed into the quadratic buck converter which is controlled by a 50 kHz,
4.3 The Primary System and Circuits

PWM signal generated by the controller. The DC-voltage output from the converter is thus adjustable.

2. The output voltage from the quadratic buck converter is fed into the half-bridge inverter, which in turn generates a square wave voltage signal of 2.777 MHz. The half-bridge inverter is driven by a specialized high-frequency MOSFET driver circuit.

3. The winding commutation circuit connects the output of the half-bridge inverter to the different primary windings and their associated series resonant capacitors.

4. The series RLC-circuit (primary winding inductance, series capacitance, and resistance) forms a band-pass filter tuned to 2.777 MHz. This allows only the fundamental current component to flow through the windings. The primary winding current is thus sinusoidal.

5. The current amplitude and phase, as well as the half-bridge voltage amplitude and phase are estimated using individual estimator circuits which send the information back to the controller.

6. The FPGA controller measures the current and voltage amplitudes and phases and controls the quadratic buck converters’ PWM signal. It also controls the half-bridge MOSFET drivers’ clock phase in order to control the primary winding current amplitude and phase, as required for single-phase, three-phase and variable-phase CET platform operations.

4.3 The Primary System and Circuits

The primary system and circuits refer to the circuits connected to the variable-phase CET platform.

4.3.1 The Primary Winding Matrix

The primary winding matrix (CET platform) consists of 19 hexagon spiral windings produced as copper tracks on a printed circuit board made from FR4 [58] material. The individual windings have the dimensions and specifications as discussed in subsection 3.4.4 and shown in Table 3.2.

Figure 4.5 shows an image of the implemented CET primary winding matrix. Furthermore, to insure that no electrical contact occurs between the primary winding matrix and a secondary winding placed on the platform, the winding matrix is covered with a thin layer (approx. 1 mm - 1.5 mm) of insulating cardboard. Figure 4.6 shows a picture of the covered CET winding matrix.

The inductances and AC-resistances of the primary windings, as shown in Fig. 4.5, are measured using an Agilent 4294A Precision Impedance Analyzer (40 Hz - 110 MHz). The results are displayed in Table 4.3.
Due to an etching fault during production of the primary winding matrix, winding $W_H$ is not used and not connected to the CET circuits. These values are marked as “n/c” (meaning: not connected) in Table 4.3.

The calculated values for the primary winding inductance and AC-resistance, as shown in Table 3.2, are 5.72 $\mu$H and 2.30 $\Omega$, respectively.
Table 4.3: Measured primary winding matrix inductances and AC-resistances.

<table>
<thead>
<tr>
<th>Winding</th>
<th>Inductance $L$</th>
<th>AC Resistance $R^{AC}$</th>
<th>Winding</th>
<th>Inductance $L$</th>
<th>AC Resistance $R^{AC}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_A$</td>
<td>6.02 $\mu$H</td>
<td>3.18 $\Omega$</td>
<td>$W_K$</td>
<td>5.78 $\mu$H</td>
<td>3.07 $\Omega$</td>
</tr>
<tr>
<td>$W_B$</td>
<td>5.93 $\mu$H</td>
<td>2.95 $\Omega$</td>
<td>$W_L$</td>
<td>5.70 $\mu$H</td>
<td>2.83 $\Omega$</td>
</tr>
<tr>
<td>$W_C$</td>
<td>5.55 $\mu$H</td>
<td>2.70 $\Omega$</td>
<td>$W_M$</td>
<td>5.33 $\mu$H</td>
<td>2.67 $\Omega$</td>
</tr>
<tr>
<td>$W_D$</td>
<td>5.78 $\mu$H</td>
<td>2.99 $\Omega$</td>
<td>$W_N$</td>
<td>5.56 $\mu$H</td>
<td>2.93 $\Omega$</td>
</tr>
<tr>
<td>$W_E$</td>
<td>5.86 $\mu$H</td>
<td>2.93 $\Omega$</td>
<td>$W_O$</td>
<td>5.69 $\mu$H</td>
<td>2.96 $\Omega$</td>
</tr>
<tr>
<td>$W_F$</td>
<td>5.90 $\mu$H</td>
<td>3.21 $\Omega$</td>
<td>$W_P$</td>
<td>5.65 $\mu$H</td>
<td>2.91 $\Omega$</td>
</tr>
<tr>
<td>$W_G$</td>
<td>5.62 $\mu$H</td>
<td>2.90 $\Omega$</td>
<td>$W_Q$</td>
<td>5.55 $\mu$H</td>
<td>2.59 $\Omega$</td>
</tr>
<tr>
<td>$W_H$</td>
<td>n/c</td>
<td>n/c</td>
<td>$W_R$</td>
<td>5.57 $\mu$H</td>
<td>2.76 $\Omega$</td>
</tr>
<tr>
<td>$W_J$</td>
<td>5.97 $\mu$H</td>
<td>3.07 $\Omega$</td>
<td>$W_S$</td>
<td>5.57 $\mu$H</td>
<td>2.96 $\Omega$</td>
</tr>
<tr>
<td>$W_J$</td>
<td>5.64 $\mu$H</td>
<td>2.92 $\Omega$</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

The difference between the calculated and measured primary winding inductances is 7.06 % with an average difference of 2.48 %. The maximum difference between the calculated and measured primary winding AC-resistance is 33.03 % with an average difference of 23.57 %.

The primary winding inductance measurements show very good agreement with the calculated values, the AC-resistances, however, show some differences. The main reason for these differences come from the thin connecting wires used to connect the primary windings to the rest of the circuits. These wires have very small diameters and add to the resistances of the primary windings.

### 4.3.2 The Primary Controller

Due to the relatively high operating frequency of the variable-phase CET platform, the use of commercially available off-the-shelf control systems like for instance, the dSpace DS1104 controller board [77], are not viable options due to their limited data sampling frequencies of up to a few hundred kilohertz. These systems are not able to handle the high-frequency megahertz-range operations required by the variable-phase CET platform.

Another option for controlling the CET platform is a microcontroller. A microcontroller is a device which contains a central processing unit (CPU) which executes programming instructions sequentially, and are generally used for complex operations and applications which do not require “real-time” operations.

FPGA’s on the other hand, support concurrent logic operations through the use of multiple interconnected logic cells. FPGA’s are generally used for applications consisting of multiple logical operations and are especially useful for interfacing, manipulating and controlling multiple logic signals. Due to the concurrent logic operations, all operations in a FPGA run in parallel, which allow for “near-real-time” operations to occur.

The variable-phase CET platform prototype consists of various clock signals,
PWM signals, clocked data signals and addressing signals which all require precise timings. Most of the operations performed in the controller are relatively simple, and consists of taking certain input signals, like clock signals and ADC data signals, manipulating it through simple operations, and in turn generate appropriate output signals like PWM- and addressing signals. For these reasons, the use of a FPGA controller is chosen.

A Xilinx Spartan 3A DSP 1800 FPGA controller board [76] is chosen as controller for the variable-phase CET platform. Some of the main attributes of this FPGA controller, which makes it very useful for this application are:

- a fast 125 MHz oscillator clock ($f_{FPGA}$),
- a RS-232 serial interface (for communications with the PC),
- approximately 1800k system gates and 37440 logic cells available for programming and storing the logic circuits,
- approximately 160 programable input-output pins available via two external connectors (for connecting the various CET platform subsystem signals),
- easy to use, VHDL programming software for a Windows PC (Xilinx ISE software),
- easy to use, FPGA programming USB-to-JTAG connector.

Figure 4.7 (a) shows a image of the FPGA controller board. Here the FPGA controller, the RS-232 interface port, as well as the two large external input-output (I/O) data pin connectors are visible. Figure 4.7 (b) shows a part of the FPGA controller board, with signaling wires connected to the external connector.

The controller is programmed in VHDL [78] and Fig. 4.8 shows the various functions implemented in the FPGA controller for one hardware channel. Here the black lines indicate the flow of data, the gray lines show the various clock signals, and the gray dashed lines show the signals coming from, and going to, the various subsystems.

Table 4.4 shows the most important external IO-pin signals used for sending signals between the FPGA controller and the different variable-phase CET platform subsystems (per hardware channel).

Regarding the FPGA main clock frequency, the generation of clock signals for the various subsystems, and the resolution at which the FPGA can sample incoming signals:

The FPGA’s main clock frequency runs at 125 MHz. In order to generate the 2.777 MHz clock signal used in the variable-phase CET platform, the main clock is fed through a clock dividing function using a counter. With the counter set to 45, a clock signal of 2.777 MHz is produced ($125 \text{ MHz} \div 45 = 2.777 \text{ MHz}$). Since the counter value is not even, a 50 % duty cycle is not possible, instead, an acceptable duty cycle of 48.89 % is achieved.

The FPGA controller will need to adjust the phase of the generated 2.777 MHz clock signal in order to manipulate the primary winding current phases.
4.3. The Primary System and Circuits

Figure 4.7: (a) The Xilinx Spartan 3A DSP 1800A FPGA primary controller. (b) The primary controller connected to the CET platform subsystems through the external connector on the development board.

Figure 4.8: Block diagram showing the various functions implemented in the FPGA controller, for one hardware channel. Black lines indicate information flow, gray lines show clock signals, and dashed gray lines show inputs and output to the various connected subsystems.
Table 4.4: The main input and output (I/O) signals from the FPGA controller to the different CET platform subsystems, per channel.

<table>
<thead>
<tr>
<th>Signal [bus width]</th>
<th>I/O</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>HB-CLK [1]</td>
<td>O</td>
<td>High-frequency 2.777 MHz clock signal for controlling the half-bridge driver.</td>
</tr>
<tr>
<td>VDC-ADC-CLK [1]</td>
<td>O</td>
<td>Clock signal for the ADC measuring the half-bridge inverter input DC-voltage level.</td>
</tr>
<tr>
<td>VDC-ADC-nCS [1]</td>
<td>O</td>
<td>Chip select signal for the ADC measuring the half-bridge inverter input DC-voltage level.</td>
</tr>
<tr>
<td>VDC-ADC-DATA [1]</td>
<td>I</td>
<td>Data signal for the ADC measuring the half-bridge inverter input DC-voltage level.</td>
</tr>
<tr>
<td>ICET-ADC-CLK [1]</td>
<td>O</td>
<td>Clock signal for the ADC measuring the primary winding current amplitude.</td>
</tr>
<tr>
<td>ICET-ADC-nCS [1]</td>
<td>O</td>
<td>Chip select signal for the ADC measuring the primary winding current amplitude.</td>
</tr>
<tr>
<td>ICET-ADC-DATA [1]</td>
<td>I</td>
<td>Data signal for the ADC measuring the primary winding current amplitude.</td>
</tr>
<tr>
<td>VCET-PHASE [1]</td>
<td>I</td>
<td>Digital representation of the half-bridge output voltage phase.</td>
</tr>
</tbody>
</table>

The 2.777 MHz clock has (conveniently) a signal period of 360 ns. Thus 1 ns per phase degree. The FPGA main clock, however, has a signal period of 8 ns only. The FPGA controller will thus only be able to measure and control the primary winding current phases with a resolution of 8°.

Similarly to the generation of the 2.777 MHz clock signal, a 50 kHz pulse width modulated (PWM) digital signal is generated for controlling the quadratic buck converter. For this purpose, a clock dividing function with a counter set to 2500 is used. Due to the relatively large difference between the FPGA clock frequency and the PWM signal frequency, the PWM signal can be controlled with a duty cycle resolution of 0.04 %.

4.3.3 The Quadratic Buck Converter

The quadratic buck converters are used to convert the high input DC-voltage levels to lower DC voltages. Each hardware channel contains one converter.

With regular buck converters operating in continuous conduction mode, the relationship between the duty cycle and the input-output voltage ratio is linear. Naturally, the upper limit of the duty cycle is 100 %. The lower limit of the duty cycle however, is limited to the switching speed of the device driving the
MOSFET in the converter. Generally, duty cycles of less than 5% (for devices switching at 50 kHz or faster) become difficult to maintain. Consequently, this limited duty cycle range, also limits the output voltage range that the buck converter can achieve.

By using (3.10) together with the information given in Table 3.2, Table 3.3, as well as Table 3.5, it can be calculated that the voltages required to transfer up to 8 W over the CET platform could range from between approximately 2 V to 50 V. Taking into account the voltage drops due to the MOSFET on-resistances in the half-bridge inverter and winding commutation circuits, as well as possible resonant capacitor and inductor mismatches, this voltage difference could possibly be much larger.

To ensure that the variable-phase CET platform can supply the voltages needed to transfer the desired power, a quadratic buck converter is implemented.

Quadratic buck converters are DC-to-DC switch mode step-down converters similar to buck converters, but with an output voltage which has a quadratic dependency on the switching duty cycle. Essentially, a quadratic buck converter behaves like two series-cascaded regular buck converters; however, just one switch is required in the current topology (see Fig. 4.9).

Each quadratic buck converter feeds the half-bridge inverter which generates a switching voltage and the sinusoidal current for the primary windings. They form part of a feedback loop which is used to control the amplitude of the primary winding current.

The quadratic buck converters are designed to operate in continuous conduction mode, and are based on the designs presented in [92, 93]. They operate at a switching frequency of $f_{BUCK} = 50$ kHz, and are designed to supply up to 0.75 A (DC) of load current.

Figure 4.9 shows a circuit schematic of the device. Here, $V_i$ and $I_i$ are the input DC-voltage and input current into the converter, respectively. The output DC-voltage and output current are given as $V_o$ and $I_o$, respectively. The signal denoted as “PWM”, is the 50 kHz pulse width modulated square wave signal which is used to drive the MOSFET.
The values for the various components in Fig. 4.9 as well as their descriptions are given in Table 4.5.

Table 4.5: Values and descriptions of the various components used in the quadratic buck converters.

<table>
<thead>
<tr>
<th>Object</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_1$</td>
<td>10 $\mu$F, 400V</td>
<td>Electrolytic capacitor used for smoothing the input voltage ripple.</td>
</tr>
<tr>
<td>$D_1$, $D_2$</td>
<td>BYV26C, BYV26E</td>
<td>600 V, 1A, Ultra fast avalanche sinterglass diodes [87].</td>
</tr>
<tr>
<td>$L_1$</td>
<td>11.49 mH, 44 $\Omega$</td>
<td>Wire wound inductor on ETD34 core using 3C15 core material [89], 155 turns and two 0.12 mm air gaps. Wire diameter is 0.25 mm.</td>
</tr>
<tr>
<td>$C_2$</td>
<td>10 $\mu$F, 400V</td>
<td>Electrolytic capacitor.</td>
</tr>
<tr>
<td>$T_1$</td>
<td>IRF840</td>
<td>N-channel MOSFET [88].</td>
</tr>
<tr>
<td>$D_3$</td>
<td>BYV26E</td>
<td>1 kV, 1A, Ultra fast avalanche sinterglass diode [87].</td>
</tr>
<tr>
<td>$L_2$</td>
<td>1.78 mH, 6.3 $\Omega$</td>
<td>Wire wound inductor on ETD34 core using 3C15 core material [89], 47 turns and two 0.12 mm air gaps. Wire diameter is 0.5 mm.</td>
</tr>
<tr>
<td>$C_3$</td>
<td>10 $\mu$F, 400V</td>
<td>Electrolytic capacitor used for smoothing the output voltage ripple.</td>
</tr>
<tr>
<td>$U_1$</td>
<td>HCPL2200</td>
<td>Logic gate optocoupler [91].</td>
</tr>
<tr>
<td>$U_2$</td>
<td>IR2213</td>
<td>MOSFET half-bridge driver [90].</td>
</tr>
</tbody>
</table>

Figure 4.10 shows a photo of the implemented quadratic buck converter.

With the input voltage set at $V_i = 50$ V, various measurements are performed on the quadratic buck converter in order to determined the exact input-to-output voltage ratios for various duty cycles and load currents. Figure 4.11 shows the measurement results.

Furthermore, the efficiency of the quadratic buck converter is calculated based on the measurements performed, and the results are shown in Fig. 4.12.

The results from Fig. 4.11 show that the output voltage of the implemented quadratic buck converter does have a quadratic dependency to the duty cycle, and is generally independent of the load current (for $I_o \geq 100$ mA). The output voltage amplitude, however, is generally 7% - 9% less than expected. The main reason for the voltage difference is due to the voltage drops over the relatively large series inductor resistances.

The efficiency results from Fig. 4.12 show that the efficiencies increase for higher duty cycles, which is about 90% for duty cycles around 0.9. At duty cycles of 0.5, the efficiency drops to approx. 70%. At lower duty cycles the efficiency drops further, to minimum values of between 35% and 45%.

The performance obtained with the quadratic buck converter is good enough for testing the variable-phase CET platform. However, it should be emphasized that the device for this converter topology as such, is not the only possibility.
Figure 4.10: *The implemented quadratic buck converter.*

Figure 4.11: *Quadratic buck converter input-to-output voltage ratio measurement results for various duty cycles and load currents.*

for the realization of an extended output-to-input voltage ratio. Off-the-shelf solutions for commercial applications, like the series connection of integrated and optimized designs of standard buck converters are available in order to cope with broad output-to-input voltage ratios.
4.3.4 The High-Frequency Half-bridge MOSFET Driver

Half-bridge MOSFET drivers are specialized circuits which are used to drive the MOSFETs in half-bridge inverters. Each hardware channel contains one such driver.

Due to the relatively high half-bridge switching frequency of 2.777 MHz, normal off-the-shelf half-bridge drivers [90, 98, 99] can not be used, since they are generally limited to switching frequencies of up to 1 MHz. For this reason a new high-frequency half-bridge MOSFET resonant gate driver, based on the designs in [94, 95] is developed.

Figure 4.13 shows the schematic circuit diagram of the implemented high-frequency MOSFET half-bridge driver. Here, \( V_1 \) is a 3.3V - 2.777 MHz clock signal generated by the controller for driving the circuit. The voltages \( V_2 \) and \( V_2^* \) are two 180° phase-shifted sinusoidal voltage signals used for driving the two MOSFETs in the half-bridge inverter. Voltage \( V_3 \) is the DC-voltage output from the quadratic buck converter, and \( V_4 \) is the square wave output voltage of the half-bridge inverter.

The values for the various components in Fig. 4.13 as well as their descriptions are given in Table 4.6.

As seen in Fig. 4.13, the half-bridge driver consists of four parts. The operation of these four parts can be described as follows:

- Part 1 forms a high-frequency current buffer which is used to supply current to the rest of the circuit. The input signal \( V_1 \), is a 3.3V - 2.777 MHz square wave signal generated by the controller. This signal is first amplified to 9 V using a logic level MOSFET, whereafter it is fed into a high-speed current buffer. In this way, the high-frequency square wave signal is fed into part 2
4.3. The Primary System and Circuits

![Circuit schematic of the high-frequency MOSFET half-bridge driver.](image)

**Figure 4.13:** Circuit schematic of the high-frequency MOSFET half-bridge driver.

**Table 4.6:** Values and descriptions of the various components used in the half-bridge MOSFET drivers.

<table>
<thead>
<tr>
<th>Device</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_1$</td>
<td>750 Ω</td>
<td>1/4 W Resistor limits the input current to approx. 4 mA.</td>
</tr>
<tr>
<td>$T_1$</td>
<td>BSS138</td>
<td>Logic level MOSFET [96].</td>
</tr>
<tr>
<td>$R_2$</td>
<td>4.7 kΩ</td>
<td>1/4 W Resistor limits the MOSFET drain-source current to approx. 2 mA.</td>
</tr>
<tr>
<td>$U_1$</td>
<td>BUF634</td>
<td>A 250 mA High-speed current buffer [97].</td>
</tr>
<tr>
<td>$C_1$</td>
<td>590 pF, 100 mΩ</td>
<td>Polystyrene resonance capacitor [75].</td>
</tr>
<tr>
<td>$L_1$</td>
<td>5.44 µH, 410 mΩ</td>
<td>Wire wound inductor on a 10/6/4 mm ferrite ring using 4C65 core material [82], 10 turns.</td>
</tr>
<tr>
<td>$L_2$</td>
<td>1.74 µH, 160 mΩ</td>
<td>Wire wound inductor on a 12.5/7.5/5 mm ferrite ring using 4C65 core material [82], 4.5 turns.</td>
</tr>
<tr>
<td>$L_3$</td>
<td>430 nH, 72 mΩ</td>
<td>Wire wound inductor on a 10/6/4 mm ferrite ring using 4C65 core material [82], 3 turns.</td>
</tr>
<tr>
<td>$L_4$</td>
<td>1.53 µH, 190 mΩ</td>
<td>Wire wound inductor on a 10/6/4 mm ferrite ring using 4C65 core material [82], 6 turns.</td>
</tr>
<tr>
<td>$L_5$</td>
<td>1.55 µH, 180 mΩ</td>
<td>Wire wound inductor on a 10/6/4 mm ferrite ring using 4C65 core material [82], 6 turns.</td>
</tr>
<tr>
<td>$C_4$</td>
<td>820 pF</td>
<td>Philips polystyrene capacitor [75].</td>
</tr>
<tr>
<td>$C_7$</td>
<td>820 pF</td>
<td>Philips polystyrene capacitor [75].</td>
</tr>
<tr>
<td>$T_2^*$</td>
<td>IRF620</td>
<td>N-channel HEXFET MOSFET (high-side) [101].</td>
</tr>
<tr>
<td>$T_2$</td>
<td>IRF620</td>
<td>N-channel HEXFET MOSFET (low-side) [101].</td>
</tr>
</tbody>
</table>

of the circuit, without drawing current from the controller. Resistor $R_1$ is used to limit the current drawn from the controller, and $R_2$ is used to limit the MOSFET drain-source current.
Chapter 4. Implementation of the Variable-Phase CET Platform

- Part 2 forms a series resonant LC circuit. The inductor and capacitor form a series resonant circuit tuned to approx. 2.78 MHz. They act as a band-pass filter, allowing only the fundamental frequency current component to flow. The current $I_1$ is thus sinusoidal.

- Part 3 is an adaptation of the resonant-gate half-bridge driver used in the self-oscillating hybrid inverter presented in [94]. Here, $L_3$, $L_4$, and $L_1^*$ form a three-winding toroidal transformer. The three windings on the transformer have similar couplings of approx. 0.76. With the half-bridge MOSFETs’ input capacitances and resistances known, the values of the inductances and capacitances in this part are calculated as to generate two 20 V (peak-to-peak) out-of-phase sinusoidal voltage signals for driving the two MOSFETs, from a sinusoidal input current of approx. 200 mA (peak).

- Part 4 is the actual MOSFET half-bridge. It will not be discussed here, but in subsection 4.3.5 instead.

Figure 4.14 shows a photo of the implemented half-bridge driver.

Figure 4.14: Circuit implementation of the high-frequency MOSFET half-bridge driver.

Figure 4.15 shows voltage and current waveforms captured by the oscilloscope (Tektronix TDS 744 A) during the half-bridge MOSFET driver and half-bridge inverter operations. In this figure, trace 1 represents voltage $V_1$, the 3.3V - 2.777 MHz square wave signal generated by the controller. Trace 2 represents voltage $V_*^2$, the low-side MOSFET's ($T_2^*$) gate-source voltage. Traces 3 and 4 will be discussed in subsection 4.3.5.
Figure 4.15: Voltage and current measurements taken during operation of the half-bridge MOSFET driver and half-bridge inverter. Trace 1: the 3.3V - 2.777 MHz clock input to the half-bridge MOSFET driver, $V_1$ (5 V/div); trace 2: half-bridge low-side MOSFET ($T_2^*$) gate-source voltage, $V_{2^*}$ (10 V/div); trace 3: primary winding current $I_P$ (0.5 A/div); trace 4: voltage output of the half-bridge inverter (5 V/div). Timescale: 100 ns/div.

From this figure it can be seen that the generated MOSFET gate-source voltage signal is approximately sinusoidal. It has a peak voltage of 8.6 V which is sufficient for driving the MOSFETs, as will be shown in the next section.

4.3.5 The Half-Bridge Inverter

The half-bridge inverters convert the DC-voltage outputs from the quadratic buck converters into square wave AC-voltage signals which are used to drive the primary hexagon spiral windings. Each CET hardware channel contains one inverter. The half-bridge inverter contains two N-channel MOSFETs connected in the typical half-bridge inverter configuration.

The circuit schematic of the MOSFETs in the half-bridge inverter is shown in Fig. 4.13 part 4, while a photo of the implemented half-bridge inverter is shown in Fig. 4.14. Here, the high-side MOSFET is labeled $T_2$, while the low-side MOSFET is labeled $T_2^*$. The MOSFETs implemented in the half-bridge inverter are both IRF620 N-channel MOSFETs, as described in Table 4.6. Some of the important electrical characteristics of these MOSFETs are given in Table 4.7.

The IRF620 is appropriately chosen for the variable-phase CET platform, with a maximum allowed drain-source voltage of at least double the expected maximum half-bridge output voltage. The maximum allowed drain current is also more than six times the maximum expected winding current, as to handle...
possible current peaks, which might occur during certain current controller step responses.

Figure 4.16 shows the output characteristics of the IRF620 MOSFET, and by studying it, it can be determined whether the MOSFETs will be sufficiently switched on during their operations.

![Figure 4.16: The IRF620 output characteristics for various gate-source voltages.](image)

Firstly, in Fig. 4.15 it is shown that the gate-source voltage amplitudes of the two MOSFETs' are approx. 8.6 V (peak). Also, from the design of the half-bridge MOSFET driver (subsection 4.3.4), it is shown that the two MOSFETs in the half-bridge inverter are driven by two 180° out-of-phase gate-source voltages. The IRF620s have gate-source threshold voltages of 3.5 V, and due to the sinusoidal
4.3. The Primary System and Circuits

The nature of the applied gate-source voltages, there exists brief periods when the applied gate-source voltages are lower than the threshold values. This occurs right when the MOSFETs are being switched on, and also when they are being switched off. During the deactivation of one MOSFET and the activation of the other, there exists thus a short period of time that neither MOSFET will conduct any current. This creates a 48 ns dead-time during the switching times, which will avoid short circuits between the applied DC-voltage and the ground signal from occurring.

Furthermore, due to the series resonance between the primary winding and the resonant capacitor, the primary winding current (also the MOSFET drain current) is sinusoidal and approximately in-phase with the half-bridge inverter’s gate-source voltage signal. At the switching instant, both MOSFETs’ drain currents, and thus, their drain-source voltages will be approximately zero. The MOSFETs thus operate with zero-voltage switching, which reduces associated switching losses [102].

Additionally, Fig. 4.15 trace 4, shows the measured output voltage signal ($V_4$ in Fig. 4.13) of the half-bridge inverter while a sinusoidal current of 0.5 A (peak) is flowing through the primary winding $W_A$. The output voltage signal is approximately square wave shaped. Trace 3 shows the current waveform, which is approximately sinusoidal.

4.3.6 The Primary Current Estimator

The primary current estimator circuits are used to estimate the primary winding current amplitudes and phases, and to report the information back to the main controller. Each hardware channel contains one such estimator. The current estimator consists of two distinct parts: the current amplitude estimator, and the current phase estimator.

The current amplitude estimator is used to estimate the amplitude of the primary winding current. It uses a small toroidal current transformer to convert the primary current into a voltage signal. Using an envelope detector, a resistor divider network, and a diode clamping circuit, the AC-current is converted into a DC-voltage signal. An analog-to-digital converter (ADC) is then used to measure and report the voltage amplitude back to the main controller. The current amplitude-to-ADC data value transfer function is estimated, and the inverse transfer function is implemented in the controller, in order to estimate the correct current amplitude.

The current phase estimator is used to estimate the phase of the primary winding current. It uses the same above mentioned toroidal core to induce a voltage in another secondary winding. This AC-voltage is clamped to $\pm 0.7$ V using a diode clamping circuit and afterwards amplified to $\pm 4$ V using an operational amplifier. Finally the negative sequence is removed using a diode circuit. The remaining signal is a positive square wave signal with approx. 3.3 V amplitude which is $90^\circ$ out-of-phase with the current signal. This signal is fed back to the controller, which samples it, and estimates the current phase by detecting the low-to-high and the high-to-low signal transitions. Due to the
limited clock frequency and sampling rate of the main controller, this sampled
signal has a resolution of only 8°. Moreover, the ADC’s run at a clock frequency
of 1.25 MHz, but due to their serial interfaces, the ADC data registers in the
main controller are updated at a rate of 50 kHz. In this way, the PWM signals
of the quadratic buck converters are synchronized with the ADC data registers’
update frequency.

Figure 4.17 shows a circuit schematic of the primary current estimator. Here,
I_P is the primary winding current, V_1 is an DC-voltage representation of the
current amplitude, and V_2 is a clamped 3.3 V square wave signal representation
of the primary winding current phase.

The values for the various components in Fig. 4.17, as well as their descriptions
are given in Table 4.8 and Table 4.9.

A photo of the implemented current estimator circuit is shown in Fig. 4.18.
The current amplitude estimators’ primary current amplitude-to-induced volt-
age and ADC data transfer function is measured and shown in Fig. 4.19.

4.3.7 The Primary Voltage Estimator

The primary voltage estimator circuits are used to estimate the output DC-
voltage amplitudes of the quadratic buck converters, as well the phases and am-
plitudes of the square wave signals generated by the half-bridge inverters. The
information is then reported back to the main controller. Each hardware channel
contains one such estimator. The voltage estimator consists of two parts: the
voltage amplitude estimator, and the voltage phase estimator.
The voltage amplitude estimator is used to estimate the DC-voltage amplitude of the quadratic buck converter’s output voltage, and from that, estimate the half-bridge inverter’s square wave output voltage signal amplitude. Using a resistor divider network, as shown in Fig. 4.20, the output from the quadratic buck converter is divided by about 30 times, so that a voltage of 100 V (DC) is
reduced to 3.33 V. An 12-bit analog-to-digital converter (ADC) is then used to measure and report the voltage amplitude back to the main controller. The voltage amplitude-to-ADC-data-value transfer function is linear, and is represented by the equation \( D = (40.96)(V_o) \). Here, \( D \) is the ADC data value, and \( V_o \) is the
quadratic buck converter DC-voltage amplitude. The inverse transfer function is implemented in the controller, in order to estimate the correct voltage amplitude.

The voltage phase estimator is used to estimate the phase of the square wave voltage signal generated by the half-bridge inverter. Here, a resistor divider network (similar to the above mentioned one) is used to reduce the voltage amplitude to under 3.33 V. A high-speed operational amplifier configured as a non-inverting amplifier with a large gain, is used to amplify this voltage signal. The output signal, which is limited by the operational amplifier to 4.40 V, is connected to two forward biased series diodes and grounded through a resistor. The two diodes produce a combined forward voltage drop of 1.40 V, and the resistor is chosen to limit the operational amplifier’s output current to approx. 5 mA. In this way, the voltage over the resistor is limited to 3.00 V and is approximately in-phase with the half-bridge inverter output signal. This signal is fed back to the controller, which samples it, and estimates the voltage phase by detecting the low-to-high and the high-to-low signal transitions.

Figure 4.20 shows a circuit schematic of the primary voltage estimator. Here, \( V_1 \) is the quadratic buck converter’s output voltage, \( V_2 \) is the voltage into the analog-to-digital converter, and \( V_5 \) is the 3.0 V square wave voltage signal representing the half-bridge output voltage.

![Circuit schematic of the voltage estimator circuit.](image)

The values for the various components in Fig. 4.20 as well as their descriptions are given in Table 4.10.

Photos of the implemented voltage amplitude and phase estimator circuits are shown in Fig. 4.21 (a) and Fig. 4.21 (b), respectively.
Table 4.10: Values and descriptions of the various components used in the voltage estimator circuits.

<table>
<thead>
<tr>
<th>Device</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>( T_1 )</td>
<td>IRF620</td>
<td>Half-bridge inverter IRF620 N-channel MOSFETs [101].</td>
</tr>
<tr>
<td>( T_1^* )</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( R_1 )</td>
<td>1.47 kΩ</td>
<td>1/4 W Axial resistor.</td>
</tr>
<tr>
<td>( R_2 )</td>
<td>50 Ω</td>
<td>1/4 W Axial resistor.</td>
</tr>
<tr>
<td>( R_3 )</td>
<td>1.47 kΩ</td>
<td>1/4 W Axial resistor.</td>
</tr>
<tr>
<td>( R_4 )</td>
<td>50 Ω</td>
<td>1/4 W Axial resistor.</td>
</tr>
<tr>
<td>( C_1 )</td>
<td>10 µF, 16V</td>
<td>Electrolytic capacitor.</td>
</tr>
<tr>
<td>( U_1 )</td>
<td>ADCS7476</td>
<td>12-bit 1-MSPS Analog-to-digital converter [81].</td>
</tr>
<tr>
<td>( U_2 )</td>
<td>AD8011</td>
<td>300 MHz Current feedback amplifier [80].</td>
</tr>
<tr>
<td>( R_5 )</td>
<td>5.6 kΩ</td>
<td>1/4 W Axial resistor.</td>
</tr>
<tr>
<td>( R_6 )</td>
<td>150 Ω</td>
<td>1/4 W Axial resistor.</td>
</tr>
<tr>
<td>( D_1, D_2 )</td>
<td>1N4148</td>
<td>Small signal diode, 4 ns reverse recovery time [79], ( V_R = 100 \text{ V}, ) ( V_F = 700 \text{ mV}, ) ( I_O = 200 \text{ mA}. )</td>
</tr>
<tr>
<td>( R_7 )</td>
<td>560 Ω</td>
<td>1/4 W Axial Resistor.</td>
</tr>
</tbody>
</table>

Figure 4.21: Circuit implementation of the (a) voltage amplitude estimator circuit, (b) the voltage phase estimator circuit.

4.3.8 The Primary Winding Commutation Circuit

The winding commutation circuits allow the square wave voltage signals generated by the half-bridge inverters to be directed to any primary winding connected to its associated hardware channel. Each hardware channel contains one commutation circuit. Pairs of anti-series connected N-channel MOSFETs are
connected to the different primary windings, and act as switches, which allow current to flow into whichever switch is activated. The MOSFETs are activated using photovoltaic MOSFET drivers, controlled by a 3-to-8 line demultiplexer. The demultiplexer in turn, receives a three bit address from the FPGA controller, indicating which MOSFET to activate. A red LED is placed in parallel with each MOSFET driver, which gives visual feedback to indicate which MOSFET switch is activated.

Figure 4.22 shows a circuit schematic of the winding commutation circuit. Here, \( n \) indicates the amount of primary windings connected to the associated hardware channel. For channel I, \( n = 6 \), for channel II, \( n = 7 \), and for channel III, \( n = 5 \).

![Circuit Diagram](image)

**Figure 4.22:** Electric circuit of the winding commutation circuit.

The values and descriptions of all the various components in Fig. 4.22 are given in Table 4.11.

Figure 4.23 shows a photo of the implemented winding commutation circuit for hardware channel II.
Table 4.11: Values and descriptions of the various components used in the winding commutation circuits.

<table>
<thead>
<tr>
<th>Device</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_1$, $R_3$, $R_5$</td>
<td>1 kΩ</td>
<td>1/4 W Pull-up resistors.</td>
</tr>
<tr>
<td>$R_2$, $R_4$, $R_6$</td>
<td>12 Ω</td>
<td>1/4 W Series current limiting resistors.</td>
</tr>
<tr>
<td>$U_0$</td>
<td>74HC238</td>
<td>3-to-8 Line demultiplexer [84].</td>
</tr>
<tr>
<td>$R_{7+}$, ..., $R_{6+n}$</td>
<td>270 Ω</td>
<td>1/4 W Current limiting series resistors.</td>
</tr>
<tr>
<td>$R_{7+n}$, ..., $R_{6+2n}$</td>
<td>270 Ω</td>
<td>1/4 W Current limiting series resistors.</td>
</tr>
<tr>
<td>$D_1$, ..., $D_n$</td>
<td>LEDs</td>
<td>5 mm Standard red LEDs.</td>
</tr>
<tr>
<td>$U_1$, ..., $U_n$, $AVP1121$</td>
<td></td>
<td>Photovoltaic MOSFET driver [86].</td>
</tr>
<tr>
<td>$T_1$, ..., $T_{2n}$</td>
<td>STP11NK40Z</td>
<td>400V, 9A, N-channel MOSFETs [85].</td>
</tr>
</tbody>
</table>

Figure 4.23: Implemented winding commutation circuit.

4.4 The Secondary System and Circuits

The secondary system and circuits refer to the circuits which receive power from the variable-phase CET platform through magnetic induction. These are the secondary winding, the rectifier circuit, the DC-to-DC converter and the load.
4.4. The Secondary System and Circuits

4.4.1 The Secondary Winding

The secondary winding is a single layer hexagon spiral winding produced as copper tracks on a printer circuit board, with the dimensions and specifications as discussed in subsection 3.4.4 and shown in Table 3.2. Two secondary windings are produced, and are labeled $W_Y$ and $W_Z$, respectively. Winding $W_Z$ is primarily used for power transfer, and $W_Y$ is a backup.

Figure 4.24 shows a picture of the secondary winding $W_Z$.

![ Implemented secondary hexagon spiral winding $W_Z$. ]

The inductances and AC-resistances of the two implemented secondary windings as shown in Fig. 4.24, are measured using an Agilent 4294A Precision Impedance Analyzer (40 Hz - 110 MHz). The results are displayed in Table 4.12.

Table 4.12: Measured secondary winding inductances and AC-resistances.

<table>
<thead>
<tr>
<th>Winding</th>
<th>Inductance $L$</th>
<th>Resistance $R^{AC}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_Y$</td>
<td>7.07 $\mu$H</td>
<td>2.08 $\Omega$</td>
</tr>
<tr>
<td>$W_Z$</td>
<td>7.17 $\mu$H</td>
<td>2.00 $\Omega$</td>
</tr>
</tbody>
</table>

The calculated values for the primary winding inductance and AC-resistance, as shown in Table 3.2, are 7.16 $\mu$H and 1.95 $\Omega$, respectively. The maximum difference between the calculated and measured primary winding inductances is 1.27 % with an average difference of 0.70 %. The maximum difference between the calculated and measured primary winding AC-resistance is 6.45 % with an average difference of 4.49 %.

4.4.2 The Rectifier Circuit

The full-bridge rectifier circuit is used to convert the high-frequency 2.777 MHz secondary induced voltage into a more usable DC-voltage signal.

Figure 4.25 shows a circuit schematic of the full-bridge rectifier circuit. Here, $V_1$ is the winding voltage, and $I_1$ is the winding current. The DC-voltage output from the rectifier is $V_L$, the load resistance is $R_L$, and the load current is $I_L$.

The values and descriptions of all the various components used in Fig. 4.25,
are given in Table 4.13.

**Table 4.13:** Values and descriptions of the various components used in the full-bridge rectifier circuit.

<table>
<thead>
<tr>
<th>Device</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1, R_1$</td>
<td>$W_X$</td>
<td>The secondary windings' inductance and AC-resistance.</td>
</tr>
<tr>
<td>$C_1$</td>
<td>465 pF</td>
<td>Philips polystyrene resonant capacitor [75].</td>
</tr>
<tr>
<td>$D_1, D_2$, $D_3, D_4$</td>
<td>1N4148</td>
<td>Small signal diode, 4 ns reverse recovery time [79], $V_R = 100$ V, $V_F = 700$ mV, $I_O = 200$ mA.</td>
</tr>
<tr>
<td>$C_2$</td>
<td>3.3 $\mu$F</td>
<td>A 3.3 $\mu$F, 400 V, Electrolytic capacitor.</td>
</tr>
</tbody>
</table>

Figure 4.26 shows a photo of the implemented full-bridge rectifier circuit.

**Figure 4.25:** Circuit schematic of the full-bridge rectifier.

**Figure 4.26:** The implemented full-bridge rectifier circuit.
4.4. The Secondary System and Circuits

Figure 4.27 shows voltage and current waveforms captured by the oscilloscope (Tektronix TDS 744 A) during inductive power transfer using the rectifier circuit. The secondary winding $W_Z$ is placed above primary winding $W_A$ with a mutual inductance of approximately $M_{AZ} = 2.45 \, \mu\text{H}$. The rectifier is connected to a load resistor of 81 $\Omega$. In this figure, trace 1 (at the bottom) represents the load voltage $V_L$. Trace 3, (at the top) is the primary winding current. Trace 2 is voltage $V_2$, the input to the rectifier circuit, and is measured using a differential voltage probe (Tektronix P5205). Trace 4 represents the secondary winding current $I_1$ as shown in Fig. 4.25.

With the load voltage at approx. 20.3 V, 5.10 W of power is transferred to the load resistor. The primary current amplitude is 414 mA (RMS), and the secondary winding current amplitude is approx. 275 mA (RMS).

Figure 4.27: Voltage and current waveforms measured during inductive power transfer through the secondary rectifier circuit. Trace 1: the rectifier’s DC-voltage output (20 V/div), trace 3: primary winding current (1 A/div), trace 2: the rectifier’s input voltage (25 V/div), trace 4: secondary winding current (0.5 A/div). Timescale: 100 ns/div.

4.4.3 The DC-to-DC Converter

After the high-frequency secondary induced voltage is rectified, the resultant voltage is not constant, but varies depending on the primary current amplitude and the specific secondary winding position. The DC-to-DC converter is used to convert the output from the full-bridge rectifier to a stable DC-voltage of 5 V, which is more suitable for powering the mobile electronic devices. The voltage conversion is accomplished through the use of a small step-down LM2953-based [103] power converter.

Figure 4.28 shows a circuit schematic of the secondary circuit, including the
secondary winding, the rectifier circuit, and the DC-to-DC converter. Here \( V_1 \) and \( I_1 \) is the induced secondary winding voltage and winding current, respectively. The voltage \( V_2 \) is the input voltage to the rectifier circuit. Voltage \( V_3 \) and current \( I_3 \) is the rectifier output voltage and output current, respectively. Additionally, \( V_L \) is the DC-to-DC converter’s output voltage as well as the voltage over the load. The load current is \( I_L \).

\[
\begin{align*}
V_1 & \quad \text{secondary winding voltage} \\
I_1 & \quad \text{secondary winding current} \\
V_2 & \quad \text{input voltage to the rectifier circuit} \\
V_3 & \quad \text{rectifier output voltage} \\
I_3 & \quad \text{rectifier output current} \\
V_L & \quad \text{DC-to-DC converter output voltage} \\
I_L & \quad \text{load current}
\end{align*}
\]

The values and descriptions of the DC-to-DC converter’s components as shown in Fig. 4.28, are given in Table. 4.14.

**Table 4.14: Values and descriptions of the various components used in the DC-to-DC converter circuit.**

<table>
<thead>
<tr>
<th>Device</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>( U_1 )</td>
<td>LM2593HV</td>
<td>150 kHz 2A Step-down voltage regulator [103].</td>
</tr>
<tr>
<td>( D_5 )</td>
<td>1N4006</td>
<td>General purpose rectifier, ( V_R = 800 \text{ V}, I_F = 1 \text{ A} ) [105].</td>
</tr>
<tr>
<td>( L_2 )</td>
<td>33 ( \mu \text{H} )</td>
<td>SMT 33 ( \mu \text{H} ) power inductor [104].</td>
</tr>
<tr>
<td>( C_3 )</td>
<td>820 ( \mu \text{F}, 16\text{V} )</td>
<td>820 ( \mu \text{F}, 16 \text{ V} ) Electrolytic capacitor.</td>
</tr>
</tbody>
</table>

Figure 4.29 shows a photo of the implemented DC-to-DC converter circuit. Figure 4.30 shows voltage and current waveforms captured by the oscilloscope (Tektronix TDS 744 A) during inductive power transfer using the rectifier and DC-to-DC converter circuits. The secondary winding \( W_Z \) is placed above the primary winding \( W_A \) with a mutual inductance of approximately \( M_AZ = 2.13 \mu \text{H} \). The DC-to-DC converter output is connected to a variable resistor. In this figure, trace 1 (at the bottom) represents the 5 V DC-to-DC converter output voltage (also the load voltage), \( V_L \). Trace 3 is the primary winding current, measured using a Tektronix TM501 - AM503 current probe. Trace 4 represents the load current \( I_L \).
4.4. The Secondary System and Circuits

Figure 4.29: The implemented DC-to-DC converter circuit.

Figure 4.30: Voltage and current waveforms captured during inductive power transfer using the secondary rectifier and DC-to-DC converter. Trace 1: load voltage (5 V/div), trace 3: primary winding current (0.5 A/div), trace 4: load current (0.1 A/div). Timescale: 100 ns/div.

With a load voltage of 5 V (DC), and the load current of about 300 mA, approx. 1.50 W is transferred to the load resistor. The primary winding current amplitude is approx. 675 mA (RMS).
4.5 The Primary Current Controller

The primary current controllers are used to control the amplitudes and phases of the primary windings’ currents. Each hardware channel contains one current controller, which is implemented in the FPGA controller. The primary current controller consists of two parts: the current amplitude controller, and the current phase controller.

4.5.1 The Current Amplitude Controller

In order to maintain a constant current amplitude through an activated primary winding, the primary winding voltage needs to be actively controlled in order to compensate for the reflected voltage caused by current flowing in the secondary winding during power transfer.

Moreover, in Table 4.3 it is shown that the resistances and inductances of the primary windings in the CET platform matrix are not all the same. This means that for the same winding current amplitude, that the primary winding voltages will differ. This further supports the need for a current controller, in order to maintain a steady primary winding current.

While subsection 3.2.3 discusses the intended implementation of the PI current controller, only now after the implementation of the variable-phase CET platform is presented, and the hardware components and digital controller are known, can the compensator be implemented.

The current amplitude controller is implemented as a PI compensator as shown in Fig. 3.7. Figure 4.31 shows a block diagram of the FPGA controller with its implemented PI compensator and the process (or plant) which generates the primary winding current, and which the compensator will attempt to control. Here, the dashed gray lines indicate the input and output of the process. Resistance $R_{EQ}$ represents the combined resistances of the primary winding, and the resistances of the MOSFETs used in the half-bridge inverter as well as the winding commutation circuits. The disturbance $\delta$ is the influence of the current circulating in the secondary winding on the primary winding’s reflected voltage (the reflected impedance).

The development of the PI controller starts by determining the step response of the process operating in open-loop mode.

Firstly, with the quadratic buck converter input voltage set to 50 V, a positive step is placed on the signal $u$ in Fig. 4.31, as to generate a duty cycle of 38.40%, which will allow approx 0.5 A (peak) current to flow through primary winding $W_A$. The open-loop positive step response is measured using an oscilloscope (Tektronix TDS 744 A), and the captured voltage waveforms are show in Fig. 4.32. Here, trace 1 is the output voltage of the quadratic buck converter ($V_o$ as shown in Fig. 4.9 in subsection 4.3.3). Trace 2 is the output voltage of the primary current amplitude estimator circuit, the voltage that the ADC measures ($V_1$ in Fig. 4.17 from subsection 4.3.6). Trace 3 is a trigger signal, indicating the precise point at which the PWM signal is enabled. Trace 4 is the quadratic buck converter’s PWM signal.
4.5. The Primary Current Controller

Figure 4.31: Block diagram of the process (plant) which the PI controller will attempt to control.

Figure 4.32: Voltage measurements taken during the first open-loop current controller test performed on winding $W_A$. Trace 1: quadratic buck converter output voltage (5 V/div); trace 2: output voltage of the current amplitude estimator circuit (0.5 V/div); trace 3: trigger signal (5 V/div); trace 4: the quadratic buck converter’s PWM signal (5 V/div). Timescale: 5 ms/div.
From these results it can be concluded that the voltage outputs of both the quadratic buck converter and the current amplitude estimator circuit have overshoots of approximately 25.62 % and 19.70 %, accordingly. The settling times for both waveforms are about 6.00 ms.

Secondly, with the quadratic buck converter input voltage set to 50 V, a PWM duty cycle of 38.40 %, and a primary winding current of approx 0.5 A (peak), operating in steady state, a negative step is placed on \( u \), as to lower the duty cycle, the quadratic buck converter output voltage, and the primary current, to zero. The open-loop negative step response is measured using the same measurement setup as in the previous experiment, and all the traces are the same. Figure. 4.33 shows the captured waveforms.

![Figure 4.33: Voltage measurements taken during the second open-loop current controller test performed on winding \( W_A \). Trace 1: quadratic buck converter output voltage (5 V/div); trace 2: output voltage of the current amplitude estimator circuit (0.5 V/div); trace 3: trigger signal (5 V/div); trace 4: the quadratic buck converter’s PWM signal (5 V/div). Timescale: 5 ms/div.](image)

From these results it can be concluded that the voltage outputs of the quadratic buck converter and the current amplitude estimator circuit both have exponential decays, with time constants of approximately 4.4 ms and 8.5 ms, respectively. The settling time for the waveforms are about 10 and 20 ms, respectively.

Since no specifications are presented for the current controller, a reasonable maximum current settling time of approximately 1.00 second is proposed. The controller parameters are chosen according to the rule of thumb method for tuning PI controllers, as presented in [106]. The parameters of the PI controllers are set to:

- Proportional gain, \( K_P = 0.625 \).
• Integrating time is set to 20 µs.

• Integrator gain, $K_I = 76.30 \times 10^{-6} \ (1 \div 13107)$.

In order to test the current controller, a closed-loop test is performed on the variable-phase CET platform. Here, the secondary winding $W_Z$ is placed above the primary winding $W_A$ with a mutual inductance of about $M_{AZ} = 2.25 \mu H$. The secondary winding is connected to its resonant capacitor, the full-bridge rectifier circuit, and a variable resistor, as shown in Fig. 4.25. With the input voltage of the quadratic buck converter set to 65 V, the current controller is set to maintain a 750 mA (peak) current amplitude through primary winding $W_A$. The variable resistor is set to 96 Ω in order to transfer about 5 W to it. A peak primary winding current of 750 mA corresponds to an current amplitude estimator circuit output voltage of approximately 1.0 V.

The secondary winding is periodically placed-on and removed-from the CET platform, in order to test a step-like response of the PI controller.

Figure 4.34 shows the voltage and current waveforms captured by the oscilloscope (Tektronix TDS 744 A) during this test. In this figure, trace 1 is the output voltage of the quadratic buck converter, $V_o$ as shown in Fig. 4.9. Trace 2 is the voltage output of the current amplitude estimator circuit, the voltage that the ADC measures, $V_1$ in Fig. 4.17. Trace 3 is the load voltage $V_L$, and trace 4 is the load current $I_L$, measured using a Tektronix TM501 - AM503 current probe. During interval 1 and interval 3 of the measured waveforms, the secondary load is removed from the platform, and during interval 2 the secondary load is placed back on top of the platform.

Essentially, the current amplitude controller will attempt to maintain a constant primary current amplitude, shown as trace 2 in Fig. 4.34.

During interval 1, the secondary winding is not placed on the CET platform. The duty cycle of the quadratic buck converter is 16.70 %, and its output voltage is 10.53 V. The primary winding current is about 750 mA (peak).

During the transition between interval 1 and interval 2, when the secondary winding is placed back on the platform, the primary current drops as the reflected voltage increases due to current circulating the in the secondary winding. The output of the quadratic buck converter increases to compensate, until the primary current stabilizes after about 500 ms.

During interval 2, the quadratic buck converter’s duty cycle is 82.67 % and the half-bridge input voltage is 42.50 V. With the load voltage at 21.60 V, and the load current at 226 mA, approx. 4.88 W is transferred to the load resistor.

During the transition between interval 2 and interval 3, the secondary winding is suddenly removed from the platform. Here, the reflected voltage sharply decreases causing the primary current to rise. In reaction, the output voltage of the quadratic buck converter is reduced in order to maintain a constant primary winding current. After about 450 ms the primary current stabilizes to its correct value.

During interval 3, when the secondary winding is absent from the CET platform, the results are similar to that of interval 1.
The primary winding current settling time of about 500 ms is acceptable and smaller then the proposed maximum of 1.00 second.

4.5.2 The Current Phase Controller

In the half-bridge MOSFET driver circuit shown in Fig. 4.13, three resonant circuits are used. In part 2 of this figure, a series resonant LC circuit is used to produce a sinusoidal current, while in part 3, the parallel placement of the transformer windings, the capacitors, and the input capacitances of the MOSFETs, form parallel resonant LC circuits, producing sinusoidal voltage waveforms over the gate-source pins of the MOSFETs used in the half-bridge inverters.

Due to possible mismatches in the values of the capacitors and inductors used in these resonant circuits in the three hardware channels, it is possible that the sinusoidal currents in the series resonant circuits, and the sinusoidal gate-source voltages driving the MOSFETs, might not all be in phase, even though their respective driving signals might be.

For this reason, current phase controllers are used to estimate the phases of the primary winding currents, and adjust them, in order to set the correct primary winding current phases when operating in three-phase and variable-phase mode.

Since no disturbances in these signals are expected, simple and relatively slow acting integrator-type controllers are implemented.

Each primary current phase controller works by first calculating the error
between the current phase and the reference phase. If the phase error is positive, the phase of the clock signal driving the associated half-bridge MOSFET driver is increased. Similarly, if the phase error is negative, the phase of the clock signal is decreased.

The output of the primary current phase estimator circuit, $V_1$ in Fig. 4.17 is sampled at a rate of 125 MHz, which gives the controller an $8^\circ$ phase resolution, as discussed in subsection 4.3.2. The current phase controller however, is implemented at a relatively slow rate of 1 kHz, which translates to 1.00 ms per phase degree, or 360 ms for a complete $360^\circ$ phase revolution. The current phase controller thus operates at a rate 50 times slower than the current amplitude controller’s operating frequency of 50 kHz.

With no proportional component, and a relatively small integration gain, the current phase controller will not display the typical “step-like” response to a large phase error value, but instead, show a smoother and slower reaction. This smoother and slower compensator reaction will give the primary current amplitude compensators enough time to control and maintain a constant primary winding current, while the primary winding phases are adjusted. This is especially useful when the variable-phase CET platform encounters “dead spots” (subsection 3.6.2) where sudden changes in primary winding current phases can result in large secondary induced voltage swings, which in turn can affect the primary winding reflected voltages and the primary current amplitude controllers.

Since no specifications are presented for the current phase controller, the time length of 360 ms for one complete $360^\circ$ phase revolution is considered reasonable, and also in the order of the current amplitude controller’s settling time.

4.6 Discussions and Conclusions

The variable-phase CET platform has the ability to power small mobile electronic devices without the use of “plug-and-socket” connectors, but by using magnetic induction. In Chapter 3, a synthesis of the variable-phase CET platform is presented. As such, a solid theoretical background for the development of the variable-phase CET platform system is given. In order to experimentally verify the various results obtained through the calculations and simulations, however, the variable-phase CET platform needs to be physically implemented in the form of a hardware prototype system.

This chapter focusses on the implementation of the variable-phase CET platform, and discusses the design considerations taken during its development.

Firstly, an overview of the complete variable-phase CET platform is presented. The CET platform is designed in a modular fashion with various subsystems, each with a specific role to perform. Here, the CET platform is figuratively “broken-down” into its modular subsystems, in order to explain their operation and their interactions with the other systems.

Secondly, the primary-side subsystems are discussed. These systems and circuits make up the variable-phase CET platform and include the primary winding matrix, the primary controller, the quadratic buck converters, the high-frequency
MOSFET half-bridge drivers, the half-bridge inverters, the half-bridge output voltage amplitude and phase estimators, the winding current amplitude and phase estimators, as well as the winding commutation circuits.

Thirdly, the secondary-side subsystems are discussed. These systems and circuits receive power from the primary winding matrix and convert the AC-voltage signal induced in the secondary winding, into more useful DC-voltage levels for the mobile devices to use. These systems include the secondary winding, the full-bridge rectifier, the DC-to-DC converter, and the load.

The primary- and secondary-side subsystems are all implemented and separately tested. Their operations are verified through various measurements, in order to ensure that they work as expected, within the larger system.

Finally, the primary current amplitude and phase controllers are discussed. These compensators are used to control the primary winding current amplitudes and phases, in order for the variable-phase CET platform to operate in single-phase, three-phase or variable-phase mode. The primary current amplitude controller is implemented as a PI compensator. It is tested, and is shown to be able to maintain a relatively constant primary winding current amplitude for loaded and unloaded situations. However, with the current amplitude controller operating at only 50 kHz, much lower than the primary winding current frequency of 2.777 MHz, some current peaks are still visible during certain current controller step responses. The current controller settling time is approximately 500 ms.

Furthermore, the current phase controller is implemented as an integrator controller with a small integrator gain. An integrator controller is used to avoid the step-like responses produced by proportional control elements. In this way, the controller reacts in a smooth and slow-acting fashion, to large phase errors. With an effective current phase controller frequency of 1 kHz, the current amplitude controller still has enough time to maintain a steady primary current amplitude, while the current phase is adjusted. This helps to avoid certain problems when the variable-phase CET platform encounters “dead spots”. The current phase controller takes 360 ms for one complete 360° phase revolution, which is also in the order of the current amplitude controller’s settling time.
5
Experimental Verification

“The strongest arguments prove nothing so long as the conclusions are not verified by experience. Experimental science is the queen of sciences and the goal of all speculation.” - Roger Bacon (1214 - 1294)

5.1 Introduction

In the previous chapter, the implementation of the variable-phase CET platform is presented. To begin with, a system overview of the complete variable-phase CET platform is presented, and it is shown that the platform is designed in a modular fashion, consisting of various specialized subsystems. The system concept is broken down, and the subsystems, together with their interactions are discussed. Then, the primary-side subsystems and circuits are presented and discussed. These include the primary winding matrix, the primary controller, the quadratic buck converters, the high-frequency MOSFET half-bridge drivers, the half-bridge inverters, the primary winding current amplitude and phase estimator circuits, the half-bridge output voltage amplitude and phase estimators, as well as the winding commutation circuits. Next, the secondary-side subsystems and circuits are presented and discussed. These include the secondary winding, the full-bridge rectifier, the DC-to-DC converter, and the load. Further, the primary winding current amplitude and phase controllers are discussed, and finally, the variable-phase CET platform is implemented in a prototype system.

The purpose of this chapter is to verify the operation of the variable-phase CET platform through measurements performed on the implemented prototype.

In this chapter, firstly, the operation of the load detection process, as presented in section 3.5, is verified through various experimental measurements.
Here, various proof of principle measurements are performed, to show how conductive materials, soft-magnetic materials, and valid resonant load circuits influence the primary windings’ equivalent impedances. Afterwards, two sets of load detection experiments are performed to show how the CET platform can locate and distinguish between these various objects placed on top of the platform surface.

Secondly, measurements are performed while the CET platform is operating in single-phase mode. Here, all the windings inside the activated winding cluster are excited with in-phase currents. The secondary winding is placed at various positions above the winding cluster, while voltage and current measurements are taken for various power transfer levels. The measurements are then compared against calculated values. Furthermore, in order to obtain a quantitative representation of the stray magnetic field generated by the CET windings, a circular coil (sensing coil) is placed 100 mm above the center of the primary winding cluster. The sensing coil’s induced voltage is measured during each power transfer experiment and the results are stored for later comparison.

Thirdly, measurements are performed while the platform is operating in three-phase mode. Here, the windings inside the activated winding cluster are excited with currents of equal amplitude, but with 120° phase shifts. As in the above mentioned experiment, the secondary winding is placed at various positions above the winding cluster, while voltage and current measurements are taken for various power transfer levels. The measurement results are again compared against calculated values. Moreover, the voltages induced in the sensing coil are also measured and saved.

Fourthly, measurements are performed on the CET platform while it is operating in variable-phase mode. Here, the windings inside the activated winding cluster are excited with precalculated currents of different amplitudes and phases. These values, of which some are presented in subsection 3.6.5 and in Tables 3.6 and 3.7, are calculated to give minimum peak stray magnetic field values for certain secondary winding placements and power transfer levels. As in the two above mentioned experiments, the secondary winding is placed at various positions above the winding cluster, while voltage and current measurements are taken for various power transfer levels. The measurements are also compared against calculated values. Furthermore, the sensing coil’s induced voltage is measured during each power transfer experiment and the results are stored.

Fifthly, the sensing coil’s induced voltage results obtained from the three different measurement experiments are compared and their results are discussed.

Finally, the experimental verification of the variable-phase CET platform is concluded.

5.2 Load Detection

Load detection is the process in which the variable-phase CET platform attempts to find the positions of load devices placed on the winding matrix surface. The load detection process can locate and distinguish between three types of ob-
jacts placed on the CET platform, namely: valid resonant load devices, magnetic materials, and conductive materials.

To demonstrate the load detection method, as presented in section 3.5, various experiments are conducted on the prototype.

Firstly, various proof of principle measurements are performed. These measurements show how the equivalent impedance of a primary winding is influenced by the presence (or lack) of magnetic materials, conductive materials, and valid resonant loads placed on top of it.

Secondly, the complete load detection process is carried out on the variable-phase CET platform, and it is shown that the prototype can locate and distinguish the three different types of objects placed on the winding matrix.

5.2.1 Proof of Principle Measurements

The proof of principle measurements are performed on the variable-phase CET platform to show actual primary winding voltage and current waveforms during the various stages of the load detection process.

Figures 5.1 - 5.4 show voltage and current waveforms captured by the oscilloscope (Tektronix TDS 744 A) during four different load detection phases.

Firstly, measurements are taken during the calibration of the primary winding $W_A$. Here, no load devices are placed on, or near the winding. Figure 5.1 shows the measurement results. In this figure, trace 1 is the output of the half-bridge inverter, and trace 2 is the output of the voltage phase estimator circuit. Trace 3 is the output of the current phase estimator, and trace 4 is the primary winding current, measured using a Tektronix TM501 - AM503 current probe. Here, the primary winding current amplitude is 0.5 A (peak), and the phase is $220^\circ$, according to the current phase estimator. The half-bridge square wave voltage signal is approximately 6.10 V (peak-to-peak), which gives a 3.88 V (peak) fundamental voltage. The oscilloscope is triggered on the output of the voltage phase estimator circuit, so the voltage signal has a phase of $0^\circ$. The calibrated impedance of $W_A$ is thus 7.77 $\Omega$ with a $-220^\circ$ phase angle.

Secondly, measurements are taken when a sheet of copper (piece of conductive material) is placed on top of winding $W_A$. Figure 5.2 shows the measurements, and the trace definitions are the same as in the previously shown calibration measurement figure. Here, the primary winding current is measured at 0.5 A (peak), with a phase angle of $320^\circ$ according to the current phase estimator. The half-bridge square wave voltage signal is about 18.00 V (peak-to-peak), which gives a 11.46 V (peak) fundamental voltage. The oscilloscope is triggered on the output of the voltage phase estimator circuit, so the voltage signal has a phase of $0^\circ$. The calibrated impedance of $W_A$ is thus 22.92 $\Omega$ with a $-320^\circ$ phase angle.

Thirdly, measurements are taken when two ferrite E-cores (material ETD39-N87 [107]) are placed on top of winding $W_A$. Figure 5.3 shows the measurement results, and the trace definitions are the same as in the previous two figures. Here, the primary winding current is approximately 0.5 A (peak), with a phase angle of $152^\circ$ according to the current phase estimator. The half-bridge square wave voltage signal is about 27.00 V (peak-to-peak), which gives a 17.19 V (peak)
Figure 5.1: Voltage and current measurements during calibration of winding $W_A$. Trace 1: output voltage of the half-bridge inverter (10 V/div); trace 2: output of the voltage phase estimator circuit (2 V/div); trace 3: output of the current phase estimator circuit (5 V/div); trace 4: primary winding current (0.5 A/div). Timescale: 100 ns/div.

Figure 5.2: Voltage and current measurements with conductive material placed on top of winding $W_A$. The trace definitions is the same as in Fig. 5.1.

fundamental voltage. With the voltage phase angle at $0^\circ$, the impedance of the
winding, when it is loaded with magnetic material, is estimated as 34.38 Ω with a -152° phase angle.

Figure 5.3: Voltage and current measurements with magnetic material placed on top of winding \( W_A \). The trace definitions are the same as in Fig. 5.1.

Finally, measurements are taken when a resonant load is placed on top of winding \( W_A \). Here, the secondary winding is connected to its resonant capacitor, the full-bridge rectifier, and a variable-resistor. Figure 5.4 shows the measurements. Here, trace 1 is the output of the voltage phase estimator circuit, and trace 2 is the output of the current phase estimator. Trace 3 is the load voltage, and trace 4 is the load current. With the load current at 179 mA (DC) and the load voltage at about 25 V (DC), approximately 4.50 W of power is transferred to the load resistor. The primary winding current amplitude is approximately 0.6 A (peak), with a phase angle of approx. 223°, according to the current phase estimator. The half-bridge output square wave voltage signal is about 36.70 V (peak-to-peak), which gives a 23.36 V (peak) fundamental voltage. The primary winding current amplitude is 0.6 A (peak). With the voltage phase angle at 0°, the impedance of the winding, when it is loaded with a resonant load device, is estimated as 38.94 Ω with a -223° phase angle.

The impedance results from the four experiments are summarized and shown in Table 5.1. The results illustrate that the conductive material causes the equivalent winding impedance amplitude to rise, and the impedance phase to decrease, with 195 % and 45 %, respectively. Furthermore, the results show that the magnetic material causes the equivalent winding impedance amplitude and phase both to increase with 342 % and 31 %, respectively. Finally, the results also show that the valid resonant load, causes the equivalent winding impedance amplitude to increase with 401 %, while the impedance phase changed with a negligible 1.36 %.
Figure 5.4: Voltage and current measurements while approx. 4.50 W is transferred to a resonant load placed above winding $W_A$. Trace 1: output of the voltage phase estimator circuit (2 V/div); trace 2: output of the current phase estimator (5 V/div); trace 4: load voltage (10 V/div); trace 3: load current (100 mA/div).

Table 5.1: A summary of the equivalent winding impedances measured during the proof of principle load detection experiments performed on the variable-phase CET platform.

<table>
<thead>
<tr>
<th>Load detection phase</th>
<th>Winding impedance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Calibration</td>
<td>$7.77 , \Omega \angle -220^\circ$</td>
</tr>
<tr>
<td>Conductive load</td>
<td>$22.92 , \Omega \angle -320^\circ$</td>
</tr>
<tr>
<td>Magnetic load</td>
<td>$34.38 , \Omega \angle -152^\circ$</td>
</tr>
<tr>
<td>Resonant load</td>
<td>$38.94 , \Omega \angle -223^\circ$</td>
</tr>
</tbody>
</table>

Concluding, the four proof of principle measurements in Table. 5.1 confirm that the three different load types influence the equivalent winding impedance of winding $W_A$, as predicted in section 3.5. The results show predictable and measurable changes in the equivalent winding impedance that can be used by the variable-phase CET platform to locate and distinguish between these different object types.

5.2.2 Primary Winding Calibration

The first step in the load detection process is the CET platform calibration. The platform is calibrated to determine the exact primary winding impedances, when the platform is void of any conductive materials, soft-magnetic materials,
5.2. Load Detection

or resonant load devices. The eighteen primary windings are individually excited and their impedances are estimated through measurements. The values are shown in Table 5.2.

Table 5.2: Calibrated primary winding impedances estimated during the calibration process.

| Winding | Measured impedance $|X_n^C|\angle\phi_n^C$ |
|---------|------------------------------------------|
| $W_A$   | $7.76 \Omega \angle -219^\circ$          |
| $W_B$   | $8.91 \Omega \angle -186^\circ$          |
| $W_C$   | $8.47 \Omega \angle -184^\circ$          |
| $W_D$   | $8.02 \Omega \angle -202^\circ$          |
| $W_E$   | $9.80 \Omega \angle -192^\circ$          |
| $W_F$   | $7.07 \Omega \angle -230^\circ$          |
| $W_G$   | $7.38 \Omega \angle -204^\circ$          |
| $W_I$   | $7.07 \Omega \angle -232^\circ$          |
| $W_J$   | $7.64 \Omega \angle -204^\circ$          |
| $W_K$   | $8.91 \Omega \angle -186^\circ$          |
| $W_L$   | $7.77 \Omega \angle -212^\circ$          |
| $W_M$   | $7.64 \Omega \angle -196^\circ$          |
| $W_N$   | $8.91 \Omega \angle -186^\circ$          |
| $W_O$   | $8.60 \Omega \angle -206^\circ$          |
| $W_P$   | $9.29 \Omega \angle -186^\circ$          |
| $W_Q$   | $6.62 \Omega \angle -218^\circ$          |
| $W_R$   | $7.92 \Omega \angle -196^\circ$          |
| $W_S$   | $8.66 \Omega \angle -180^\circ$          |

From these results, and the results shown in Table 5.1, it can be seen that the measured impedances do not have zero phase angles as one would expect from resonant RLC-circuits. This is the result of the various delays in the current phase and voltage phase estimation circuits. However, since only relative changes in current and voltage phases are important for the load detection process, this is not a problem.

5.2.3 Conductive Material, Ferrites, and Soft-Magnetic Material Detection

In the first set of experiments, various conductive and ferrite materials are placed on the CET platform. Figure 5.5 (a) shows the arrangement of the various materials placed over the CET winding matrix. Figure 5.5 (b) shows the image of the actual materials placed onto the CET prototype.

In Fig. 5.5, object A is an office key (60 mm long, from aluminium), object B is a ferrite toroidal core (radius = 10 mm, material is 3C11, $\mu_r = 4150$, [108]), object C are two ferrite cores (material ETD39-N67 [109]), and object D is a piece of copper plate (45 mm x 45 mm).

The load position detection process is conducted and the results are shown in Fig. 5.6 and Fig. 5.7. In these two figures, the solid lines show the calibrated values, while the dotted lines are the calibration threshold values. The dots show the measured impedance amplitude and phase values.

Regarding the load-position detection: From the measured results it can be seen that windings $W_A$, $W_C$, $W_E$, $W_L$, $W_M$, $W_N$, $W_P$, $W_Q$, and $W_R$ clearly fall outside the threshold limits. These are the possible load positions as determined
Figure 5.5: (a) The placement of various ferrites and conductive materials during the load detection experiments, and (b) an image of the object placements.

Figure 5.6: Impedance amplitude results for the load detection experiment involving conductive and magnetic materials.

by the load-position detection. They all correspond correctly with objects placed on the CET platform.

Regarding the load-type estimation: Windings $W_A$ and $W_E$, corresponding to the placement of the metallic key, both show an increase in the impedance amplitude and a decrease in the impedance phase. Winding $W_C$, corresponding to the toroidal ferrite core placement, shows an increase in both the impedance amplitude as well as phase. Windings $W_L$ and $W_P$, corresponding to the ferrite E-core placement, both show an increase in the impedance amplitude and impedance phase values. Windings $W_M$, $W_N$, $W_Q$, and $W_R$ corresponding to the
5.2. Load Detection

Figure 5.7: Impedance phase results for the load detection experiment involving conductive and magnetic materials.

large copper plate placement, all show impedance amplitude increase, as well as impedance phase decrease.

The results from the load-type estimation experiment, for conductive and ferrite materials, all show correct and measurable changes in impedance amplitude and impedance phase values as predicted in section 3.5.

5.2.4 Resonant Load Detection

In the second set of experiments, the secondary winding $W_Z$, is connected to its resonant capacitor, the full-bridge rectifier and a variable-resistor. Together, the secondary circuit and load are referred to as the resonant load, which is placed at different locations on the CET platform. Figure 5.8 (a) shows the positions of the different resonant load placements during the experiments. Figure 5.8 (b) shows an image of the resonant load placed at position $F$.

The load resistances are individually tuned so that in every test, the circuit will transfer approximately 3 W to 5 W of power to the load, at primary winding currents of about $I_P = 500$ mA - 750 mA (RMS).

The load-position detection process is conducted and the results are shown in Fig. 5.9 and Fig. 5.10. In these two figures, the solid lines show the calibrated values, while the dotted lines are the calibration threshold values. The dots show the measured impedance and phase values.

From these results, it can be seen that the impedance amplitudes of windings $W_A, W_J, W_K, W_N, W_Q$ and $W_R$ clearly fall outside the threshold limits. These are the possible load positions as determined by the load-position detection. They all correspond correctly with the secondary windings placed on the CET platform.

Regarding the load-type estimation, all the windings show similar results. They all show definite increases in impedance amplitudes, while impedance phase
Figure 5.8: (a) The placement of different resonant loads during the load detection experiments, and (b), an image of the resonant load placed at position $F$.

Figure 5.9: Impedance amplitude results for the load detection experiment involving valid resonant loads.

values are all within the threshold limits.

The results from the load-type estimation test for resonant load devices, all show correct and measurable changes in impedance amplitude and impedance phase values as predicted in section 3.5.
5.3 Single-Phase Power Transfer Experiments

The single-phase power transfer experiments are performed on the variable-phase CET platform in order to measure and verify power transfer values when the system is operating in single-phase mode. Here, all the windings inside the activated winding cluster are excited with in-phase currents with the same amplitude. The secondary winding is placed at various position above the winding cluster, while voltage and current measurements are taken for various power transfer levels.

The primary winding cluster used in this experiment, consists of primary windings $W_F, W_J,$ and $W_K,$ as shown in Fig. 4.4 and Fig. 5.11 (a). Figure 5.11 (b) shows the primary windings’ associated hardware channels.

The secondary winding $W_Z,$ is connected to its resonant capacitor, full-bridge rectifier and a variable-resistor, and is placed at one of three different locations above the primary winding cluster during the measurements. These three positions are labeled $P_1, P_2$ and $P_3,$ and are shown in Fig. 5.11 (c). At each secondary winding position, power levels of 0 W, 2.5 W, and 5 W are transferred to the load resistor.

In order to verify the various measurements, power transfer calculations are performed, but to do so, the various inter-winding mutual inductances must first be determined. The primary winding matrix is covered in a layer of insulating cardboard, with a thickness of approx 1.5 mm. The mutual inductances between the primary windings and secondary winding placed at the different positions are first calculated and then measured, and the results are shown in Table 5.3. As it can be seen, the results are in good agreement, with a maximum error of 5% between calculated and measured values.

![Figure 5.10: Impedance phase results for the load detection experiment involving valid resonant loads.](image-url)
Figure 5.11: (a) The primary winding cluster consists of windings $W_F$, $W_J$, and $W_K$, while winding $W_E$ is used for the “dead spot” detection and removal during three-phase and variable-phase operations. (b) The hardware channels associated with the primary windings. (c) The three secondary winding positions $P_1$, $P_2$ and $P_3$.

Table 5.3: Measured and calculated, primary-to-secondary winding mutual inductances.

<table>
<thead>
<tr>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td>$M_{EZ} [\mu H]$</td>
<td>$P_1$</td>
</tr>
<tr>
<td>-0.145</td>
<td>-0.168</td>
</tr>
<tr>
<td>$M_{FZ} [\mu H]$</td>
<td>2.470</td>
</tr>
<tr>
<td>$M_{JZ} [\mu H]$</td>
<td>-0.155</td>
</tr>
<tr>
<td>$M_{KZ} [\mu H]$</td>
<td>-0.147</td>
</tr>
</tbody>
</table>

Furthermore, in order to demonstrate the effects of the destructive wave interference on the magnetic field created by the variable-phase CET platform, the stray magnetic field is indirectly observed by measuring the voltage induced in a circular coil placed above the primary winding cluster. In this way, a quantitative representation of the stray magnetic field generated by the CET windings is obtained. The circular sensing coil is labeled $W_X$, and has a radius of 40 mm and 3 turns. It is placed above and in parallel to the primary CET platform surface at a height of 100 mm. Figure 5.12 (a) shows a three-dimensional view of the circular sensing coil in relation to the primary windings, and Fig. 5.12 (b) shows a two-dimensional (top view) of the sensing coil and the primary windings. Moreover, Fig. 5.13 shows a photo of the sensing coil placed above the primary winding matrix.

The mutual inductances between the primary windings and the sensing coil are calculated and measured, and shown Table 5.4. Moreover, the mutual inductances between the secondary winding placed at the three different test positions, and the sensing coil are also calculated and is shown in Table 5.5. These values will be used in the following sections for the indirect measurement of the stray
5.3. Single-Phase Power Transfer Experiments

Figure 5.12: (a) A three-dimensional view of the circular sensing coil in relation to the primary windings. (b) A two-dimensional (top view) of the sensing coil and the primary windings.

Figure 5.13: The implemented stray field sensing coil above the CET platform’s primary winding matrix.

The first set of single-phase experiments are performed while the secondary winding $W_Z$, is placed at position $P_1$. Various currents and voltages are measured while the CET platform is transferring approx. 0 W, 2.5 W and 5 W to the load resistor. The results are shown in Table 5.6.

The various parameters in Table 5.6 can be explained as follows: on the primary side $|I_F|$, $|I_J|$, and $|I_K|$, are the current amplitudes of the three primary windings, respectively. Similarly, $\angle \theta_F$, $\angle \theta_J$, and $\angle \theta_K$, are the phase angles (in degrees) of the currents in the three primary windings, respectively. The voltages $V_{DC}$ (I), $V_{DC}$ (II), and $V_{DC}$ (III), are the DC-voltage outputs of the
Table 5.4: Measured and calculated, primary winding-to-sensing coil mutual inductances.

<table>
<thead>
<tr>
<th></th>
<th>Measured mutual inductance</th>
<th>Calculated mutual inductance</th>
</tr>
</thead>
<tbody>
<tr>
<td>$M_{EX}$</td>
<td>10.10 nH</td>
<td>10.40 nH</td>
</tr>
<tr>
<td>$M_{FX}$</td>
<td>10.00 nH</td>
<td>10.84 nH</td>
</tr>
<tr>
<td>$M_{JX}$</td>
<td>10.10 nH</td>
<td>10.84 nH</td>
</tr>
<tr>
<td>$M_{KX}$</td>
<td>10.40 nH</td>
<td>10.84 nH</td>
</tr>
</tbody>
</table>

Table 5.5: Calculated secondary winding-to-sensing coil mutual inductances.

<table>
<thead>
<tr>
<th></th>
<th>Calculated mutual inductance, $M_{ZX}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_1$</td>
<td>22.19 nH</td>
</tr>
<tr>
<td>$P_2$</td>
<td>22.93 nH</td>
</tr>
<tr>
<td>$P_3$</td>
<td>22.74 nH</td>
</tr>
</tbody>
</table>

three channels’ quadratic buck converters, respectively (The roman numerals in brackets indicate the associated hardware channel).

On the secondary side, voltage $V_L$ is the load voltage, and current $I_L$ is the load current. The load resistance is $R_L$ and the power dissipated in the load resistor is $P_L$. Furthermore, the power transfer efficiency between the primary coils and the load resistance is given as $\eta_{CET}$ and finally, the sensing coil’s induced voltage is presented as $V_X$.

In the second set of experiments, the secondary winding $W_Z$ is placed on the CET platform surface at position $P_2$ as indicated in Fig. 5.11 (c), while 0 W, 2.5 W and 5 W of power is transferred to the load resistor. The results are shown in Table 5.7.

In the third set of experiments, the secondary winding $W_Z$, is placed on the CET platform surface at position $P_3$ as indicated in Fig. 5.11 (c), while 0 W, 2.5 W, and 5 W of power is transferred to the load resistor. The results are shown in Table 5.8.

In Tables 5.6 - 5.8, as well as the tables providing experimental results in the following two sections, the abbreviations “o/c” mean open circuit, and is used in the 0 W power transfer tests, indicating that the load resistor is disconnected from the secondary circuit.

The measured results as shown in Tables 5.6 - 5.8 all show good agreement with the calculated values. The measured and calculated secondary winding current and voltage values, as well as the sensing coil’s induced voltages are all accurate within 5 %. The primary circuits’ voltage measurements and calculated
Table 5.6: Measured and calculated circuit parameters for single-phase CET power transfer, with the secondary winding placed at position $P_1$. All the current and voltages values are in RMS.

<table>
<thead>
<tr>
<th>Primary windings</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
<td>$ [mA]</td>
</tr>
<tr>
<td>$\angle \theta_F$ [deg]</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC}$ (I) [V]</td>
<td>18.46</td>
<td>31.14</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>$ [mA]</td>
</tr>
<tr>
<td>$\angle \theta_J$ [deg]</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
<td>$ [mA]</td>
</tr>
<tr>
<td>$\angle \theta_K$ [deg]</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC}$ (III) [V]</td>
<td>18.83</td>
<td>18.45</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Secondary winding</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
<td>$ [V]</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>$ [mA]</td>
</tr>
<tr>
<td>$P_L$ [W]</td>
<td>~0</td>
<td>2.50</td>
</tr>
<tr>
<td>$R_L$ [Ω]</td>
<td>o/c</td>
<td>355</td>
</tr>
<tr>
<td>$\eta_{CET}$ [%]</td>
<td>0 %</td>
<td>35 %</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Sensing coil</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
<td>$ [mV]</td>
</tr>
</tbody>
</table>

values are accurate with in 20 %. The differences here can all be contributed to measurement tolerances and primary circuits’ component tolerances. It should be noted that the primary currents are imposed, and that all the uncertainties in the primary impedances are reflected in the adjustments of the primary DC voltages.
Table 5.7: Measured and calculated circuit parameters for single-phase CET power transfer, with the secondary winding placed at position $P_2$. All the current and voltages values are in RMS.

<table>
<thead>
<tr>
<th>Primary windings</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC} (I) [V]$</td>
<td>18.50</td>
<td>19.50</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC} (II) [V]$</td>
<td>21.68</td>
<td>26.00</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_K [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC} (III) [V]$</td>
<td>19.06</td>
<td>23.21</td>
</tr>
</tbody>
</table>

<table>
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<tr>
<th>Secondary winding</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
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<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
<td>[V]$</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$P_L [W]$</td>
<td>~0</td>
<td>2.50</td>
</tr>
<tr>
<td>$R_L [\Omega]$</td>
<td>o/c</td>
<td>481</td>
</tr>
<tr>
<td>$\eta_{CET} [%]$</td>
<td>0 %</td>
<td>35 %</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Sensing coil</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
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<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
<td>[\text{mV}]$</td>
</tr>
</tbody>
</table>
Table 5.8: Measured and calculated circuit parameters for single-phase CET power transfer, with the secondary winding placed at position $P_3$. All the current and voltages values are in RMS.

### Primary windings

<table>
<thead>
<tr>
<th></th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
<td>:[mA]$</td>
</tr>
<tr>
<td>$\angle\theta_F:[\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC\ (I)}:[V]$</td>
<td>18.60</td>
<td>23.20</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>:[mA]$</td>
</tr>
<tr>
<td>$\angle\theta_J:[\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC\ (II)}:[V]$</td>
<td>22.28</td>
<td>27.37</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
<td>:[mA]$</td>
</tr>
<tr>
<td>$\angle\theta_K:[\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC\ (III)}:[V]$</td>
<td>18.80</td>
<td>19.35</td>
</tr>
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### Secondary winding

<table>
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<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
<td>:[V]$</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>:[mA]$</td>
</tr>
<tr>
<td>$P_L:[W]$</td>
<td>~0</td>
<td>2.50</td>
</tr>
<tr>
<td>$R_L:[\Omega]$</td>
<td>o/c</td>
<td>470</td>
</tr>
<tr>
<td>$\eta_{CET}:[%]$</td>
<td>0 %</td>
<td>35 %</td>
</tr>
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</table>

### Sensing coil

<table>
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<tr>
<th></th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
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<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
<td>:[mV]$</td>
</tr>
</tbody>
</table>
5.4 Three-Phase Power Transfer Experiments

Power transfer experiments are performed on the variable-phase CET platform in order to measure and verify power transfer values when the system is operating in three-phase mode. Here, all the windings inside the activated winding cluster are excited with currents of the same amplitude, but with 120° phase shifts. The secondary winding is placed at various position above the winding cluster, while voltage and current measurements are taken for various power transfer levels.

The first set of three-phase experiments are performed while the secondary winding $W_Z$, is placed at position $P_1$ as shown in Fig. 5.11 (c). Here, the primary winding cluster consists of the same primary windings as in the single-phase experiments, namely, $W_F$, $W_J$, and $W_K$, as shown in Fig. 5.11 (a). The various currents and voltages are measured while the CET platform transfers approx. 0 W, 2.5 W and 5.0 W of power to the load resistor. The results are shown in Table 5.9. Here, the various parameters are the same as in the tables presenting the single-phase experimental results.

In the second set of experiments, the secondary winding $W_Z$, is placed on the CET platform surface at position $P_2$ as indicated in Fig. 5.11 (c), while 0 W, 2.5 W and 5.0 W of power is transferred to the load resistor. The “dead spot” present at this position is removed by disabling primary winding $W_K$, and using winding $W_E$, instead. The results are shown in Table 5.10.

In the third set of experiments, the secondary winding $W_Z$, is placed on the CET platform surface at position $P_3$ as shown in Fig. 5.11 (c), while 0 W, 2.5 W, and 5.0 W of power is transferred to the load resistor. Here, the original primary winding cluster, consisting of windings $W_F$, $W_J$, and $W_K$ are used. The results are shown in Table 5.11.

The measured results as shown in Tables 5.9 - 5.11 all show acceptable agreement with the calculated values. The measured and calculated secondary winding current and voltage values are accurate within 5 %. The primary circuits’ current and voltage measurements and calculated values however, show again less accurate results with up to 20 % differences for some values. Here, as explained earlier, the differences are contributed to by the primary circuits’ component tolerances as well as measurement tolerances. The sensing coil’s induced voltage measurements and calculations also show less accurate results with up to 30 % differences. The major contributing factor to these differences are possible non-exact 120° phase shifts between the primary currents during the measurements. When the three phases are not completely balanced the resultant field will contribute to the stray magnetic field, while not necessarily interfering with the power transfer results.
Table 5.9: Measured and calculated circuit parameters for three-phase CET power transfer, with the secondary winding placed at position $P_1$. All the current and voltages values are in RMS.

<table>
<thead>
<tr>
<th>Primary windings</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td>$</td>
<td>I_F</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
<td>0°</td>
<td>0° 0° 0°</td>
</tr>
<tr>
<td>$V_{DC} (\text{I}) [V]$</td>
<td>25.16</td>
<td>15.93 33.14</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>120°</td>
<td>120° 120° 120°</td>
</tr>
<tr>
<td>$V_{DC} (\text{II}) [V]$</td>
<td>19.77</td>
<td>18.15 19.02</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_K [\text{deg}]$</td>
<td>240°</td>
<td>240° 240° 240°</td>
</tr>
<tr>
<td>$V_{DC} (\text{III}) [V]$</td>
<td>21.74</td>
<td>21.13 20.70</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
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<th>Calculations</th>
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</thead>
<tbody>
<tr>
<td>$</td>
<td>V_L</td>
<td>[V]$</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$P_L [W]$</td>
<td>∼0</td>
<td>0 W 2.5 W 5.0 W</td>
</tr>
<tr>
<td>$R_L [\Omega]$</td>
<td>o/c</td>
<td>o/c o/c o/c</td>
</tr>
<tr>
<td>$\eta_{CET} [%]$</td>
<td>0 %</td>
<td>0 % 35 % 51 %</td>
</tr>
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<table>
<thead>
<tr>
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<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td>$</td>
<td>V_X</td>
<td>[mV]$</td>
</tr>
</tbody>
</table>

Table 5.10: Measured and calculated circuit parameters for three-phase CET power transfer, with the secondary winding placed at position $P_2$. All the current and voltages values are in RMS.

<table>
<thead>
<tr>
<th>Primary windings</th>
<th>Measurements</th>
<th>Calculations</th>
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<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
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<tr>
<td>$</td>
<td>I_F</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC} (I) [\text{V}]$</td>
<td>17.00</td>
<td>23.92</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>120°</td>
<td>120°</td>
</tr>
<tr>
<td>$V_{DC} (II) [\text{V}]$</td>
<td>18.40</td>
<td>22.57</td>
</tr>
<tr>
<td>$</td>
<td>I_E</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_E [\text{deg}]$</td>
<td>240°</td>
<td>240°</td>
</tr>
<tr>
<td>$V_{DC} (III) [\text{V}]$</td>
<td>21.61</td>
<td>26.66</td>
</tr>
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</table>

<table>
<thead>
<tr>
<th>Secondary winding</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
<td>[\text{V}]$</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>[\text{mA}]$</td>
</tr>
<tr>
<td>$P_L [\text{W}]$</td>
<td>~0</td>
<td>2.50</td>
</tr>
<tr>
<td>$R_L [\Omega]$</td>
<td>o/c</td>
<td>70</td>
</tr>
<tr>
<td>$\eta_{CET} [%]$</td>
<td>0 %</td>
<td>34 %</td>
</tr>
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</table>

<table>
<thead>
<tr>
<th>Sensing coil</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
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<td></td>
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<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
<td>[\text{mV}]$</td>
</tr>
</tbody>
</table>
Table 5.11: Measured and calculated circuit parameters for three-phase CET power transfer, with the secondary winding placed at position $P_3$. All the current and voltages values are in RMS.

### Primary windings

<table>
<thead>
<tr>
<th>Measurements</th>
<th>Calculations</th>
</tr>
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<tbody>
<tr>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
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<tr>
<td>$V_{DC} \text{ (I) [V]}$</td>
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</tr>
<tr>
<td>$</td>
<td>I_J</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>120°</td>
</tr>
<tr>
<td>$V_{DC} \text{ (II) [V]}$</td>
<td>20.33</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
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<tr>
<td>$\angle \theta_K [\text{deg}]$</td>
<td>240°</td>
</tr>
<tr>
<td>$V_{DC} \text{ (III) [V]}$</td>
<td>26.54</td>
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### Secondary winding

<table>
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<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
</tr>
<tr>
<td>$P_L \text{ [W]}$</td>
<td>~0</td>
</tr>
<tr>
<td>$R_L \text{ [}\Omega\text{]}$</td>
<td>o/c</td>
</tr>
<tr>
<td>$\eta_{CET} \text{ [%]}$</td>
<td>0 %</td>
</tr>
</tbody>
</table>

### Sensing coil

<table>
<thead>
<tr>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
</tr>
</tbody>
</table>
5.5 Variable-Phase Power Transfer Experiments

Also, power transfer experiments are performed on the variable-phase CET platform in order to measure and verify the power transfer values when the system is operating in variable-phase mode. Here, the windings inside the activated winding cluster are excited with precalculated currents of different amplitudes and phases. These values, of which some are presented in subsection 3.6.5, are calculated to give minimum peak stray magnetic field values for certain secondary winding placements and power transfer levels. The secondary winding is placed at various positions above the winding cluster, while voltage and current measurements are taken for various power transfer levels.

The first set of variable-phase experiments are performed, while the secondary winding $W_Z$, is placed at position $P_1$ as shown in Fig. 5.11 (c). Here, the primary winding cluster consists of the original primary windings, namely $W_F$, $W_J$, and $W_K$, as shown in Fig. 5.11 (a). The various currents and voltages are measured while the CET platform transfers approx. 0 W, 2.5 W and 5.0 W of power to the load resistor. The results are shown in Table 5.12. Here, the various parameters are the same as in the single-phase and three-phase result tables.

In the second set of experiments, the secondary winding $W_Z$, is placed on the CET platform surface at position $P_2$ as indicated in Fig. 5.11 (c), while 0 W, 2.5 W and 5.0 W of power is transferred to the load resistor. The “dead spot” present at this location is removed by disabling primary winding $W_K$, while using winding $W_E$, in its stead. The results are shown in Table 5.13.

In the third set of experiments, the secondary winding $W_Z$, is placed on the CET platform surface at position $P_3$ as indicated in Fig. 5.11 (c), while 0 W, 2.5 W, and 5.0 W of power is transferred to the load resistor. Here, the original primary winding cluster consisting of windings $W_F$, $W_J$, and $W_K$, as shown in Fig. 5.11 (a), is used. The results are presented in Table 5.14.

The measured results as shown in Tables 5.12 - 5.14 all show acceptable agreement with the calculated values. The measured and calculated secondary winding current and voltage values are accurate within 10 %. The primary circuits’ current and voltage measurements and calculated values however, show less accurate results with up to 20 % differences for some values. Here, again, the differences are contributed to by the primary circuits’ component tolerances as well as measurement tolerances. The sensing coil’s induced voltage measurements and calculations also show results with up to 30 % differences. Since the attenuation of the stray magnetic field is very much dependent on the primary winding current amplitudes and phases, the major contributing factor to these differences are possible non-exact implemented current amplitudes and phase shifts during the measurements.
Table 5.12: Measured and calculated circuit parameters for variable-phase CET power transfer, with the secondary winding placed at position $P_1$. All the current and voltages values are in RMS.

<table>
<thead>
<tr>
<th>Primary windings</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
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<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC} \text{ (I) [V]}$</td>
<td>5.46</td>
<td>25.32</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>180°</td>
<td>215°</td>
</tr>
<tr>
<td>$V_{DC} \text{ (II) [V]}$</td>
<td>2.23</td>
<td>5.03</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_K [\text{deg}]$</td>
<td>180°</td>
<td>215°</td>
</tr>
<tr>
<td>$V_{DC} \text{ (III) [V]}$</td>
<td>2.68</td>
<td>5.80</td>
</tr>
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</table>

<table>
<thead>
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<th>Secondary winding</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
<td>[\text{V dc}]$</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>[\text{mA dc}]$</td>
</tr>
<tr>
<td>$P_L [\text{W}]$</td>
<td>~0</td>
<td>2.50</td>
</tr>
<tr>
<td>$R_L [\Omega]$</td>
<td>o/c</td>
<td>132</td>
</tr>
<tr>
<td>$\eta_{CET} [%]$</td>
<td>0 %</td>
<td>71 %</td>
</tr>
</tbody>
</table>

<table>
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<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
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<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
<td>[\text{mV}]$</td>
</tr>
</tbody>
</table>
Table 5.13: Measured and calculated circuit parameters for variable-phase CET power transfer, with the secondary winding placed at position $P_2$. All the current and voltages values are in RMS.

<table>
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<tr>
<th>Primary windings</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC}$ (I) [V]</td>
<td>6.48</td>
<td>21.50</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>0°</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC}$ (II) [V]</td>
<td>7.43</td>
<td>24.30</td>
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<tr>
<td>$</td>
<td>I_E</td>
<td>[mA]$</td>
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<tr>
<td>$\angle \theta_E [\text{deg}]$</td>
<td>180°</td>
<td>230°</td>
</tr>
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<td>$V_{DC}$ (III) [V]</td>
<td>14.96</td>
<td>17.51</td>
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<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
<td>[V]$</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
<td>[mA]$</td>
</tr>
<tr>
<td>$P_L [W]$</td>
<td>~0</td>
<td>2.50</td>
</tr>
<tr>
<td>$R_L [\Omega]$</td>
<td>o/c</td>
<td>49</td>
</tr>
<tr>
<td>$\eta_{CET} [%]$</td>
<td>0 %</td>
<td>45 %</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Sensing coil</th>
<th>Measurements</th>
<th>Calculations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
<td>[mV]$</td>
</tr>
</tbody>
</table>
Table 5.14: Measured and calculated circuit parameters for variable-phase CET power transfer, with the secondary winding placed at position $P_3$. All the current and voltages values are in RMS.

<table>
<thead>
<tr>
<th>Primary windings</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Measurements</td>
<td>Calculations</td>
</tr>
<tr>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>I_F</td>
</tr>
<tr>
<td>$\angle \theta_F [\text{deg}]$</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC}$ (I) [V]</td>
<td>5.39</td>
</tr>
<tr>
<td>$</td>
<td>I_J</td>
</tr>
<tr>
<td>$\angle \theta_J [\text{deg}]$</td>
<td>0°</td>
</tr>
<tr>
<td>$V_{DC}$ (II) [V]</td>
<td>5.96</td>
</tr>
<tr>
<td>$</td>
<td>I_K</td>
</tr>
<tr>
<td>$\angle \theta_K [\text{deg}]$</td>
<td>180°</td>
</tr>
<tr>
<td>$V_{DC}$ (III) [V]</td>
<td>13.05</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Secondary winding</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Measurements</td>
<td>Calculations</td>
</tr>
<tr>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_L</td>
</tr>
<tr>
<td>$</td>
<td>I_L</td>
</tr>
<tr>
<td>$P_L$ [W]</td>
<td>~ 0</td>
</tr>
<tr>
<td>$R_L$ [Ω]</td>
<td>o/c</td>
</tr>
<tr>
<td>$\eta_{CET}$ [%]</td>
<td>0 %</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Sensing coil</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Measurements</td>
<td>Calculations</td>
</tr>
<tr>
<td>0 W</td>
<td>2.5 W</td>
</tr>
<tr>
<td>$</td>
<td>V_X</td>
</tr>
</tbody>
</table>
5.6 Stray Magnetic Field Comparisons

In section 3.6, the concept of stray magnetic field attenuation through the use of destructive wave interference is presented and discussed. Through various simulations it is concluded in Fig. 3.37, that the peak stray magnetic field generated by the CET platform can be reduced (for the same amount of power transferred to the load) when the system is operating in three-phase and variable-phase mode.

The peak stray magnetic field values for the three operational modes of the CET platform is calculated and shown in Fig. 3.30, Fig. 3.33, and Fig. 3.36. In these graphs, the magnetic field values are mostly in the order of one micro tesla, but go as low as 0.1 micro tesla at some points.

These values are extremely small and difficult to accurately measure using hand-held magnetic field meters, so instead of direct measurements, the stray magnetic fields are rather indirectly measured using a sensing coil (as presented in the previous sections). By measuring the induced voltage in a sensing coil with 40 mm radius and 3 turns, a quantitative representation of the stray magnetic field is obtained. Moreover, the induced voltages are in the range of 10 - 400 millivolts which can easily be measured using an oscilloscope.

The sensing coil’s induced voltages as measured in the various power transfer experiments and presented in the previous 3 sections are combined in one graph and shown in Fig. 5.14.

![Figure 5.14](image-url)

**Figure 5.14:** Sensing coil’s induced voltage results for the various power transfer experiments performed in section 5.3, section 5.4, and section 5.5.

The results obtained from the sensing coil verify some of the results presented
Firstly, when the CET platform is operating in single-phase mode, the sensing coil voltages stay approximately constant. This means that the stray magnetic field is mostly dominated by the magnetic fields generated by the primary windings. This conclusion is also reached in subsection 3.6.3 and presented in Fig. 3.30.

Secondly, when the CET platform is operating in its three-phase mode, the sensing coil voltages are reduced by up to 3-8 times (compared to the single-phase results). Furthermore, these voltages increase as the power transfer level and the secondary current increases. This verifies the fact that only the magnetic fields generated by the three primary winding currents contribute the destruction of the stray magnetic field. Here, the secondary winding current is still uncompensated and contributes to the stray magnetic field. These conclusions are also reached in subsection 3.6.4 and shown in Fig. 3.33.

Finally, when the CET platform is operating in variable-phase mode, the sensing coil voltages are reduced by up to 10 times (compared to the single-phase results). Moreover, these voltages stay relatively constant, compared to the results obtained from the three-phase operation. In variable-phase mode, the primary winding currents together with the secondary winding current all contribute to the destruction of the stray magnetic field. For low power transfer levels (low secondary current) the variable-phase and three-phase mode's sensing voltages give similar results. For higher power transfer levels however, the variable-phase results show lower values compared to the three-phase mode. These conclusions are also reached in subsection 3.6.5 and are clearly shown in Fig. 3.36 and Fig. 3.37.

From the various results obtained by the sensing coil, the effects of the destructive wave interference on the stray magnetic fields are indirectly shown. The measurement results closely follow and confirm the results and conclusions reached from the simulations performed in section 3.6 during the synthesis of the variable-phase CET platform.

5.7 Discussions and Conclusions

The variable-phase CET platform is developed according to the concepts in Chapter 3, and finally implemented as a hardware prototype system. Chapter 4 deals with the implementation issues of the CET platform and also discusses in detail the various design considerations taken during this process.

This chapter focusses on the experimental verification through measurements performed on the implemented CET platform prototype.

Firstly, through various experiments performed on the variable-phase CET platform, the operation of the load detection process is verified. Initially, four proof of principle measurements are performed on the platform. The results show that the presence of conductive materials, soft-magnetic materials, and valid resonant load circuits, all influence the primary windings' equivalent impedance amplitudes and phases in unique and measurable ways, as predicted in section 3.5. Afterwards, two load detection experiments are performed on the variable-phase
CET platform. The results of these two experiments show that the CET platform is not only able to correctly locate all the positions of the possible load devices placed on the CET platform, but are also able to correctly identify and distinguish all the different objects on its surface.

Secondly, various power transfer experiments are performed while the CET platform is operating in single-phase, three-phase, and variable-phase mode. During these experiments, various voltage and current measurements are taken in order to verify the results against calculated values. The secondary winding is placed at different positions above the primary winding cluster, while 0 W, 2.5 W and 5 W of power are transferred to the load. The results show that all measured secondary circuit values are accurate within 5 % - 10 %. The measurement results of the primary circuits are less accurate, with differences of up to 20 % in some cases. The differences here are mostly due to the primary circuits’ component tolerances as well as possible measurement tolerances.

Finally, in order to verify the influence of the three-phase and variable-phase modes on the stray magnetic field, the stray field is indirectly measured using a sensing coil. By measuring the induced voltage in the sensing coil, a quantitative representation of the stray magnetic field is obtained, which can then be used to compare the influence of the single-phase, three-phase, and variable-phase operation on the stray magnetic field. The sensing coil’s induced voltages are measured during the various power transfer experiments. The results confirm many of the conclusions reached about the influence of destructive wave interference on the stray magnetic fields during the three- and variable-phase operations, as presented in section 3.6. To summarize: the three-phase and variable-phase operation results show clear reductions in the sensing coil voltages, of up to 8 times and 10 times, for the two modes, respectively. This implies that the average magnetic field through the sensing coil area is also reduced. Furthermore, the three-phase test results show that the sensing coil voltage increases as the power transfer level (and secondary current) increases. As expected, the variable-phase test results show low and steady sensing voltage values, that do not change significantly with the different power transfer levels, compared to the three-phase mode.

The various experimental measurements taken from the variable-phase CET platform verifies that the system operates the way it is designed to, and that it can transfer up to 5 W of power to a load resistor connected to the secondary winding through the rectification circuit. It also experimentally verifies the different conclusions reached in Chapter 3, regarding the load detection process, and the attenuation of the stray magnetic field through destructive wave interference.
6.1 Conclusions

Contactless energy transfer, as used in this thesis, describes the process in which electrical energy is transferred among two or more galvanically isolated electrical circuits by means of magnetic induction (magnetic energy).

The potential applications for such a technology can range from the transfer of energy between low power home and office devices to high power industrial applications. Medical, marine, and other applications where physical electrical contact might be dangerous, impossible or at the very least problematic, are all prospective candidates for using CET.

Contactless energy transfer platforms attempt to remove the different cables, plugs and adaptors used to power and recharge various mobile electronic devices, by transferring power from inductor-embedded primary platforms, to the mobile devices, through the use of magnetic fields.

The variable-phase CET platform is a specialized implementation of the CET concept, and consists of multiple hexagon spiral windings arranged to form a matrix of windings. When a mobile electronic device, embedded with a similar “power receiving” winding, is placed on the platform surface, power is transferred from the platform to the mobile device through magnetic induction.

In this thesis, a CET platform using air-cored planar PCB inductors has been investigated, and this research has resulted in a fully operational and successfully tested prototype. To this end, several thesis objectives were formulated in section 1.6, which can be summarized as:
• Building up a library of analytical and numerical methods and procedures for modeling planar inductors constructed as copper tracks on printed circuit boards, so that the use of FEM software can be minimized or avoided completely.

• Developing a load detection scheme which can be used to locate mobile devices placed on top of the CET platform.

• Investigate the effectiveness of a novel magnetic field shielding methodology using destructive wave interference.

• Developing the theory, and synthesize the variable-phase CET platform based on given user requirements, in order for it to transfer the appropriate amount of power to the mobile devices. Additionally, is the development of the hardware and the implementation of the variable-phase CET platform in the form of a hardware prototype.

The purpose of this chapter is to conclude on the variable-phase CET platform system as presented in this thesis, and to discuss possible future developments in the field.

Firstly, the conclusions to the above stated thesis objectives are given in sections 6.2 - 6.5, where they are discussed in detail. Secondly, the thesis contributions are presented in section 6.6. Thirdly, a critical discussion about the variable-phase CET platform is given in section 6.7, and finally, an outlook towards future developments are given in section 6.8, also concerning CET platforms in general, and other related research activities.

6.2 Modeling Planar Inductors on Printed Circuit Boards

At the center of any CET system, lays the primary and secondary inductors which transfer power from the primary circuits to the secondary circuit and load, through their share magnetic field. In order to predict the behavior of these CET systems, the inductors they employ must be accurately modeled.

For this reason, a library of analytical methods and procedures for modeling planar PCB inductors has been developed and presented in Chapter 2. These methods can easily be implemented as algorithms in software programs like Matlab and C++, so that the use of time consuming FEM software can be minimized or avoided completely. The purpose of this “library of techniques” is not only to model the inductors used in the variable-phase CET platform presented in this thesis, but to act as a single repository for information on modeling similar planar PCB inductors for various future CET applications.

The developed models include: estimation of the DC- and AC-resistances, the mutual inductances, self-inductances, magnetic field intensities and parasitic capacitances of planar PCB inductors. Furthermore, the inductor excitation frequency limit due to the magneto-static approximation and the parasitic capacitances have also been studied and discussed.
Measurements have been performed on actual implemented PCB inductors, in order to verify the various developed inductor models. All measurement results show good agreement with calculated values. These models have been used extensively during the development of the variable-phase CET platform.

6.3 Development of a Load Detection Scheme

The variable-phase CET platform uses a matrix of hexagon spiral windings to create a large surface whereupon CET-enabled mobile devices can be placed for charging. However, in order to transfer power to these valid CET devices, the system must first detect their positions, and determine which primary windings they are closest to.

In order for the system to locate these CET-enabled devices placed on the platform, an innovative load detection scheme has been developed. The load detection scheme “scans” the primary windings while estimating their equivalent impedances through measurements. By comparing and detecting differences between the measurements and previously calibrated values, the system can find and locate the valid devices. Furthermore, this process has the added advantage of being able to distinguish between valid CET devices, metallic materials, as well as soft-magnetic materials placed on the platform surface. Objects like, soft-drink cans, lighters, pens, coins, and certain soft-magnet containing devices, which might have been accidently placed on the CET platform, can thus be detected, and the activation of the windings closest to it, avoided.

The development of the load detection scheme is presented in section 3.5. Various load detection measurements have been performed on the implemented variable-phase CET platform, and the results are presented in section 5.2. The results show correct and measurable changes in the equivalent winding impedances which allowed the system to correctly locate and distinguish valid loads, and different metallic and soft-magnetic objects.

6.4 Magnetic Field Shielding through Destructive Wave Interference

Currently, many conventional CET platforms use one or both of the two well known magnetic field shielding methodologies for shielding the internal circuitry of the CET-enabled mobile devices, from the potentially dangerous alternating magnetic fields which are used to transfer the energy. Firstly, magnetic field shielding can be accomplished through “flux-guiding”, where high permeability magnetic materials are used to “guide” the magnetic fields around and away from sensitive electronic circuits and components. Secondly, magnetic fields can be shielded through eddy current cancelation, where the eddy currents induced in conductive materials (caused by the applied alternating field) generate magnetic fields which tend to oppose the applied magnetic field.

The variable-phase CET platform system, in terms of shielding, contains no
soft-magnetic or conductive materials, and does not make use of any of the above mentioned shielding methodologies. The CET platform, however, investigates the effectiveness of a third, novel method of magnetic field shielding, using destructive wave interference.

In section 3.6, the concept of wave interference and its application for magnetic field shielding is proposed. Here it is shown, however, that within the framework of the CET platform, that magnetic field shielding through destructive wave interference only becomes effective as the distance from the windings increases. This methodology is thus useful for reducing stray magnetic fields, but has a relatively small impact on the fields close the windings. Moreover, this method does not require any additional shielding materials, but instead, uses the existing CET platform windings to create magnetic fields which counteract the fields created during power transfer.

The variable-phase CET platform operates by transferring power from a cluster of three primary windings, to a load connected to the secondary winding, and has the ability to operate in three distinct modes, namely: single-phase mode, three-phase mode, and variable-phase mode.

The first mode of operation of the variable-phase CET platform, is the single-phase operational mode, which is developed in subsection 3.6.3. During single-phase operation, all the windings inside an activated winding cluster are excited with in-phase currents. Here, no additional reduction of the magnetic field is expected. The power transfer and magnetic field results obtained from the single-phase operation are used as a baseline for comparing the magnetic field attenuations, the capabilities, as well as the efficiencies of the other operational modes.

The second mode of operation is the three-phase operational mode, which is developed in subsection 3.6.4. During three-phase operation, the windings inside an activated winding cluster are excited with currents of the same amplitude, but with 120° phase-shifted currents. Here, the phase-shifted magnetic fields will destructively add together, leading to the attenuation of the stray field.

During three-phase operation, however, when power is transferred to the secondary load device, the current in the secondary winding also produce a magnetic field, which adds to the primary fields. This additional secondary magnetic field, however, is not taken into account by the three-phase mode. The third mode of operation of the CET platform is the variable-phase operational mode, which is developed in subsection 3.6.5. The variable-phase mode is an extension of the three-phase mode, and attempts to further reduce the stray magnetic field, by taking into account the secondary windings’ magnetic field. Here, the primary windings are excited with precalculated currents of different amplitudes and phases. These values, of which some are presented in the same section, are calculated to give minimum peak stray magnetic field values for certain secondary winding placements and power transfer levels. In this way, all the primary windings as well as the secondary winding, contribute to the reduction of the stray magnetic field.

Various simulations have been performed on the CET platform in order to
6.5. Synthesis and Implementation of the Variable-Phase CET Platform

determined the effectiveness of the magnetic field attenuation through destructive wave interference. The simulation results have been presented in subsection 3.6.6, and show that both the three-phase and variable-phase modes are able to reduce the stray magnetic fields (for the same amount of power transferred to the load resistor) compared to the single-phase mode. During low power transfer, the three-phase and variable-phase modes show similar results. During higher power transfer levels, however, the variable-phase mode results show better attenuation of the stray field, compared to the three-phase operations. This stems from the fact that the variable-phase mode takes into account the magnetic field created by the secondary winding, whereas the three-phase mode does not. Stray field attenuation values of up to 12 dB are estimated.

The effects of the destructive wave interference on the stray magnetic field have also been verified through measurements, of which the results are presented in section 5.6. Here, the stray magnetic field is indirectly observed by measuring the voltage induced in a circular sensing coil placed above the primary winding cluster. The measurement results closely follow and confirm the results and conclusions reached from the simulations performed during the synthesis of the CET platform.

6.5 Synthesis and Implementation of the Variable-Phase CET Platform

The synthesis and implementation of the variable-phase CET platform is performed in order to get from the basic concept, to a fully working platform prototype.

To begin with, a synthesis of the variable-phase CET platform has been performed and is presented in Chapter 3. Here, firstly, various inductive power transfer details are discussed in section 3.2. It is shown that capacitive compensated series resonant primary and secondary circuits and a current controlled primary voltage source can be used to removed inherent disadvantages of the power transfer process between weakly coupled inductors, increasing the power transfer capability and efficiency. The controllability of the power transfer process is also increased in this way. Secondly, the geometry of the CET inductor platform is discussed in section 3.3. Here, it is shown that fully turned planar spiral windings shaped as hexagons are best suited for the platform, and that the unique arrangement of windings in the “honey-comb”-like winding lattice, strongly supports a three-phase operational methodology. Thirdly, a synthesis of the variable-phase CET platform is performed in section 3.4. Here, the user specifications and constraints, together with the planar windings models as presented in Chapter 2, and the power transfer equations are used to simulate and create the CET platform. Through a parametric search process, the set of winding-geometric and lumped parameters as well as the electrical parameters which will allow successful power transfer between a primary and secondary winding are calculated. The final platform geometry is chosen, and consists of primary windings with 12.5 mm radii, secondary windings of 20 mm radii, and an operating
frequency of 2.777 MHz. Finally, the load detection process is presented in section 3.5, and the stray field attenuation through wave interference is presented in section 3.6.

Next, the implementation of the variable-phase CET platform has been performed and is presented in Chapter 4. This chapter is dedicated to the implementation of the system, and to discuss in detail the various design considerations taken in developing it. Firstly, an overview of the complete concept is presented in section 4.2. The CET platform is designed in a modular fashion with various subsystems, each with a specific role to perform. The prototype system consists of three CET hardware channels, which allows it to simultaneously excite and control three primary windings at a time. Secondly, the primary-side subsystems are discussed in section 4.3. These systems and circuits make up the variable-phase CET platform and include the primary winding matrix, the primary controller, the quadratic buck converters, the high-frequency MOSFET half-bridge drivers, the half-bridge inverters, the half-bridge output voltage amplitude and phase estimators, the winding current amplitude and phase estimators, as well as the winding commutation circuits. Thirdly, the secondary-side subsystems are discussed in section 4.4. These systems and circuits receive power from the primary winding matrix and convert the AC-voltage signal induced in the secondary winding, into more useful DC-voltage levels for the mobile devices to use. These systems include the secondary winding, the full-bridge rectifier, the DC-to-DC converter, and the load. The primary- and secondary-side subsystems are all implemented and separately tested. Their operations are verified through various measurements, in order to ensure that they work as expected, within the larger system.

Finally, the primary current amplitude and phase controllers are developed and discussed in section 4.5. These compensators are used to control the primary winding current amplitudes and phases, in order for the variable-phase CET platform to operate in its different modes. The primary current amplitude controllers are implemented as PI compensators. They are tested, and are shown to be able to maintain relatively constant primary winding current amplitudes for loaded and uploaded situations. Furthermore, the current phase compensators are implemented as integrator controllers. In this way, the controllers avoid step-like responses produced by proportional control elements, and reacts in a smooth and slow-acting fashion, to large phase errors. This gives the current amplitude controllers enough time to maintain steady primary current amplitudes, while the current phases are adjusted to their required values.

The operation of the variable-phase CET platform has been verified through various measurements performed on the prototype while it is operating in its three modes. The measurement results are presented in section 5.3, section 5.4, and section 5.5. It is shown that the prototype can successfully transfer up to 5 W of power (DC) to a load, in all three operational modes. All the measured and calculated results show good agreement.
6.6 Thesis Contributions

The major scientific and technological contributions of this thesis can be summarized as follows:

- A library of analytical and numerical methods and procedures for modeling planar inductors constructed as copper tracks on printed circuit boards.

- New load detection scheme which allows the CET platform to find the positions of CET-enabled load devices placed on top of the winding matrix.

- A novel stray magnetic field shielding methodology, which uses destructive wave interference to attenuate the stray magnetic field generated by the CET platform.

- The theoretical background, the synthesis, and the implementation of a working CET platform prototype, of which the knowledge and designs presented in this thesis can be used by future developers in other CET-related projects.

6.7 Discussion: The Variable-Phase CET Platform

The variable-phase CET platform has been implemented, and through various experiments it is shown that the prototype can transfer up to 5 W of power (DC) to a load, when operating in any of its three modes.

In discussion, firstly, the original idea of the CET platform, as stated in Chapter 1, was for the prototype to transfer power to a mobile electronic device placed on the winding platform surface. To this end, the hardware was developed and presented in Chapter 4. In Chapter 5, however, the experimental measurements went only as far as transferring power to a resistor connected to the secondary winding through the full-bridge rectifier. Power transfer to a mobile device was not fully tested. The reason for this is due to the occurrence of stability problems with the current amplitude controllers when the LM2953-based DC-to-DC converter is connected to the secondary circuit, which causes oscillations of the primary winding currents. However, the transfer of power to a mobile device is an implementation problem, which can be solved in time, by redesigning the secondary circuit and perhaps choosing a different DC-to-DC converter. By transferring power to a load resistor connected to the full-bridge rectifier instead, the CET power transfer principle is still proven, albeit, not in the originally intended way.

Secondly, the variable-phase CET platform essentially reduces the stray magnetic field produced during power transfer, by generating more magnetic fields which interfere with the already existing fields in a destructive manner. Shielding is thus attained, but at the expense of increased power loss, and reduced efficiency. In today’s society, people are becoming more aware of the potential health risks regarding the use of devices generating electromagnetic fields,
such as cell phones in particular. On the other hand, with the fears of possible global warming, people are also becoming conscious about the negative impacts of unnecessarily wasting energy. The variable-phase CET platform thus raises an interesting philosophical question: which is better, an efficient device with larger stray magnetic fields, or an safer inefficient system with smaller stray fields?

Thirdly, other important questions can be raised: How does the shielding efficiency of the variable-phase CET platform compare against the other existing methods? How dangerous are these fields to humans, and the internal electronics of the CET-enabled mobile devices? How effective are the other existing methods for shielding these devices? These are all very important questions, which will need to be addressed before the CET systems become commercially available.

Finally, it should however, be reiterated that the variable-phase CET platform as presented is not intended for direct commercial use, but rather, to server as a research platform which is used to investigate some novel CET-related concepts, which could possibly be used to improve future CET charging platforms.

6.8 Outlook Towards Future Developments

Based on literature studies, as well as information gathered during the development of the variable-phase CET platform, the following projections for the future developments are done:

6.8.1 Standardization

Recently, a handful of innovative companies have started researching, producing, and selling CET charging platforms for charging different mobile devices through magnetic induction. Although these systems use very similar technology, they are most probably not cross compatible. Due to the lack of standardization in this field, no clear set of operational rules exist to ensure cross compatibility and conformation between the systems designed by different companies. Although more and more CET devices will appear, large scale deployment will probably not occur until the technology is standardized and backed up by either a large respectful professional organization, like the IEEE or ISO, or a consortium of large electronics manufacturing companies.

6.8.2 CET and Communications

RFID systems operate on the same principles than CET systems, albeit, on lower power levels. Here, power is transferred from primary RFID readers to the receiver coils inside the RFID tags. Passive RFID tags contain no batteries and use inductive power transfer to power the internal circuity of the tags. The RFID tags, in turn, communicate with the RFID readers, by modulating data onto their secondary coil currents, which is then reflected back to the RFID readers’ primary windings. This technology is called “backscatter modulation” and can also be used in CET devices to facilitate communications between the CET platform and the valid CET mobile device. This inductive communications
channel can be used by the mobile device to communicate its power demands to the CET platform, and can even be used to create a local area network, where data can be transferred between the various CET-enabled devices placed on the platform. Furthermore, a communication system such as this can also be used to create another, smarter load detection scheme, where valid devices can be located via their transmitted signals.

6.8.3 CET Platforms for Automotive Industry

With the depletion of fossil fuels, and the increasing oil prices, new and interesting technologies are emerging in the automotive industry. One of these technologies are electric vehicles. Electric vehicles like electric cars, use electric motors for propulsion, powered by electrical energy stored in batteries, in opposed to standard cars using internal combustion engines powered by hydrocarbon fuels. Since these electric cars use batteries which can only store a finite amount of energy, they need to be recharged periodically. Here, instead of using power cables and plugs, CET can be used to recharge electric car batteries through inductive means. By embedding large power coils inside a garage or motor parking bay floor, power can be transferred from the embedded CET platform to the car batteries without the use of cables and plugs.
A.1 Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Delta_l$</td>
<td>numerical integration element on primary filament contour</td>
<td>-</td>
</tr>
<tr>
<td>$\Delta_k$</td>
<td>numerical integration element on secondary filament contour</td>
<td>-</td>
</tr>
<tr>
<td>$\epsilon_0$</td>
<td>permittivity of vacuum</td>
<td>F/m</td>
</tr>
<tr>
<td>$\epsilon_r$</td>
<td>relative permittivity</td>
<td>-</td>
</tr>
<tr>
<td>$\varepsilon$</td>
<td>error</td>
<td>-</td>
</tr>
<tr>
<td>$\eta$</td>
<td>efficiency</td>
<td>-</td>
</tr>
<tr>
<td>$\eta_{CET}$</td>
<td>inductive power transfer efficiency</td>
<td>%</td>
</tr>
<tr>
<td>$\theta$</td>
<td>current phase angle</td>
<td>deg</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>flux linkage</td>
<td>Wb</td>
</tr>
<tr>
<td>$\lambda_P$</td>
<td>amount of primary winding layers</td>
<td>-</td>
</tr>
<tr>
<td>$\lambda_S$</td>
<td>amount of secondary winding layers</td>
<td>-</td>
</tr>
<tr>
<td>$\mu_0$</td>
<td>permeability of vacuum</td>
<td>H/m</td>
</tr>
<tr>
<td>$\mu_r$</td>
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</tr>
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<td>$\rho$</td>
<td>charge density</td>
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</tr>
<tr>
<td>$\rho_r$</td>
<td>resistivity</td>
<td>$\Omega$m</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>conductivity</td>
<td>$(\Omega m)^{-1}$</td>
</tr>
<tr>
<td>$\Upsilon_P$</td>
<td>primary winding radius</td>
<td>m</td>
</tr>
<tr>
<td>$\Upsilon_S$</td>
<td>secondary winding radius</td>
<td>m</td>
</tr>
<tr>
<td>$\omega$</td>
<td>radial frequency</td>
<td>rad/s</td>
</tr>
<tr>
<td>$\varphi$</td>
<td>electric potential</td>
<td>V</td>
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### List of Symbols and Abbreviations

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<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>Unit</th>
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</thead>
<tbody>
<tr>
<td>(a)</td>
<td>total amount of straight line segments</td>
<td>-</td>
</tr>
<tr>
<td>(a^k)</td>
<td>total amount of straight line segments on secondary filament contour</td>
<td>-</td>
</tr>
<tr>
<td>(A)</td>
<td>magnetic vector potential</td>
<td>Wb/m</td>
</tr>
<tr>
<td>(A)</td>
<td>cross-sectional area of conductor</td>
<td>m²</td>
</tr>
<tr>
<td>(A_z)</td>
<td>magnetic vector potential (z)-component</td>
<td>Wb/m</td>
</tr>
<tr>
<td>(b^l)</td>
<td>amount of vertices on primary filament contour</td>
<td>-</td>
</tr>
<tr>
<td>(b^k)</td>
<td>amount of vertices on secondary filament contour</td>
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</tr>
<tr>
<td>(B)</td>
<td>magnetic field (magnetic flux density)</td>
<td>T</td>
</tr>
<tr>
<td>(B_A)</td>
<td>absolute magnetic field</td>
<td>T</td>
</tr>
<tr>
<td>(B_A^{close})</td>
<td>absolute magnetic field between the primary and secondary windings</td>
<td>T</td>
</tr>
<tr>
<td>(B_{A^{mid}})</td>
<td>absolute magnetic field at 10 mm above the secondary winding</td>
<td>T</td>
</tr>
<tr>
<td>(B_{A^{stray}})</td>
<td>absolute stray magnetic field at 100 mm above the secondary winding</td>
<td>T</td>
</tr>
<tr>
<td>(c)</td>
<td>speed of light</td>
<td>m/s</td>
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<tr>
<td>(C)</td>
<td>capacitance</td>
<td>F</td>
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<td>(C_{air})</td>
<td>partial capacitances between two tracks’ cross-sections in the absence of a substrate</td>
<td>F/m</td>
</tr>
<tr>
<td>(C_{iss})</td>
<td>MOSFET input capacitance</td>
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</tr>
<tr>
<td>(c^l)</td>
<td>Euclidean lengths of primary contour filaments</td>
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</tr>
<tr>
<td>(c^k)</td>
<td>Euclidean lengths of secondary contour filaments</td>
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</tr>
<tr>
<td>(C_{P}^{par})</td>
<td>parasitic capacitance of primary winding</td>
<td>F</td>
</tr>
<tr>
<td>(C_{P}^{pr})</td>
<td>primary parallel resonant capacitor</td>
<td>F</td>
</tr>
<tr>
<td>(C_{P}^{sr})</td>
<td>primary series resonant capacitor</td>
<td>F</td>
</tr>
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<td>(C_{sub})</td>
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<td>F/m</td>
</tr>
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<td>(C_{S}^{par})</td>
<td>parasitic capacitance of secondary winding</td>
<td>F</td>
</tr>
<tr>
<td>(C_{S}^{pr})</td>
<td>secondary parallel resonant capacitor</td>
<td>F</td>
</tr>
<tr>
<td>(C_{S}^{sr})</td>
<td>secondary series resonant capacitor</td>
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</tr>
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<td>(C_T)</td>
<td>total inter-track capacitance</td>
<td>F/m</td>
</tr>
<tr>
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<td>duty cycle</td>
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</tr>
<tr>
<td>(d)</td>
<td>winding track thickness</td>
<td>m</td>
</tr>
<tr>
<td>(d_P)</td>
<td>primary winding track thickness</td>
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</tr>
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</tr>
<tr>
<td>(D)</td>
<td>diode</td>
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</tr>
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<td>(D)</td>
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<tr>
<td>(D)</td>
<td>electric displacement field</td>
<td>C/m²</td>
</tr>
<tr>
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<td>m</td>
</tr>
<tr>
<td>(D_O)</td>
<td>outer width of a PCB inductor</td>
<td>m</td>
</tr>
<tr>
<td>(e)</td>
<td>position vector</td>
<td>m</td>
</tr>
<tr>
<td>(E)</td>
<td>electric field strength</td>
<td>V/m</td>
</tr>
<tr>
<td>(f)</td>
<td>frequency</td>
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<tr>
<td>(f)</td>
<td>position vector</td>
<td>m</td>
</tr>
<tr>
<td>(f_{BUCK})</td>
<td>buck converter PWM frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>(f_{CET})</td>
<td>CET operating frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>(f_{FPGA})</td>
<td>FPGA controller frequency</td>
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### A.1. Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>Unit</th>
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<tr>
<td>$f_p^m$</td>
<td>primary winding characteristic frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>$f_p^{SR}$</td>
<td>primary winding self-resonating frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>$f_s^m$</td>
<td>secondary winding characteristic frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>$f_s^{SR}$</td>
<td>secondary winding self-resonating frequency</td>
<td>Hz</td>
</tr>
<tr>
<td>$\mathbf{g}$</td>
<td>position vector</td>
<td>m</td>
</tr>
<tr>
<td>$h$</td>
<td>distance in $z$-direction</td>
<td>m</td>
</tr>
<tr>
<td>$h_{PS}$</td>
<td>distance between primary and secondary winding in $z$-direction</td>
<td>m</td>
</tr>
<tr>
<td>$H$</td>
<td>magnetic field intensity</td>
<td>A/m</td>
</tr>
<tr>
<td>$H_A$</td>
<td>absolute magnetic field intensity</td>
<td>A/m</td>
</tr>
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<td>$h_I$</td>
<td>inner height of PCB inductor</td>
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<td>$h_O$</td>
<td>outer height of PCB inductor</td>
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<td>$I$</td>
<td>current</td>
<td>A</td>
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<td>$I_{I, II, III}$</td>
<td>CET platform prototype hardware channels</td>
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<td>$I_D$</td>
<td>MOSFET drain current</td>
<td>A</td>
</tr>
<tr>
<td>$I_L$</td>
<td>load current</td>
<td>A</td>
</tr>
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<td>$I_P$</td>
<td>primary winding current</td>
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</tr>
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<td>estimated primary winding current</td>
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</tr>
<tr>
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<td>primary winding current reference</td>
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<td>$J$</td>
<td>current density</td>
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<td>Bessel function of the first kind</td>
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<tr>
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<td>current density $z$-component</td>
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<td>$k$</td>
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<td>$K(\cdot)$</td>
<td>elliptic integral of the first kind</td>
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<td>$K_I$</td>
<td>integral gain of PI controller</td>
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<td>$K_P$</td>
<td>proportional gain of PI controller</td>
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<td>secondary winding inductance</td>
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</tr>
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<td>mutual inductance</td>
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<td>primary-to-secondary winding mutual inductance</td>
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<td>normalized position vectors of the primary filamentary structure</td>
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</tr>
<tr>
<td>$n^b$</td>
<td>normalized position vectors of the secondary filamentary structure</td>
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<td>$P$</td>
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<td>$P_{PR}$</td>
<td>power loss in primary winding resistance</td>
<td>W</td>
</tr>
<tr>
<td>$P_{PS}$</td>
<td>power loss in secondary winding resistance</td>
<td>W</td>
</tr>
<tr>
<td>$P_L$</td>
<td>power dissipated in the load</td>
<td>W</td>
</tr>
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<td>$P_1, \ldots, P_s$</td>
<td>secondary winding positions</td>
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<td>Symbol</td>
<td>Description</td>
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<td>--------</td>
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<tr>
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<td>coordinates of primary winding’s contour vertices</td>
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</tr>
<tr>
<td>$q^k$</td>
<td>coordinates of secondary winding’s contour vertices</td>
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<td>resistance</td>
<td>Ω</td>
</tr>
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<td>$R_{DS(on)}$</td>
<td>drain-source resistance of MOSFET in conduction state</td>
<td>Ω</td>
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<tr>
<td>$R_I$</td>
<td>inner radius of a PCB inductor</td>
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<tr>
<td>$R_L$</td>
<td>load resistance</td>
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<td>$R_O$</td>
<td>outer radius of a PCB inductor</td>
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<td>$R_P$</td>
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<td>$R_{PAC}$</td>
<td>AC-resistance of primary winding</td>
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</tr>
<tr>
<td>$R_{PDC}$</td>
<td>DC-resistance of primary winding</td>
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<td>$R_S$</td>
<td>secondary winding resistance</td>
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<tr>
<td>$R_{SAC}$</td>
<td>AC-resistance of secondary winding</td>
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</tr>
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<td>$R_{SDC}$</td>
<td>DC-resistance of secondary winding</td>
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<td>secondary winding inter-track spacing</td>
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<td>$V$</td>
<td>voltage</td>
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<td>$V_{DS}$</td>
<td>MOSFET drain-source voltage</td>
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<tr>
<td>$V_{GS}$</td>
<td>MOSFET gate-source voltage</td>
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<td>$V_{GS(th)}$</td>
<td>MOSFET gate-source threshold voltage</td>
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<td>$V_L$</td>
<td>load voltage</td>
<td>V</td>
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<td>$V_P$</td>
<td>primary winding voltage</td>
<td>V</td>
</tr>
<tr>
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<td>secondary winding voltage</td>
<td>V</td>
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</tr>
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<td>$x, y, z$</td>
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</tr>
<tr>
<td>$x', y', z'$</td>
<td>cartesian coordinate</td>
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<td>$Z_{P}^{SCAN}$</td>
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A.2 Abbreviations

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<tr>
<th>Abbreviation</th>
<th>Definition</th>
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<td>2D</td>
<td>two-dimensional</td>
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<tr>
<td>3D</td>
<td>three-dimensional</td>
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<tr>
<td>AC</td>
<td>alternating current</td>
</tr>
<tr>
<td>ADC</td>
<td>analog to digital converter</td>
</tr>
<tr>
<td>C++</td>
<td>“C Plus Plus” (a programming language)</td>
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<tr>
<td>CET</td>
<td>contactless energy transfer</td>
</tr>
<tr>
<td>CIET</td>
<td>contactless inductive energy transfer</td>
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<tr>
<td>CISPR</td>
<td>Comité International Spécial des Perturbations Radioélectriques</td>
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<td>CPT</td>
<td>contactless power transfer</td>
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<tr>
<td>CPU</td>
<td>central processing unit</td>
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<td>DC</td>
<td>direct current</td>
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<td>DCM</td>
<td>distributed capacitance model</td>
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<td>digital signal processing</td>
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<td>EMF</td>
<td>electromotive force</td>
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<td>FCC</td>
<td>Federal Communications Commission</td>
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<td>FEM</td>
<td>finite element method</td>
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<td>FPGA</td>
<td>field-programmable gate array</td>
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<td>FR4</td>
<td>flame retardant 4</td>
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<td>I/O</td>
<td>input-output</td>
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<td>inductive energy transfer</td>
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<tr>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
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<td>IPT</td>
<td>inductive power transfer</td>
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<tr>
<td>ISO</td>
<td>International Organization for Standardization</td>
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<td>JTAG</td>
<td>joint test access group</td>
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<td>LCR</td>
<td>inductance, capacitance, resistance</td>
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<td>LED</td>
<td>light-emitting diode</td>
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<td>MOSFET</td>
<td>metal-oxide-semiconductor field-effect transistor</td>
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<td>PC</td>
<td>personal computer</td>
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<td>PCB</td>
<td>printed circuit board</td>
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<td>PDA</td>
<td>personal digital assistant</td>
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<td>PI</td>
<td>proportional-integral</td>
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<td>PWM</td>
<td>pulse width modulation</td>
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<td>RFID</td>
<td>radio-frequency identification</td>
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<td>RMS</td>
<td>root mean square</td>
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<td>RS-232</td>
<td>recommended standard 232</td>
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<td>UART</td>
<td>universal asynchronous receiver/transmitter</td>
</tr>
<tr>
<td>USA</td>
<td>United States of America</td>
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<td>USB</td>
<td>universal serial bus</td>
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<td>VHDL</td>
<td>VHSIC hardware description language</td>
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<td>VHSIC</td>
<td>very high speed integrated circuits</td>
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<td>WET</td>
<td>wireless energy transfer</td>
</tr>
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<td>WPT</td>
<td>wireless power transfer</td>
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Bibliography


[75] Phillips Polystyrene Film Capacitors. Not produced or sold anymore.


Samenvatting

Draadloos Energieoverdrachtplatform met Vlakke Luchtspoelen en Variabele Faseaansturing

Draadloze Energieoverdracht (DEO) is het proces waarbij elektrische energie bidirectioneel uitgewisseld kan worden tussen verschillende elektrische netwerken of apparaten die galvanisch gescheiden zijn. De uitwisseling van energie vindt plaats via wederzijdse magnetische inductie. Er is een breed scala van toepassingen denkbaar waarin een dergelijke technologie kan worden toegepast, variërend van kleine huishoudelijke- of kantoorapparatuur, tot grote industriële applicaties met grote vermogens. Ook wanneer een galvanisch contact een direct gevaar vormt voor de mens (medische apparatuur) of waarbij galvanisch contact problematisch of gewoonweg onmogelijk is (maritieme toepassingen), kunnen DEO-systemen een uitkomst bieden.

In dit proefschrift wordt de nadruk gelegd op de fundamentele concepten van DEO-systemen bestaand uit een oppervlak met spoelen die gemaakt zijn met conventionele printplaattechnologie. Indien er een elektrisch apparaat op het platform wordt gelegd kan er energie overgedragen worden van het platform naar het apparaat. Het apparaat is voorzien van een “ontvangende” spoel, vergelijkbaar met de spoelen in het platform. Enkele spoelen in het platform en de ontvangende spoel zijn nu via het magnetische veld inductief gekoppeld. Voor de ontvangende spoel wordt ook een platte spoel op een printplaat gebruikt. Dergelijke spoelen zijn goedkoop te fabriceren, licht en makkelijk aan te brengen in de bestaande apparatuur. Op deze wijze wordt het gebruik van kabels, stekkers en adapters vermeden.

Er wordt een overzicht gegeven van analytische en numerieke magnetostatische methoden om de vlakke printplaatsspoelen te kunnen modelleren. Tevens worden er modellen afgeleid waarmee o.a. de DC- en AC-weerstand, wederzijdse inductie, zelfinductie, magnetische veldsterkte en parasitaire capaciteiten bepaald kunnen worden. Met behulp van deze modellen wordt vervolgens de maximale excitatiefrequentie van de spoelen berekend in relatie tot de resonantiefrequenties. Het overzicht van de modellen heeft niet alleen als doel het modelleren van de in dit proefschrift gebruikte printplaatスポール, maar kan ook als vademecum fungeren voor de ontwikkeling van vergelijkbare toekomstige DEO-toepassingen. De geldigheid van de modellen is getoetst aan de hand van eindige elementen simulaties en metingen.

Daarna wordt er ingegaan op de ontwikkeling van de theorie en de synthese van het DEO-systeem. Uitgaande van de gebruikersspecificaties is het DEO-systeem zodanig ontworpen, dat een groep van drie primaire spoelen in staat is om 8 W over te dragen naar de belaste secundaire spoel over een afstand van 5 mm. Na het gelijkrichten en verlagen van de spanning aan de secundaire kant, is er nog 5 W aan vermogen beschikbaar voor een applicatie. Het primaire spoelenplatform is opgebouwd uit een matrix van zeshoekige printplaatsspoelen, bestaand uit spiraalvormige windingen met een buiten-
straal van 12,5 mm. De spiraalvormige secundaire spoel is ook zeshoekig en heeft een buitenstraal van 20 mm. De excitatiefrequentie van het systeem is 2,777 MHz. Het systeem is ontworpen om tenminste 10 V geïnduceerde RMS spanning te leveren aan de uitgang van de secundaire spoel. Na omzetting, is een DC uitgangsspanning van tenminste 5 - 6 V gegarandeerd. Het DEO-systeem is geïmplementeerd als prototype. De hoogfrequentie vermogenselectronica is modulair van opzet en bevat meerdere subsystemen, waaronder een FPGA-regeling, quadratic buck omzetters, hoogfrequente halve bruggen, spanning- en stroomschatters en een commutatiecircuit voor het schakelen van de spoelen.

Een innovatieve belastingdetectieprocedure is ontwerpen en gerealiseerd die kan achterhalen wat de locatie van een apparaat op het spoelenplatform is. Tijdens de belastingdetectieprocedure worden de primaire spoelen gescand en de equivalente impedantie gemeten. Door verschillen in de metingen en de eerder gekalibreerde impedantiewaarden te detecteren, kan de locatie van een geldig apparaat bepaald worden. Bovendien kan de belastingdetectie een onderscheid maken tussen een geldig DEO-apparaat en ongewenste metalen en ferromagnetische materialen. Voorwerpen zoals blikjes, aanstekers, pennen, en munten die abusievelijk op het platform geplaatst zijn, kunnen dus worden gedetecteerd. Zo kan voorkomen worden dat spoelen in de buurt van deze voorwerpen geactiveerd worden.

Een nieuwe techniek wordt toegepast voor het afschermen van het magnetische strooiveld, die gebaseerd is op destructieve interferentie. In tegenstelling tot conventionele afschermingsmethoden wordt hierbij geen gebruik gemaakt van ferromagnetische of geleidende materialen. De fases en amplitudes van de stromen door de primaire spoelen worden aangepast om het magnetische veld boven de spoelen te beïnvloeden. Op deze wijze kan het strooiveld verminderd worden. Het systeem kan in drie modi opereren: éénfasig, driefasig en met een variabele faseaansturing. In de éénfasige modus worden alle primaire spoelen bekrachtigd met stromen die allemaal in fase zijn en een gelijke amplitude hebben. In dit geval zal er geen verlaging in de magnetische veldsterkte plaatsvinden. Indien het systeem zich in de driefasige modus bevindt, worden de spoelen bekrachtigd met stromen met gelijke amplitudes, maar met een onderling faseverschil van 120°. Het magnetische veld boven de spoelen zal nu onderhevig zijn aan destructieve interferentie. Hierdoor zal het strooiveld afnemen. Tijdens de energieoverdracht zorgt de stroom in de secundaire spoel toch voor een toename van het strooiveld. In de variabele faseaansturingmodus wordt getracht het strooiveld verder te reduceren door de stroom, die geïnduceerd wordt in de secundaire spoel gedurende de energieoverdracht, te verdisconteren. De spoelen binnen een geactiveerde groep van drie primaire spoelen worden nu bekrachtigd met stromen met vooraf berekende amplitudes en onderlinge faseverschuivingen, die het minimale strooiveld opleveren afhankelijk van de positie van de secundaire spoel en het vermogensoverdrachtsniveau. Simulaties hebben aangetoond dat de driefasige en variabele faseaansturingsmodus het strooiveld verzwakken met maximaal 12 dB ten opzichte van de enkelfase modus voor een gelijkblijvende belasting.

De vermogensoverdrachtscapaciteit, de belastingdetectieprocedure en de afscherming van het strooiveld via destructieve interferentie zijn geverifieerd met de simulaties en door metingen aan het prototype. Het prototype is in staat om een vermogen van 5 W draadloos over te dragen aan een apparaat dat aangesloten is op het secundaire circuit, opererend in de drie verschillende modi. De metingen komen goed overeen met de simulaties.
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Curriculum Vitae

Christoph L. W. Sonntag was born on 13th March 1979, in Bellville, South Africa. He attended secondary school at Randburg High School, Johannesburg. In 2002 he received his B.Eng degree in Electronic and Electrical Engineering from the University of Stellenbosch, South Africa. In 2005 he received his M.Sc.Eng degree at the same university. From 2002 until 2004 he worked as a development engineer with a South African electronics research and manufacturing company in Sandton, South Africa. In September 2005 he started his research in the field of low power contactless energy transfer systems in the EPE-group at the Eindhoven University of Technology, Netherlands, and started working towards his Ph.D degree, which resulted in this thesis.