DESIGN OF INDUCTORS FOR POWER CONVERTERS OPERATING AT INTERMEDIATE SWITCHING FREQUENCIES

A thesis submitted in fulfilment of the requirements for the degree of Doctor of Philosophy

by

Samira Janghorban

M. E. (2009) RMIT University
B.Sc. (2004) IAUN University

School of Engineering
College of Science, Engineering and Health
RMIT University Australia
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Samira Janghorban
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________________________________________

Samira Janghorban
ABSTRACT

Magnetic components (i.e. inductors and transformers) are essential elements of modern power converters. However, the design of magnetic components is a complex and iterative process, requiring trade-offs between a large number of parameters and careful consideration of their various interactions. To date, the research that has been conducted in this area has mainly targeted either low or high frequency converter regions of operation, focusing on issues such as material selection, loss modelling and high frequency effects. Consequently, there remains a gap in the body of knowledge regarding optimal design and construction of higher-power (i.e. > 1 kW) magnetic components for converters operating in the intermediate switching frequency range (i.e. 1 kHz-25 kHz).

The present-day design methodologies for magnetic components typically provide guidelines for a user to design magnetic components using a basic set of fundamental rules. However, intermediate-frequency magnetic components are more difficult to design because of constraints in the selection of suitable core materials, conductor types and problems of dealing with non-sinusoidal excitation waveform. Strategies which suit other frequency ranges are often used based on a series of assumptions that date back over two decades, many of which are valid only in a lower power higher frequency context. Consequently, direct usage of these techniques for high power intermediate frequency applications can require a significant number of design iterations and even then can result in either an unconstructable design or a poor performance solution.

This thesis develops an improved methodology for the design of higher power inductors operating at the intermediate frequency range. It first creates a multivariable optimising type system using an expert system approach that addresses the complexity of the design inter-relationships by iterating and trading-off objectives to achieve a design answer. This stage of the work focuses on the development of a knowledge-based advisory system for design of magnetic components.
Abstract

The second stage is to find the limitations of present design methodologies; it examines why current state-of-the-art design methodologies are not directly applicable to this frequency range by revisiting design principles. The thesis then explores the development of an improved user friendly methodology to suit the development of physically constructible designs for power inductors for converters operating at the intermediate frequency range. The developed strategy uses a 3D graph-based error minimization approach which automatically sweeps across the key design parameters until it converges on the best possible solution. Then it introduces an evaluative comparison between simulation results and experimental implementation in a prototype converter running under full load conditions by performance evaluation of the technique in terms of loss optimisation, temperature rise and the overall dimension of the final design.

A significant part of the contributions presented in this thesis have been published in peer reviewed papers, and are identified accordingly as appropriate.
ACKNOWLEDGEMENTS

First and foremost, I would like to thank my supervisors Prof. Grahame Holmes and Prof. Xinghuo Yu for giving me the opportunity to do my PhD under their supervision and their guidance and inspiration. Very special thanks to Grahame for his invaluable ideas and advice during this period. It has been a privilege to work with him. I also would like to express my sincere gratitude to Dr. Brendan McGrath for his precious feedback and comments.

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Several parts of the work and ideas presented in this thesis have been published by the author during the course of the research. These publications are listed below.


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<th>Description</th>
</tr>
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<tbody>
<tr>
<td>AC</td>
<td>Alternative Current</td>
</tr>
<tr>
<td>ADC</td>
<td>Analogue-to-Digital Converter</td>
</tr>
<tr>
<td>ANN</td>
<td>Artificial Neural Network</td>
</tr>
<tr>
<td>APDL</td>
<td>ANSYS Parametric Design Language</td>
</tr>
<tr>
<td>BB</td>
<td>Branch and Bound</td>
</tr>
<tr>
<td>BP</td>
<td>Back Propagation</td>
</tr>
<tr>
<td>CAE</td>
<td>Computer Aided Engineering</td>
</tr>
<tr>
<td>CCLR</td>
<td>Core to DC Copper Loss Ratio</td>
</tr>
<tr>
<td>CPU</td>
<td>Central Processing Unit</td>
</tr>
<tr>
<td>CPT</td>
<td>Creative Power Technologies</td>
</tr>
<tr>
<td>CSI</td>
<td>Current Source Inverter</td>
</tr>
<tr>
<td>DAB</td>
<td>Dual Active Bridge</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processor</td>
</tr>
<tr>
<td>DT</td>
<td>Decision Tree</td>
</tr>
<tr>
<td>EMI</td>
<td>Electro Magnetic Interference</td>
</tr>
<tr>
<td>ES</td>
<td>Expert System</td>
</tr>
<tr>
<td>EV</td>
<td>Electric Vehicles</td>
</tr>
<tr>
<td>FEA</td>
<td>Finite Element Analysis</td>
</tr>
<tr>
<td>FEM</td>
<td>Finite-Element Method</td>
</tr>
<tr>
<td>GA</td>
<td>Genetic Algorithm</td>
</tr>
<tr>
<td>GP</td>
<td>Geometric Programming</td>
</tr>
<tr>
<td>GSE</td>
<td>Generalized Steinmentz Equation</td>
</tr>
<tr>
<td>GUI</td>
<td>Graphical User Interface</td>
</tr>
<tr>
<td>HF</td>
<td>High Frequency</td>
</tr>
<tr>
<td>HV</td>
<td>High Voltage</td>
</tr>
<tr>
<td>Acronym</td>
<td>Definition</td>
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<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
</tr>
<tr>
<td>iGSE</td>
<td>Improved GSE</td>
</tr>
<tr>
<td>i²GSE</td>
<td>Improved iGSE</td>
</tr>
<tr>
<td>KBE</td>
<td>Knowledge-Based Engineering</td>
</tr>
<tr>
<td>KBS</td>
<td>Knowledge-Based Systems</td>
</tr>
<tr>
<td>L</td>
<td>Inductive (filter)</td>
</tr>
<tr>
<td>LF</td>
<td>Low Frequency</td>
</tr>
<tr>
<td>LV</td>
<td>Low Voltage</td>
</tr>
<tr>
<td>MATLAB</td>
<td>MATLAB mathematical package</td>
</tr>
<tr>
<td>MINP</td>
<td>Mixed Integer Nonlinear Programming</td>
</tr>
<tr>
<td>MSE</td>
<td>Modified Steinmetz Equation</td>
</tr>
<tr>
<td>NLP</td>
<td>Non-Linear Programming</td>
</tr>
<tr>
<td>NSE</td>
<td>Natural Steinmetz Extension</td>
</tr>
<tr>
<td>OOD</td>
<td>Object-Oriented Design</td>
</tr>
<tr>
<td>PEG</td>
<td>Power and Energy Group</td>
</tr>
<tr>
<td>PSIM</td>
<td>PowerSIM simulation package</td>
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<tr>
<td>SCI</td>
<td>Serial Communication Interface</td>
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<tr>
<td>SE</td>
<td>Steinmetz Equation</td>
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<tr>
<td>SFD</td>
<td>Squared-Field-Derivative</td>
</tr>
<tr>
<td>SMPS</td>
<td>Switched Mode Power Supplies</td>
</tr>
<tr>
<td>THD</td>
<td>Total harmonic distortion</td>
</tr>
<tr>
<td>VSI</td>
<td>Voltage Source Inverter</td>
</tr>
<tr>
<td>WCSE</td>
<td>Waveform Coefficient Steinmetz Equation</td>
</tr>
<tr>
<td>WUF</td>
<td>Window Utilisation Factor</td>
</tr>
<tr>
<td>ZVS</td>
<td>Zero Voltage Switching</td>
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# List of Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>$A$</td>
<td>Defined in Equation 5.1</td>
</tr>
<tr>
<td>$A_c$</td>
<td>Cross-sectional area of magnetic core</td>
</tr>
<tr>
<td>$A_L$</td>
<td>Inductance per turn</td>
</tr>
<tr>
<td>$A_p$</td>
<td>Area product</td>
</tr>
<tr>
<td>$A_P$</td>
<td>Bare wire conduction area</td>
</tr>
<tr>
<td>$A_{str}$</td>
<td>Defined in Equation 4.18</td>
</tr>
<tr>
<td>$A_t$</td>
<td>Total surface area of wound magnetic device</td>
</tr>
<tr>
<td>$A_w$</td>
<td>Cross-sectional area of conductor</td>
</tr>
<tr>
<td>$a$</td>
<td>Transformer turns ratio</td>
</tr>
<tr>
<td>$\hat{B}$</td>
<td>Peak flux amplitude</td>
</tr>
<tr>
<td>$B_{max}$</td>
<td>Maximum allowable flux density</td>
</tr>
<tr>
<td>$B_o$</td>
<td>Optimum flux density</td>
</tr>
<tr>
<td>$B_{sat}$</td>
<td>Saturation flux density</td>
</tr>
<tr>
<td>$C_{lv}$</td>
<td>LV dc-link capacitor</td>
</tr>
<tr>
<td>$C_{hv}$</td>
<td>HV dc-link capacitor</td>
</tr>
<tr>
<td>$D$</td>
<td>Duty cycle</td>
</tr>
<tr>
<td>$d$</td>
<td>Layer thickness</td>
</tr>
<tr>
<td>$d_{str}$</td>
<td>bare strand diameter litz-wire</td>
</tr>
<tr>
<td>$f$</td>
<td>Frequency</td>
</tr>
<tr>
<td>$F_R$</td>
<td>Ratio of ac to dc resistance in round conductor</td>
</tr>
<tr>
<td>$F_R^{foil}$</td>
<td>Ratio of ac to dc resistance in foil</td>
</tr>
<tr>
<td>$F_R^{litz}$</td>
<td>Ratio of ac to dc resistance in litz-wire</td>
</tr>
<tr>
<td>$g_{max}$</td>
<td>Maximum gap length</td>
</tr>
<tr>
<td>$h$</td>
<td>Foil thickness</td>
</tr>
</tbody>
</table>
### List of Symbols

- $i_{dclv}$: Battery current
- $I_{\text{max}}$: Maximum inductor current
- $I_{\text{rms}}$: RMS value of inductor current
- $J_0$: Current density
- $K_c$: Steinmetz equation material parameter
- $K_i$: Current waveform factor
- $K_t$: Dimensional constant
- $k$: Number of strands of each litz bundle
- $k_a$: Dimensional constant
- $k_u$: Window utilisation factor
- $k_{u,\text{sel}}$: Selected values of $k_u$
- $L_{\text{cal}}$: Calculated inductance
- $L_{\text{leak}}$: Leakage inductance
- $L_m$: Magnetising inductance
- $L_{\text{target}}$: Target inductance
- $L_{20}$: Inductance @ 20 kHz
- $l_c$: Magnetic path length of core
- $l_{\text{gap}}$: Air gap length
- $\text{MLT}$: Mean Length of a Turn
- $N$: Number of turns
- $N_l$: Number of layers
- $N_{\text{pri}}$: Number of turns of primary winding
- $N_{\text{sec}}$: Number of turns of secondary winding
- $N_{ll}$: Effective number of layers
- $P$: Distance between centres of strands in a layer
- $P_{\text{cu}}$: Copper loss
- $P_{\text{cu, max}}$: Maximum copper loss
- $P_{fe}$: Core loss
- $P$: Power
- $P_D$: Maximum allowable heat dissipation
- $P_{\text{eddy}}$: Eddy current loss
- $P_{\text{excess-eddy}}$: Excess eddy current loss
<table>
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<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>$P_{\text{hysteresis}}$</td>
<td>Hysteresis loss</td>
</tr>
<tr>
<td>$P_v$</td>
<td>Power loss per unit volume</td>
</tr>
<tr>
<td>$P_{\text{winding, ac}}$</td>
<td>Winding AC power loss</td>
</tr>
<tr>
<td>$P_{\text{winding, dc}}$</td>
<td>Winding DC power loss</td>
</tr>
<tr>
<td>$R_{\text{ac}}$</td>
<td>AC resistance of a round wire winding</td>
</tr>
<tr>
<td>$R_{\text{foil}}$</td>
<td>AC resistance of foil winding</td>
</tr>
<tr>
<td>$R_{\text{ac}}^\text{foil}$</td>
<td>AC resistance of foil winding</td>
</tr>
<tr>
<td>$R_{\text{foil}}$</td>
<td>AC resistance of litz-wire winding</td>
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<tr>
<td>$R_{\text{ac}}^\text{litz}$</td>
<td>AC resistance of litz-wire winding</td>
</tr>
<tr>
<td>$R_{\text{dc}}$</td>
<td>DC resistance of a round wire winding</td>
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<tr>
<td>$R_{\text{foil}}$</td>
<td>DC resistance of foil winding</td>
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<tr>
<td>$R_{\text{dc}}^\text{foil}$</td>
<td>DC resistance of foil winding</td>
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<tr>
<td>$R_{\text{dc}}^\text{litz}$</td>
<td>DC resistance of litz-wire winding</td>
</tr>
<tr>
<td>$R_L$</td>
<td>Resistance @ 20 kHz</td>
</tr>
<tr>
<td>$r$</td>
<td>Radius of round conductor</td>
</tr>
<tr>
<td>$\Re_{\text{core}}$</td>
<td>Reluctance of core</td>
</tr>
<tr>
<td>$\Re_{\text{gap}}$</td>
<td>Reluctance of airgap</td>
</tr>
<tr>
<td>$\Re_{\text{total}}$</td>
<td>Total reluctance</td>
</tr>
<tr>
<td>$T$</td>
<td>Period of waveform</td>
</tr>
<tr>
<td>$T_{\text{amb}}$</td>
<td>Ambient temperature</td>
</tr>
<tr>
<td>$T_{\text{max}}$</td>
<td>Maximum operating temperature</td>
</tr>
<tr>
<td>$VA$</td>
<td>Voltage rating of winding</td>
</tr>
<tr>
<td>$V_c$</td>
<td>Volume of core</td>
</tr>
<tr>
<td>$v_{\text{dchv}}$</td>
<td>High voltage side</td>
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<tr>
<td>$v_{\text{dchv}}$</td>
<td>Low voltage side</td>
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<tr>
<td>$V_w$</td>
<td>Volume of winding</td>
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<tr>
<td>$W_a$</td>
<td>Window winding area of core</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Steinmetz equation material parameter</td>
</tr>
<tr>
<td>$\alpha_{20}$</td>
<td>Temperature co-efficient of resistivity at 20°C</td>
</tr>
<tr>
<td>$\beta$</td>
<td>Steinmetz equation material parameter</td>
</tr>
<tr>
<td>$\Delta$</td>
<td>Ratio $d/\delta_0$</td>
</tr>
<tr>
<td>$\Delta_B$</td>
<td>Flux density ripple</td>
</tr>
<tr>
<td>$\Delta_T$</td>
<td>Temperature rise</td>
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<tr>
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<td>Skin depth</td>
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<tr>
<td>$\phi$</td>
<td>Phase displacement</td>
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<tr>
<td>Symbol</td>
<td>Description</td>
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<td>--------</td>
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<tr>
<td>$\theta$</td>
<td>Small phase displacement</td>
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<tr>
<td>$\gamma$</td>
<td>Core to DC copper loss ratio</td>
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<tr>
<td>$\gamma_{sel}$</td>
<td>Selected values of $\gamma$</td>
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<tr>
<td>$\mu$</td>
<td>Permeability</td>
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<tr>
<td>$\mu_0$</td>
<td>Magnetic permeability of free space</td>
</tr>
<tr>
<td>$\mu_{\text{eff}}$</td>
<td>Effective relative permeability</td>
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<tr>
<td>$\mu_{\text{opt}}$</td>
<td>Optimum value of effective relative permeability</td>
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<tr>
<td>$\mu_r$</td>
<td>Relative permeability</td>
</tr>
<tr>
<td>$\eta$</td>
<td>Porosity factor</td>
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<tr>
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<td>Electrical resistivity at 20°C</td>
</tr>
<tr>
<td>$\rho_w$</td>
<td>Electrical resistivity</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>Electrical conductivity</td>
</tr>
<tr>
<td>$\xi$</td>
<td>Defined in Equation 2.8</td>
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</table>
Chapter 1

INTRODUCTION

1.1. Background

The process of transforming electrical energy from one level of voltage or frequency to another is called electrical energy conversion. Power electronic converters are a class of electronic systems that perform this transformation using power semiconductor devices and magnetic components for switching, isolation and filtering. Power electronic converters operate from power levels of a few Watts up to Megawatts depend on the given application. They are the major building blocks for electronic power supplies, battery chargers, distributed generation systems, electric vehicles (EV) and renewable energy generation using solar and wind power, to name just a few applications.

Two major objectives for power electronic converters are higher efficiency and improved reliability. These can be achieved by reducing converter losses using better switching devices and lower loss magnetics. The increasing adoption of new wide band-gap semiconductor devices is currently leading to lower conduction and switching losses, allowing converter operation at much higher switching frequencies. However, because the use of magnetic components is essential for power electronic converters, a reduction in the losses of semiconductor devices alone is not sufficient to make adequate continuing advances in the efficiency of a power electronic converter. Hence matching loss reductions for magnetic components are also required to achieve significant improvements in overall converter efficiency.

Magnetic components (i.e. inductors and transformers) provide energy storage and energy transfer functionality as well as a basic low pass filtering role, and are indispensable elements of modern power electronic converters. Often they are the bulkiest and the most costly components in a power converter. In addition, they can incur considerable losses, which may be similar or even greater than the semiconductor device
switching and conduction losses. Hence the design of magnetic components is critical to avoid the overall design of a power electronic converter being less effective or possibly even unviable.

The design of magnetic elements has been a major and challenging research area for decades, and is typically considered in the following categorisation:

- The low frequency range (50Hz).
- High switching frequencies, ranging up to the MHz region.
- More recently, for higher power converters switching at intermediate frequencies.

The continuing evolution of higher power switching devices and higher frequency magnetic materials allow for continually increasing the switching frequency of the converters. However, this continuing increase in switching frequency makes the design of magnetic components a more and more challenging task as increasing levels of higher frequency losses need to be taken into account and minimised.

There is substantial published literature exploring the design of high frequency magnetic components, particularly within low power range. However, there is no clear equivalent design methodology for the higher power intermediate frequency range. Essentially, the challenge with this power range is that the operating frequency of the switching devices cannot yet be readily increased above 20kHz-50kHz, and the direct translation of the higher frequency magnetic component design methodologies into this higher power intermediate frequency range often leads to unsatisfactory results. Consequently, there is a need to modify the low power high frequency design approaches to suit the intermediate frequency range.

One important consideration for the design of magnetic components in the intermediate frequency region is the converter application context, including in particular the harmonic currents produced by the converter switching processes. Hence the next section of this chapter reviews the two major types of converter switching topologies in this context, and the harmonic currents that flow through their magnetic filter component as a result of their switching action. Following this review, the major focus of the research work presented in this thesis is summarised, together with an outline of the various thesis chapters and identification of the original contributions of the work.
1.2. Harmonic Currents produced by Power Electronic Converters

Magnetic design is conventionally based on sinusoidal excitation, a legacy of low frequency 50/60 Hz design processes. However, a critical starting point for the design of a higher frequency power converter magnetic component is the harmonic waveforms produced by the converter. Fundamentally, two types of converter output currents can be identified:

- **Low Frequency (LF) fundamental with HF ripple**: AC converters such as Voltage Source Inverters are typically controlled by a Pulse Width Modulation (PWM) strategy, which produces a low frequency more or less sinusoidal fundamental output with high frequency harmonics clustered around multiples of the converter switching frequency.

- **High Frequency (HF) fundamental and baseband harmonics**: High frequency direct conversion systems such as a Dual Active Bridge (DAB) produce an essentially square wave high frequency current through the magnetic component which has a high frequency fundamental component with associated baseband harmonics.

While this sharp contrast between these two types of converter output currents does not significantly change the magnetic component design process in principle, it does have a major influence on the practical design process. It is also important to appreciate the changing role of the magnetic elements of the converter for these two types of conversion systems. For an AC inverter, the switching frequency harmonics are generally significantly higher than the primary fundamental harmonic, and the inductor can be designed to substantially reduce the magnitude of these harmonics with little impact on the primary fundamental component. For a direct conversion system, the inductor’s role becomes more to mitigate the baseband harmonics as much as possible, but not at the expense of a significant reduction in the magnitude of the primary high frequency fundamental component.

Since these two types of converters produce quite different levels of harmonic components, it is important to understand their operation and the harmonics they create before proceeding to the magnetic design process, as will now be explained.
1.2.1. VSI Topology

One of the popular topology choices for high power density converters used in grid interfaces is the Voltage Source Inverter (VSI). Figure 1.1 shows the schematic of a typical single phase Voltage Source Inverter (VSI) which consists of two phase legs connected to a common DC bus. Each phase leg is switched in a complementary fashion by comparing reference waveforms against a triangular carrier waveform [5]. The process is called Pulse Width Modulation (PWM), and offers a number of features including the use of a fixed DC supply from a rectifier, stable open-loop operation using a constant volt-per-hertz control and low cost [6-9]. Refs [10-12] address this concept in detail.

Figure 1.2 illustrates the operating waveforms (voltage and current waveforms) of a VSI. The switched output voltage is a high frequency variable duty cycle switched waveform as shown in Figure 1.2(a) that contains both the desired fundamental sinusoidal ac component (the reference waveform) and a rich harmonic voltage spectrum clustered around multiples of the carrier frequency [5]. From Figure 1.2(b) it can be seen that the current waveform is a sinusoidal (fundamental) waveform, with a superimposed high frequency ripple created by the switching harmonics. Hence the harmonic components that need to be filtered by the inductor occur at much higher frequencies than the primary fundamental component, as shown in Figure 1.2(c), and the magnitude of the harmonic currents compared to this fundamental current component are relatively small as a consequence of this filtering action.

![Figure 1.1. Schematic of single phase Voltage Source Inverter (VSI) with R-L load.](image-url)
Figure 1.2. Single phase VSI operating waveforms (a) DC link voltage waveform (b) Current waveform (c) Harmonic current (frequency domain).
1.2.2. DAB Topology

DC-DC converters are a major element for a wide variety of power supplies, with bidirectional alternatives being the main building block for the intermediate storage systems that are now becoming popular in grid connected applications [13]. One of the most attractive bidirectional systems is an isolated DC-DC converter, which can transfer energy in either direction between two independent DC voltage sources. Bidirectional isolated DC-DC converters and their topology variations are well described in [14-22], with the most common topology choice for higher power applications being the Dual Active Bridge (DAB) structure proposed in [14]. This structure has the advantages of inherent isolation, bidirectional power flow, step up or step down voltage ratios, soft Switching (ZVS), high efficiency and high reliability [23, 24]. Comparisons and evaluations of the DAB with other DC-DC topologies are presented in [14, 18, 24-29].

The basic schematic of a DAB is shown in Figure 1.3, and comprises two single-phase H-bridges linked through an isolating/scaling transformer $T$ and an intermediate inductor $L$. Each bridge phase leg is made up of complementary switch pairs $(S_1, S_2)$ and $(S_3, S_4)$ for the primary high-side H-bridge, and $(S_5, S_6)$ and $(S_7, S_8)$ for the low-side secondary H-bridge. The intermediate-frequency transformer $T$ provides galvanic isolation between the two bridges and converts the primary voltage level to a secondary side voltage level while the intermediate inductor $L$ smooths the current flow between the two H-bridges (note that often the coupling transformer is deliberately designed with a significant leakage inductance that then also acts as the intermediate filter inductance). The LV dc-link capacitor $C_{lv}$ provides the bulk energy storage on the LV side, while the HV dc-link capacitor $C_{hv}$ creates a decoupled DC link to a separate AC-DC conversion stage (not shown), which is typically a voltage-source converter.

![Figure 1.3. Schematic of Dual Active Bridge (DAB).](image-url)
Figure 1.4 shows the major operating waveforms of a DAB, where it can be seen that the voltage waveforms produced by each H-bridge are three-level square waves that are displaced by a phase shift to control the power flow between the bridges, as shown in Figure 1.4(a). The difference between these two bridge output voltages generates a rectangular waveform across the intermediate inductor with significant periods of zero voltage, which creates a trapezoidal current flowing through the inductor with the fundamental switching frequency as the dominant harmonic as shown in Figure 1.4(b). Hence in terms of harmonic considerations for the magnetic component design, a DAB converter produces a full magnitude harmonic current at its switching frequency as its fundamental frequency, with a reducing set of harmonic multiples of this frequency, as shown in the last plot of Figure 1.4(b). This is in sharp contrast to the much wider frequency range of the harmonic currents that are created by a PWM controlled inverter.

The main advantage of the DAB is its low number of components, which allows a more compact converter design compared to other bidirectional DC-DC converter topologies. However, its drawback is the presence of a high rms/peak circulating current flowing through the high frequency AC link [18, 30-33]. This circulating current increases the conduction losses and so requires a more sophisticated design of magnetic components to avoid impacting on the overall efficiency of the converter [32].

This section has reviewed two basic types of converters – a VSI and a DAB – in terms of the harmonic switching components that they produce. For a VSI type converter, where the switching harmonics are well away in frequency from the target fundamental, the magnetic filter component can be made sufficiently large to keep the higher frequency harmonic current magnitudes small and their impact on the magnetic component design and performance therefore reduces. For the DAB type converter, the magnetic design task more difficult, since the main fundamental harmonic frequency is quite high anyway, and it is also not possible to filter the low order baseband harmonic components above this frequency so much without impacting on the fundamental component. Hence the main focus for this thesis is on the design of magnetic components for this second type of application, since it is the more challenging context.
Figure 1.4. DAB operating waveforms, (a) voltage waveforms (b) voltage, current and harmonic current of the series inductor.
1.3. Aim of the Research

The aim of the research presented in this thesis is to develop an improved methodology for the design of higher power inductors operating at the intermediate switching frequency range. There are particular challenges in designing magnetic components in this frequency range because it requires a series of trade-off design decisions (design objectives) between parameters such as cost, volume and temperature rise that are often hard to balance. One strategy to address these challenges is to create a multivariable optimising type system using an expert system approach that addresses the complexity of the design inter-relationships by iterating and trading-off objectives to achieve a design answer. However, validation of a design achieved by this approach usually requires physical construction and evaluation, since decisions made by an expert system are typically supported more by experiential data than from a more fundamental understanding of intrinsic analytical theory. The alternative approach is design the magnetic component directly from analytical principles. However, as identified earlier, the application of currently identified theoretical design principles based on higher frequency lower power level designs often does not lead to an effective design outcome. Hence both approaches have been explored in this research, as follows:

The first stage of this work focuses on the development of a knowledge-based advisory system for design of magnetic components. The thesis proposes a novel inductor/transformer knowledge-based design advisor support system to integrate the outcome of an initial primary process with ANSYS® Maxwell as an FEA tool. The objective of the system is to recommend design alternatives to a user to meet a set of input specifications and advise optimum FEA configurations for the selected design. The system then evaluates the design and provides multi-objective optimization in terms of volume, temperature rise and localized hot spots.

The second part of the research is to identify the limitations of present-day analytical design approaches and then to use this understanding to develop a new design approach that is more suitable for higher power intermediate frequency inductors. A typical inductor design procedure involves several iterative steps including core selection, winding selection, power handling capacity and thermal management. One major challenge for the design process is to select an appropriately sized core that meets the design requirements, particularly in terms of volume and anticipated temperature rise.
Chapter 1  Introduction

Hence this thesis then presents an improved core selection methodology for higher power inductors. The research begins by identifying the issues that make conventional higher frequency inductor design methods less suitable for lower frequency inductors, and then proposes an improved approach that better suits the design of intermediate frequency range components. Finally, a 3D graphical method for optimum selection of core size is presented that requires only a small number of iterations to come to a final core selection (in fact often only one design pass is required). The performance and feasibility of this new design approach has been verified for the series inductor of the Dual Active Bridge (DAB) converter, chosen because of its more challenging High Frequency (HF) fundamental, as discussed previously.

1.4. Structure of the Thesis

The thesis is organised as follows.

Chapter 1 (this chapter) provides a context and overview for the work performed in this thesis and presents the structure of the thesis.

Chapter 2 presents a literature review in two main sections. The first section focuses on the loss modelling of magnetic components, separately describing core and winding loss modelling and reviewing in particular high frequency effects on winding losses with appropriate analytical expression. The second section reviews various computer aided techniques that can be used for the magnetic component design, introduces the general concept of multi objective optimisation and considers its application to the magnetic design problem.

Chapter 3 proposes an innovative knowledge based system which adapts the optimisation process to suit magnetic design in a more comprehensive way to be able to deal with the particular challenge of intermediate frequency range. It presents a case study and compares its performance at the end of the chapter. But the problem is there is no insight into the final answer in order to have more confidence that the answer actually works. There are two ways to validate the final answer; by building the component which will confirm the design but will not provide any particular insight into the underlying methodology, or by revisiting the fundamentals to come up with an analytical solution.

Chapter 4 has two main sections. The first section reviews the principles of a generic inductor design process by revisiting the primary fundamental concepts, to
establish a reference point for the work that follows. This section then shows how to proceed with a complete inductor design through a number of steps, illustrating the whole process with a design flowchart. The second section identifies the initial assumptions that underpin the core selection for a generic design process, and then presents an improved design strategy that allows core selection to be made without any heuristic assumptions. The strategy is illustrated using an exemplar practical design.

Chapter 5 presents a detailed description of a complete design process to develop the magnetic components of a commercial 2.5kW bi-directional charging system for a vanadium flow battery. The results of this design process are then validated using simulation and experimental investigations.

Chapter 6 provides a description of the simulation and experimental systems used in this research to obtain the simulated and analytical results presented throughout the thesis.

Chapter 7 provides a summary of the work presented. The important contributions are reviewed and proposals are made for future research work.

1.5. Identification of Original Contributions

This thesis presents a number of original contributions to the topic of magnetic component design, as follows:

- Development of a semi-automatic and user-friendly environment to provide assistance in design of high power inductors. It integrates a knowledge based advisory system with ANSYS-based Finite Element Analysis (FEA) to fine tune the outcome of proposed 3D graph-based technique.
- Identification of the shortcomings within the existing body of knowledge concerning the design of magnetic components operating at intermediate switching frequencies.
- Development of a novel 3D graph-based method for core selection. Unlike other techniques, it does not require any assumptions to be made for parameter initialization. This improves the design process and outcome greatly in terms of efficiency, size, reliability and total number of iterations required to complete a design.
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1.6. List of Publications

The contributions and ideas presented in this thesis have been published by the author during this research program. These publications are listed below:


Chapter 2

LITERATURE REVIEW

Magnetic design has been the topic of intensive research for decades, resulting in a substantial amount of accumulated knowledge. This chapter reviews the fundamental understandings arising from this work that can be used for the design of magnetic components, highlighting the design challenges of such a design process, and reviewing computer based analysis techniques that are appropriate for use with magnetic design and optimisation processes. The published literature is classified and reviewed in two major areas, as follows:

- Loss Modelling of magnetic components, subdivided into core and winding loss modelling;
- Multi Objective Optimisation techniques suitable for magnetic component design.

The chapter begins by reviewing loss modelling of magnetic components, introducing core and winding losses as the primary sources of power loss in a magnetic component. Section 2.1 describes the major core loss modelling approaches that have been presented in the literature, comparing their complexity and providing a summary of their classifications, advantages and limitations. Section 2.2 provides a detailed literature review of winding loss modelling, introducing high frequency effects such as skin and proximity effects, and considering their contribution to the ac resistance of a winding as well as their calculation methods. This section finishes by introducing fringing effect around an air gap, describing the challenges of accurate prediction of this effect and its contribution to the total winding loss.

Finally, Section 2.3 introduces the fundamentals of multi objective optimisation, by presenting an overview of computer-aided design techniques with a focusing on optimization for the process of magnetic component design.
2.1. Core Loss Modelling

Several different approaches are available to model the core loss of a magnetic component. These approaches can be generally subdivided into hysteresis models [34, 35], a loss separation approach [36, 37] and various empirical methods [34, 37].

2.1.1. Hysteresis Models

Hysteresis core loss models are usually based on Jiles-Atherton or Preisach models. The Jiles-Atherton model [35] is based on microscopic energy calculation using differential equations. This model uses an iterative process to determine core loss, and requires a large number of parameters to describe both the static and dynamic behaviour of ferro- and ferri-magnetic behaviour. Generally this modelling approach is not so useful in practice because it requires a large number of input parameters, many of which are usually not available from a core manufacturer [38].

Preisach’s model introduces a statistical approach for the description of the time and space distribution of domain-wall movement [34]. The magnetic material characteristics are represented by a distribution function which is known as Preisach function or weighting function. There are different procedures and measurement methods to determine this weighting function, and the choice of the measurement and evaluation method of domain-wall movement and the hysteresis curve has an impact on the precision of the distribution function of the Preisach model. However, similar to the Jiles-Atherton model, this model is of limited practical use as it requires extensive measurement and computation efforts to achieve a calculated core loss result, making it a time consuming procedure [39].

2.1.2. Loss Separation Models

Loss separation approaches [36, 37] are based on the assumption that three different effects contribute to magnetic loss - static hysteresis loss, eddy current loss and excess eddy current loss, which combine to produce an overall core loss as specified by:

\[ P_{\text{core}} = P_{\text{hysteresis}} + P_{\text{eddy}} + P_{\text{excess,eddy}} \]  

(2.1)

Static hysteresis loss relates to both major and minor hysteresis loops and their allied hysteresis losses. Eddy current loss is caused by the eddy currents that form inside
the core as a consequence of an alternating magnetic field in the core. Excess eddy current loss originates from magnetic domain wall motion inside a core material [40] and depends on the microstructure of the material. Similar to hysteresis models, the results of this approach to core loss calculation are again usually quite good, but the approach is once more not particularly practical as it requires substantial computation and a large amount of input data that is usually not available from a core manufacturer’s datasheet [38].

2.1.3. Empirical Methods

The most popular empirical equation used to determine the core losses in a magnetic component is the power law equation [41, 42], i.e.

\[ P_v = K_c f^\alpha \hat{B}^\beta \]  

(2.2)

where \( P_v \) is the power loss per unit volume, \( \hat{B} \) is the peak flux amplitude, \( f \) is the frequency of the sinusoidal excitation and \( K_c, \alpha \) and \( \beta \) are constants determined by the material characteristics usually provided by manufacturer. Equation (2.2) is known as the Steinmetz Equation (SE) since the original equation was proposed by Steinmetz in 1892 (without the frequency dependency) [43]. The SE and its material parameters (the Steinmetz parameters) are valid for a limited range of flux density and frequency. Also, (more importantly) they are only valid for a sinusoidal waveform. This is a major drawback of using the SE approach for modern power electronic converters, since the magnetic core material is routinely exposed to non-sinusoidal flux waveforms as discussed in Chapter 1 [34, 44]. Hence applying SE to non-sinusoidal waveforms can results in significant errors and/or inaccurate loss calculations [34, 38, 44]. To overcome this limitation, a variety of approaches have been explored and developed to determine core losses for a wider variety of waveforms [45], mostly by extending the original SE approach. For example, Reinert et al. [34] introduced a modified Steinmetz equation (MSE) to address the losses associated with an arbitrary flux waveform. According to [34], the loss due to domain wall motion directly depends on the rate of change of magnetic induction with time, i.e. \( d\hat{B} / dt \). Another approach was the generalised Steinmetz equation (GSE), proposed in [44] to avoid the anomalies of MSE. In GSE both the derivative of magnetic induction \( d\hat{B} / dt \) and its instantaneous value \( \hat{B}(t) \) are taken into account [39, 44]. This approach correlates better with the original SE for sinusoidal excitation. The GSE was further improved in [46] to deal with minor hysteresis loops, by
replacing the instantaneous value of magnetic induction $B(t)$ with its peak-to-peak value $\Delta B$. This results in the improved Generalised Steinmetz Equation (iGSE) of:

$$P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta - \alpha} dt$$

(2.3)

with

$$k_i = \frac{K_c}{(2\pi)^{\alpha - 1} \int_0^{2\pi} |\cos \theta|^\alpha \cos^2 \theta d\theta}$$

(2.4)

and where the parameters $K_c$, $\alpha$ and $\beta$ are the same parameters as used in the original SE approach, Eq. (2.2).

The iGSE approach provides a relatively accurate loss calculation method for any flux waveform, without requiring extra characterisation of core material properties beyond the basic Steinmetz parameters [46, 47]. However, it has the drawback of not accounting for the impact of DC bias conditions on the core loss [45, 47]. A further improvement of iGSE was developed in [48] to consider the effect of DC bias, leading to the improved iGSE approach ($i^2$GSE). This method requires five new parameters in addition to the basic Steinmetz parameters to accurately calculate the core loss, which makes it more complicated. Furthermore these parameters are not normally provided in the datasheet by the core manufacturer [47].

Other derivations of Steinmetz equation are the Natural Steinmetz Extension (NSE) [49], the Equivalent Elliptical Loop (EEL) [50] and the Waveform Coefficient Steinmetz Equation (WCSE) [51]. However, among all these empirical methods, the iGSE has proved to be the most useful practical method, without requiring parameters beyond the basic Steinmetz parameters. Hence it has become the most widely used approach in practice [39, 52].

2.1.4. Core Loss Summary

The classification of above core loss approaches, their complexity and required data is summarised in Table 2.1. Table 2.2 then summarises the characteristics of three major empirical approaches that are in practical use, identifying their advantages and limitations.
2.2. Winding Loss Modelling and Second Order Effects

The power loss of the windings of a magnetic component includes both dc losses and ac losses, contributes to the overall loss of the component and dramatically increases as the operating frequency increases, primarily due to eddy current effects [53]. Eddy current effects, also known as high-frequency (HF) effects including skin and proximity effects, are generally modelled using a frequency dependant winding ac resistance $R_{ac}$. Skin effect is a result of eddy currents flowing within the conductor itself, while proximity effect is caused by the adjacent conductors carrying current. Both effects cause the current density to be non-uniform across the cross-section of the conductor, decreasing the current-carrying ability of the conductor at high frequency and increasing its effective ac...
Chapter 2 Literature Review

resistance [54, 55]. The severity of these effects within a winding depends on the conductor wire size and converter operating frequency.

2.2.1. Skin Effect

A time-varying (ac) current flowing through a conductor generates an alternating magnetic field inside the conductor. This in turn induces eddy currents as shown in Figure 2.1 based on Faraday’s law of electromagnetic induction. Based on Lenz’s law, these eddy currents produce an opposing magnetic field inside the conductor, which makes the current density pattern non-uniform as shown in Figure 2.2(b). Accordingly, the current tends to flow near the conductor surface instead of through the entire cross-section and so the current density decreases from the surface of the conductor towards its centre [54, 55], shown in Figure 2.2. The depth of current penetration in the conductor is defined by the skin depth, $\delta_0$ which describes the radial distance from the conductor surface to the point where the current density drops to $1/e$ of its maximum value, i.e.

$$
\delta_0 = \frac{1}{\sqrt{\pi \mu_0 \sigma f}}
$$

(2.5)

where $\mu_0$ is the permeability of free space, $\sigma$ is the conductor conductivity and $f$ represents the waveform frequency. Figure 2.3 illustrates the relationship between the skin depth and frequency for a copper conductor material.

![Eddy Current Effect](image)

Figure 2.1. Eddy current effect in a round conductor (a) cross section view of a round conductor (b) self-induced eddy current effect in a round conductor.
Figure 2.2. Skin effect in a round conductor (a) cross section of a round conductor carrying current (b) current density distribution due to skin effect (c) magnetic field intensity due to skin effect.

Figure 2.3. Skin depth vs. frequency for Copper material (Cu).
2.2.2. Proximity Effect

Proximity effect occurs by the magnetic field generated by the current flowing through neighbouring conductors inducing additional current inside the conductor. This forces the current density to decrease in the vicinity of the nearby conductor and intensify in the opposite side, leading to a non-uniform current density distribution [54, 55]. Figure 2.4 illustrates this effect for two adjacent round conductors. The number of winding layers has a substantial impact on proximity effect. Hence for multi-layer windings proximity effect usually dominates over skin effect [54, 56].

Skin effect together with proximity effect gives rise to the total winding loss of the magnetic component by increasing the effective total ac resistance of the winding. Urling et al. [53] reviewed a number of analytical approaches to characterise the HF effects on windings. The first approaches were developed in 1940 by Bennett and Larson [57] to formulate the HF winding loss for sinusoidal waveforms, based on a one-dimensional (1-D) field solution. In 1966, Dowell [58] adapted their method specifically for a transformer winding, now known as Dowell’s method. This analysis was extended to predict the HF winding loss for non-sinusoidal waveforms in [59-63] using Fourier analysis. Hurley et al. [64] then developed a new formula applicable to any arbitrary waveform to find the optimum foil or layer thickness, without requiring calculation of Fourier coefficients at each harmonic frequencies. The ratio of ac to dc resistance in round conductor using Dowell’s method is given by Eq. (2.6) for a winding with \( N_l \) layers [64]:

\[
F_R = \frac{R_{ac}}{R_{dc}} = \Delta \left[ \frac{\sinh 2\Delta + \sin 2\Delta}{\cosh 2\Delta - \cos 2\Delta} + \frac{2(N_l^2 - 1)}{3} \frac{\sinh \Delta - \sin \Delta}{\cosh \Delta + \cos \Delta} \right]
\]

with

\[
\Delta = \frac{d}{\delta_0}
\]

where \( d \) is the layer thickness and \( \delta_0 \) is the skin depth.

An alternative approach to calculate the HF winding loss in round conductors using a Bessel function was presented in [65-68]. The orthogonality of skin effect and proximity effect was then used by Ferriera [67] to develop a more accurate winding loss calculation method [69]. Equation (2.8) shows their proposed expression for the skin and proximity effect in a round conductor with the radius of \( r \) [69]:
Figure 2.4. Proximity effect and current density distribution in two neighboring round conductors carrying currents: (a) in opposite directions (b) in the same direction.
Chapter 2  Literature Review

\[ F_R = \frac{R_{ac}}{R_{dc}} = \xi \left[ \frac{ber\xi ber'\xi - bei\xi ber'\xi}{ber\xi ber'\xi + bei\xi ber'\xi} \right] \]

\[ - 2\pi(2m - 1)^2 \frac{ber_2\xi ber'\xi - bei_2\xi bei'\xi}{ber_2\xi ber'\xi + bei_2\xi ber'\xi} \]

with

\[ \xi = \frac{\sqrt{2r}}{\delta_0} \]

where the Kelvin functions \( ber \) and \( bei \) are the real and imaginary parts of Bessel functions of the first kind.

A comprehensive review and comparison of various methods for determining HF winding loss can be found in [70]. Both Dowell’s formula and Ferriera’s Bessel-function expression are based on one-dimensional (1-D) field assumption to characterise HF effects in the windings. However, the accuracy of these methods was compared and evaluated in [71] using two-dimensional (2-D) finite-element method (FEM) simulation for a round conductor at HF, and the results showed substantial errors as high as 60%. Reference [71] concluded that the errors stem from the position that Dowell’s formula underestimates the proximity effect while Ferriera’s method overestimates it at high frequency. Integrating direct numerical methods with finite element analysis (FEA) tools improved their accuracy [71, 72] for different winding configurations, although increasing processing time as a consequence. Sullivan [72] introduced the squared-field-derivative (SFD) method to calculate the HF winding losses in round-wire or litz-wire. This analysis showed that the SFD method is capable of loss calculation for both 2-D and 3-D field effects in multiple windings with an arbitrary waveform in each winding, and computationally is more efficient compared to FEA.

Skin and proximity effects are highly dependent on the conductor thickness and can be diminished using litz-wire and/or interleaved windings. Litz-wire conductors are made up of many strands of individually insulated wires, which are twisted into a bundle as shown in Table 2.3. Skin and proximity effects in litz-wire windings reduce to losses at the strand-level and bundle-level [63, 73]. A number of studies have been carried out for the litz-wire construction to provide guidance for an optimum selection of the number and diameter of the litz-wire strands [63, 74, 75]. Skin and proximity effects within litz-wire and multi-strand windings have been explored in [54, 68, 73, 76-82]. Cost constraints
of the litz-wire windings have been addressed in [56, 83, 84], where the trade-off between strand diameter, cost and loss has also been described. Wojda et al. [81] adapted Dowell’s equation to describe the ratio of ac to dc winding resistance of litz wire with an effective number of layers of $N_{lt}$. The modified Dowell’s equation for litz wire winding from this reference is given by Eq. (2.10):

$$F_{ltz}^{litz} = \frac{R_{ac}^{litz}}{R_{dc}^{litz}} = A_{str} \left[ \frac{\sinh(2A_{str}) + \sin(2A_{str})}{\cosh(2A_{str}) - \cos(2A_{str})} \right. + \left. \frac{2(k \cdot N_{lt}^2 - 1) \sinh(A_{str}) - \sin(A_{str})}{3 \cosh(A_{str}) + \cos(A_{str})} \right]$$

where $k$ is the number of strands of each litz bundle, $A_{str} = \left( \frac{\pi}{4} \right)^{0.75} \frac{d_{str}}{\delta_0} \sqrt{\eta}$ where $d_{str}$ is the litz-wire bare strand diameter and $\eta$ is the porosity factor or fill factor defined as $\eta = \frac{d_{str}}{P}$. $P$ is the distance between centres of strands in a layer.

Interleaving primary and secondary windings in transformers minimises HF winding losses [85]. A useful comparison between an interleaved winding and an ordinary disk/layer winding is described in [86]. This winding arrangement has been used in many different applications [87-89], and its relationship with the transfer function of the transformer has been presented in [90, 91]. A partial interleaving method is introduced in [92] to simplify a winding’s configuration and attain a higher power density. Finally, [93] proposed a novel technique called maximum interleaving that offers less construction difficulty compared to previous methods.

### Table 2.3. Skin and Proximity effect in Litz-wire

<table>
<thead>
<tr>
<th></th>
<th>Strand-level</th>
<th>Bundle-level</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Skin Effect</strong></td>
<td><img src="image1.png" alt="Skin Effect" /></td>
<td><img src="image2.png" alt="Skin Effect" /></td>
</tr>
<tr>
<td><strong>Proximity Effect</strong></td>
<td><img src="image3.png" alt="Proximity Effect" /></td>
<td><img src="image4.png" alt="Proximity Effect" /></td>
</tr>
</tbody>
</table>

- *Internal and External field*


2.2.3. Fringing Effect

The 1-D models used for winding loss calculations at high frequency usually do not consider losses caused by fringing flux around the air gap. In some cases, this loss can be quite high and results in inaccurate loss prediction and temperature rise, degrading the overall performance of the system if it is ignored. Figure 2.5 illustrates this effect, simulated using ANSYS® MAXWELL. The 2-D effects of the non-ideal field distribution due to air gap and fringing field around the air gap, which contribute to the total winding loss [94], cannot be characterised by 1-D approaches [95]. However, these effects were treated by Wallmeier et al. [95] using 2-D field calculations for a specific geometry. A number of efforts have been carried out to address these issues using FEA tools, mainly focusing on the distributed air gap in planar magnetics [96-98]. In [99, 100] the mirror-image principle was extended to predict the winding loss of gapped cores. Another analytical method to calculate the air gap fringing loss based on the reluctance model of the air gap using a Schwarz-Christoffle transformation was presented in [101]. Accurate calculation of the 3-D air gap reluctance based on an analytical field solution was offered in [102, 103]. Roshen derived simple formulas for the fringing field of a single lumped air gap using scalar potential approach [104, 105]. In 2008 [106], these

![Gapped Inductor (With Air Gap)](image)

Figure 2.5. Fringing effect around the air gap.

24
formulas were extended to thick rectangular and round wire windings using the principle of superposition and 1-D field.

Finally, the effect of fringing field of the air gap can be alleviated by incorporating some practical modifications to the winding configuration or the air gap structure. These modifications include placing the winding conductors away from the edge of the air gap [54, 80, 107] or replacing the lumped air gap with distributed air gap material using material with lower permeability compared to the core [108, 109]. Figure 2.6 illustrates the reduced effect of air gap fringing when material with lower permeability compared to the core (here powder iron material) is inserted into the lumped air gap, simulated again using ANSYS® MAXWELL.

2.3. Multi Objective Optimization

Designing power magnetic components is well recognized as a highly complicated process because of the large number of design variables, component alternatives and mutual dependencies that have to be considered. Therefore, the power magnetic design process has more recently evolved from a quite labour intensive task to one that relies on computers to shorten the development cycle and save the costs. A significant number of

Figure 2.6. Replacing the air gap with powdered iron material to reduce the fringing effect.
studies have been conducted in the area of magnetic component design synthesis and analysis, with fundamental Computer-Aided Engineering (CAE) techniques being frequently employed. This section presents an overview of these studies, identifying their relationships with CAE theories and highlighting the CAE techniques that have been used. The section introduces the general concept of optimisation, reviews its extension to magnetic component design and investigates the potential and limitations of such CAE methods.

2.3.1. Computer Aided Design System

For industrial applications, the design of magnetic components should meet a set of standards in terms of impedance, loss and average temperature rise, weights, dimension [110]. Because of this large number of variables and their complex interrelations, which are often not completely known, designing a power transformer/inductor is a time-consuming task for human designers [111]. The design method usually involves extensive use of experience, rules-of-thumb and application tables from catalogues, and it may require a lot of iteration and time consuming calculation processes to find the optimum design [112]. To overcome this, the methodology has evolved in recent years from a very labour intensive task to one that relies heavily on Computer-Aided Engineering (CAE).

The application of CAE techniques to Inductor/Transformer design has been studied for some time. The earliest approaches used simple computer programs with sequences of calculations, to reduce the human designer’s calculation efforts [110],[113]. More recent approaches include Finite Element Analysis (FEA)-based tools for design feasibility and evaluation [114],[115],[116], knowledge-based systems for design guidance and solution generation [117],[118],[119] and a set of numerical models for design optimization according to a certain objective function [120],[121]. But one major weakness is that these systems are all separate tools that focus only on individual parts of the design process, and do not deliver complete design assistance. Also, the collaboration of those systems is often poor due to their weak interoperability.

2.3.2. Inductor/Transformer Design Challenges

Inductor/Transformer design is essentially an iterative process that looks for a balanced design which satisfies all the specified requirements, within the guidelines of
theoretical models and physical laws. In most of the existing literature and commercial software manuals, the design usually starts by choosing an appropriate magnetic core. The winding structure is then determined, followed by the construction of peripheral components, such as bobbins and external heat exchangers. If at any stage the design does not comply with the constraints or requirements for the component, or if the design is physically clumsy and/or impossible to realize, the designer tries again with a better and more educated choice of core. This process can be quite difficult for higher frequency applications, where the core performance becomes more critical and the choice of magnetic core material is more challenging.

A major difficulty with Inductor/Transformer design is that it involves large numbers of interrelated electrical, physical or geometrical parameters. The complex interrelationships between these parameters brings a number of trade-offs that need to be treated differently according to specific application constraints. For example, reductions in volume and weight can often be achieved by operating the component a higher frequency, but with the penalty of a reduction in efficiency [122]. Where the frequency cannot be varied, selecting a more efficient core material may also help to reduce volume and weight, but with the penalty of increased cost. Using a computer to make judicious trade-offs in these circumstances is a major motivation in building CAE systems for magnetic design.

In addition to the above challenges, predicting the performance of a magnetic design without actually manufacturing it is another challenge for the designers. It has long been realized that prediction using mathematical equations or regression analysis on sample data can result in significant errors for non-standard physical structures [114], especially when complex geometry shapes are involved. Hence numerical methods, such as Finite Element Method (FEM), have been frequently applied to address this issue using CAE. Data mining technologies have also been recently introduced to handle this task.

2.3.3. Overview of CAE

Computer analysis has become an important part in almost all aspects of engineering practice and research. During the past four decades, with the development of Computer Science, the initial view of computing as a means of calculating has given way to viewing computing as a means of solving comprehensive problems, generating solutions, communicating and sharing data, information and knowledge. This results in
vaguely defined boundaries for CAE. In fact, any use of computer software to solve engineering problems can be regarded as CAE. To be more specific, rather than a research topic with a clear definition, CAE is more like a generic name given to the collection of Computer Science technologies that are employed to build computer programs to, either independently or collaboratively, handle engineering tasks. In the rest of this section, a simple model of general engineering tasks is firstly outlined, followed by a list of fundamental CAE techniques. Raphael [123] proposed a simple information transformation model for general engineering tasks included in a design activity, as illustrated in Figure 2.7. Within this model, the following engineering tasks are highlighted:

1) **formulation**: formulate required behavior, based on the design objectives;

2) **synthesis**: generate design solutions (usually physical configurations or spatial arrangements) that satisfy all parameter constraints associated with the required behavior and the external environment, where optimization strategies are involved for better solution generation

3) **analysis**: the solution is analysed to obtain predicted behavior. Simulation or prediction processes are usually required

4) **construction**: build real engineering products based on the design solutions

5) **monitoring**: provide a measured behavior, by monitoring the real engineering products

![Figure 2.7. An information transformation model for general engineering tasks.](image)
6) evaluation: compare the synthesis, analysis and monitoring behavior. The evaluation results may lead to a new formulation or to new synthesis and new analysis.

Within the field of computer-aided Inductor/Transformer design, most of the exiting studies focus on the previously mentioned two challenges: 1) making judicious trade-offs in order to generate compromised design, which belongs to the synthesis task; 2) accurately predicting the performance parameters of a design, which belongs to the analysis task.

A typical list of fundamental CAE techniques is as follows [123]:

- **Database techniques**: Methods regarding the storage and manipulation of engineering data. Typical engineering database applications include: simulation data, drawings, measurement data, cost data, product models, material properties, and etc.

- **Computational mechanics**: Methods for solving the mathematic models for practical engineering problems.

- **Optimization**: Methods for finding the best solutions among all the alternatives according to certain criteria.

- **Knowledge-based engineering**: Methods for integrating human experiences and expertise in computers for them to make decisions and solve problems intelligently.

- **Data mining**: Also called machine learning in some circumstances, which refers to the methods that can automatically discover the implicit properties or relationships of data.

- **Geometric modelling**: Methods for representing and modelling the geometric shapes of the physical objects within engineering tasks.

- **Computer graphics**: Methods for generating visualization of engineering data in the format engineers find most understandable.

For Inductor/Transformer design, the topics that have attracted most interest are computational mechanics, optimization, knowledge-based engineering and data mining. A summary of the literature according to these topics is shown in Table 2.4. The survey
in the next section follows this classification, together with more detailed explanations of these topics.

### 2.3.4. Computational Mechanics

Computational Mechanics is one of the earliest applications of CAE. The significance of this topic is implicitly reflected by the fact that many engineers are still unaware of the potential for the use of computers outside Computational Mechanics [123]. Computational Mechanics is an inter-disciplinary science composed of Computer Science, mathematics and mechanics, and has been widely applied for prediction and understanding of complex engineering systems by solving their corresponding mathematical models using computer programs.

The mostly used method in Computational Mechanics for power transformer design is the Finite Element Method (FEM). FEM has been frequently used in predicting the performance parameters of transformers, especially various types of power loss, by solving differential equations in electromagnetic form. The usage of FEM is usually for the analysis task identified in the previous section. The core idea of FEM is to breakdown the overall complicated problem of solving the differential equations into a series of simpler sub-problems that correspond to an easily solved linear system of equations. The procedure for solving an electromagnetic field problem using FEM is divided into the three steps of [153].

1) Pre-processing: the electromagnetic field problem is defined and prepared to be solved.

2) Solving: the numerical solution of the physical problem is obtained.

3) Post-processing: the solution obtained is prepared to calculate the required electromagnetic field quantities or other macroscopic quantities.
The extensive use of FEM in magnetic component loss prediction starts from the prediction of the power loss. [114] and [115] present feasible FEM models for evaluation of stray eddy current losses, while in [116], calculation of hysteresis loss has been taken into account. Other studies have investigated different aspects. Lin et al. [124] paid attention to loss calculation in the tie plate of the power transformer. Loss distribution in stacked power transformer cores was studied in [125]. Current distribution and loss in foil windings of transformers are analysed in [126] in detail, while a special focus was given to the half-turn effect in [127]. FEM is a powerful Computational Mechanics method for transformer design analysis which solves the complex mathematical equations that define the electromagnetic model of a power transformer. A very strong advantage of FEM is the ability to handle time-dependent and complex geometry problems. However, a significant drawback is that it requires precise mathematical models relating all the relevant parameters, either qualitative or quantitative, to identify the final performance. This is often impractical since the analytical relationships expressing the effect of all the parameters on the transformer performance are often not known [144].

2.3.5. General Concept of Optimisation

Choosing the best option among a set of options is called optimisation. In an optimisation process, the primary step is to model the problem (i.e. the challenge of magnetic design) which includes the identification of the objective (i.e. cost, time, efficiency, loss or volume showing the performance of the system), variables and constraints [154]. The system’s variables impact the objective therefore; the values of these variables need to be found in such a way to optimise the objective.

From a mathematical point of view, optimisation is the minimization or maximization of a function subject to constraints on its variables. The optimisation problem can be expressed from [154, 155] as:

$$\min f(x), \quad x \in \mathbb{R}^n \quad \begin{cases} c_i(x) = 0, & i \in E \\ c_i(x) \geq 0, & i \in I \end{cases}$$  \hspace{1cm} (2.11)

where $x$ is the vector of variables, $f$ is the objective function, $c_i$ are constraint functions defining certain equations inequalities that the unknown vector $x$ must satisfy, $E$ and $I$ are two separate sets of indices representing the equality and inequality constraints, respectively. There are a number of different classifications for optimisation among which convexity is one the major one.
Convexity

Convexity is a fundamental concept in optimisation [154] and convex optimisation is one of the major class of mathematical optimisation problems [156]. The problem can be formulated and solved more reliably and efficiently using convex optimisation.

In a convex optimisation problem the objective and constraint functions are both convex which satisfying the following criteria as shown in (2.12) [154]:

\[ f(\alpha x + (1 - \alpha)y) \leq \alpha f(x) + (1 - \alpha)f(y), \quad \text{for all } \alpha \in [0,1] \quad (2.12) \]

where \( x, y \in \mathbb{R}^n \).

It has been broadly used in combinational optimisation and global optimisation [156]. A lot of practical problems inherently have the convex property, which makes the optimisation formulation and solution easier both in theory and practice [154].

2.3.6. Optimisation

As there may exist many good solutions to an engineering design problem, it is always important to find the best alternative between them. According to the existing literature, optimization strategies for magnetic component design fall into two categories. The first automatically generate design solutions which maximize or minimize mathematical representations that correspond to the engineer’s scope. The second are essentially generate-and-test procedures. Optimization in CAE refers to the first category, which is essentially a set of advanced synthesis strategies. It is also the focus of this section.

The optimization problem entails the optimization (minimization or maximization) of an objective function (also called a fitness function in Genetic Algorithm) subject to certain constraints (restrictions and trade-offs) [153]. There are two types of optimization methods, deterministic optimization and stochastic optimization, which are dominated by Non-Linear Programming (NLP) and Genetic Algorithm (GA) respectively in power transformer design. Figure 2.8 shows the general structure of using such models for design optimization, where the key issue is formalization of the problem, including building objective (or fitness) functions, describing constraints (for NLP) or generating chromosomes (for GA). The objective functions reflect the system’s expectations, namely, what are the designs optimized for. The generation of the objective functions are
Table 2.5. Summary of types of Objective Functions

<table>
<thead>
<tr>
<th>Objective</th>
<th>Literature</th>
</tr>
</thead>
<tbody>
<tr>
<td>Minimization of Manufacturing Cost</td>
<td>[120,128, 129,137]</td>
</tr>
<tr>
<td>Minimization of Total weight</td>
<td>[128]</td>
</tr>
<tr>
<td>Minimization of power loss</td>
<td>[128, 130, 131, 133]</td>
</tr>
<tr>
<td>Maximization of Power Output</td>
<td>[128]</td>
</tr>
</tbody>
</table>
method, a special case of NLP that is efficient and reliable, to solve multiple design optimization problems, including minimum mass, minimum loss and maximum VA capacity. Due to the fact that some of the design variables have to take integer (e.g. number of turns) or discrete values (e.g. core dimensions from magnetic core suppliers’ datasheet), recently Mixed Integer Nonlinear Programming (MINLP) method was employed to format the problem [131, 132], and was solved by Branch and Bound (BB) technique. The MINLP is further combined with heuristic method to improve the computing speed in [120].

Apart from NLP, GA has also been widely used for transformer design optimization. It was firstly employed in [134] to identify the best winding transposition to obtain the lowest current loss. It was then used to generate optimum designs which minimize the total loss based on empirical fitness functions [121, 130, 135]. Other studies have attempted to combine GA and Artificial Neural Network (ANN) to improve the efficiency and accuracy [134, 136]. A recent study has shown that combining GA and FEM helps the system converge to the global optimum in a higher speed [137].

The above optimization models are powerful tools in generating optimum power transformer design. However, a major problem is that they mostly act as "black boxes” and lack the capability to convince users to trust the system [117].

2.3.7. Knowledge-Based Engineering

Over the years, engineers’ expectations of computers have changed from tool-like devices to intelligent machines that can assist them in many aspects, such as giving advice, making decisions, providing explanations, etc. Knowledge-Based Engineering (KBE) thus bridges between CAE and Knowledge-Based Systems (KBS). Although there is no fixed or rigid definition for KBE, the following characteristics are usually present:

1) They all hold a symbolic knowledge structure, such as rules, fuzzy sets, or frame-based knowledge representations which explicitly express the design logic, rather than mathematical equations or numerical representations where knowledge is implicitly captured;

2) The generation of designs includes a reasoning (either rule-based or case-based) process, rather than purely mathematical calculations.
KBE systems in power transformer design can be generally classified into two types:

1) Design knowledge is encoded as a set of deductive rules, and is directly used to guide the design generation, e.g. expert systems [138] and fuzzy systems [112, 139];

2) Design constraints and expertise are captured to validate designs generated by other models (usually direct mapping models), and provide modification advice of the design generation process if the validation process fails, in order to automate the "trial-and-error” processes that are usually conducted manually [117, 140, 141].

For the engineering tasks listed in section 2.3.3, the first type of system is typically used for the synthesis task, while the second type refers to the evaluation task.

The first KBS for designing transformers was named Encore [140]. It generates an initial design based on traditional analytic models, and then applies heuristic modifications represented by rules to generate better design alternatives. The system also implements object-oriented programming techniques to naturally represent design information so that the knowledge captured is more transparent, or understandable, to the user. A follow up study [117] improved this structure by introducing frame-based knowledge representation and refining the modification rules in order to make the system more efficient. Other studies have been done in different directions. An Expert System (ES) [138] for designing transformer for Switched Mode Power Supplies (SMPS) was developed by inputting the design knowledge into an ES shell called Babylon. Holt [112] presented a fuzzy decision support system for magnetic component design, where design expertise is encoded as fuzzy sets. Case-based reasoning has also been combined with traditional rule-based reasoning to provide more practical solutions [119]. A recent study proposed an intelligent design assistant which combines object-oriented, rule-based and computational techniques to assist in the design of power distribution transformers [118].

According to [117], KBE systems have the following advantages:

1) They have transparent knowledge representation making the system understandable and trustable;

2) They do not require training cases;

3) They do not require trial-and-error approaches;
4) They can be systematically implemented. However, although KBE has been widely developed for many other engineering applications, the application of KBE to power magnetic design is still comparatively limited, potentially because they have a limited capability to perform numerical calculations, which is not consistent with traditional design methods for power magnetic components.

2.3.8. Data Mining

Data mining, also called machine learning in some circumstances, refers to a set of techniques that discover implicit patterns in large data sets, and apply the patterns for future use, such as classification or prediction of new data. The development of data mining is motivated by the ambition of enabling computers to summarize general principles or patterns from a large number of observations. Although this goal is still far from being achieved, the applications of data mining in some engineering applications are considered to be successful where there is much implicit knowledge or where the computational model is inaccurate or partly unknown, just like Power Transformer Design.

In the case of power transformer design, data mining is mostly used for the analysis task. Such systems usually come with a database of successful design samples with measured performance parameters. The samples are represented as vectors of input/output pairs, where input vector encodes design parameters and the output vector indicates the performance parameters of interest. Data mining techniques, primarily Decision Tree (DT) and ANN, are then applied to automatically build a model (deductive rules or mathematical equations), which map the input vectors to the output vectors with certain accuracy, from the database. This model is finally used to predict the performance of a new input design.

DT was used for performance analysis in [135, 142, 143]. The systems in these three studies have a similar structure. Namely, several attributes which are selected from design parameters comprise the input vector, and the output performance has only two symbolic values, viz: “acceptable” or “unacceptable”. The system then constructs a decision tree using a learning set of data based on the classification entropy method. The generated tree consists of a set of decision rules which decide whether a design is
"acceptable" or "unacceptable". Finally, a test set of data is used to evaluate the prediction accuracy of the DT. Similar systems have also been built for winding material selection [145].

ANN was introduced to predict core losses in [144] for the first time, where the input vector is composed of nine attributes generated from design parameters, and a Back Propagation (BP) neural network is selected to make the prediction. Nussbaum et al. [146] further evaluated multiple ANN structures for core losses and concluded that a simple BP neural network with a single hidden layer and at most five input parameters is enough to achieve effective prediction. Implementing an ANN to predict transformer temperature rise and inrush current forces on transformer windings has also been studied in [147, 148]. Recent studies have attempted to combine ANN with GA [149] and FEM [150] to improve the performance.

Data mining techniques are apparently quite practical since they directly learn from real samples. The drawback is that the accuracy of such models heavily relies on the data that has been used. Since that can only access limited sources of sample data for isolated applications, the general applicability of these techniques are questionable.

### 2.3.9. Trends and Perspective

Similar to most of the engineering applications, a clear trend for computer-aided power transformer design is the increasing number of hybrid approaches that combine multiple CAE technologies. For example, recent studies show that FEM has been combined with optimization technologies for better performance. It can either be employed as a second round validation for optimal solutions provided by NLP in order to eliminate the possibility of infeasible optimum designs [131], or integrated with GA to avoid local optimal traps [137]. Another typical combination is KBE and FEM, where the knowledge is used as an action guide for the simulation process performed by FEM [151, 152].

Another highlighted trend is the use of data-driven approaches. More and more systems are equipped with databases, and are enabled to learn high-level synthetic information from the databases, such as those reviewed in previous sections. Probably the main reason is the tremendous increase in available computing power, which makes possible the application of computationally-intensive data mining algorithms to practical
large scale problems [153]. Furthermore, besides the studies that have been used for the analysis task (reviewed earlier), data-driven approaches have also been extensively applied for the synthesis task in [145, 157], where DT and ANN techniques were employed for selecting optimal winding structures.

In addition, although the existing systems have proved to be successful, the fast evolving Computer Science technologies also offer a lot more potential for computer-aided power transformer design. For example, a recent development in KBE called Generative Modelling (GM) [158] gives a potential for dynamic synthesis. GM-based systems hold a virtual representation that maps the functional specifications of a design. This virtual representation is a dynamic object that automatically reacts to user’s instructions, and continuously updates itself. Hence the design synthesis is no longer a one-time process, but becomes an interactive process that can sense the dynamics of the environment and provide adaptive design solutions. Other potential approaches may include: employing Semantic Web technologies to enhance collaborative design for power transformer, introducing software agent theory to improve design automation, etc...

2.4. Conclusion

This chapter has reviewed the current knowledge in the literature regarding the design of magnetic components. The first and second sections reviewed the two sources of loss in the magnetic components, namely core and winding losses. The review presented different core loss modelling classifications, their advantages and limitations in terms of excitation waveforms and level of complexity with the corresponding analytical expressions. It then reviewed winding losses, exploring skin depth and its relationship to increasing operating frequency, and presenting commonly used analytical expressions to model high frequency effects. Finally, the last section reviewed multi-objective optimization, and the use of the CAE techniques for magnetic component design.

From this chapter, it can be seen that there has been a large amount of detailed work investigating loss modelling and the design of magnetic components, based on an underlying set of theoretical principles and practical approaches. However, the usage of these concepts for the design of magnetic components in the intermediate frequency area is less comprehensively addressed in the literature, and turns out to be quite challenging in practice. Hence this thesis starts from this point, developing firstly a knowledge based
approach for the design of the higher power inductors operating at the intermediate frequency range, and then revisiting theoretical design principles to develop an improved analytical approach for core selection of magnetic inductors operating in this region.
Chapter 3

Knowledge Based Design Advisory System

Chapter 2 has reviewed the major approaches that have been used to determine core and winding losses within the magnetic devices. The advantages and limitations of various methods described in the literature were identified, and the complexities of high frequency effects on the winding loss, together with their analytical methods, were addressed. The last section of Chapter 2 then focused on the multi objective optimisation techniques that can be used in magnetic design.

This chapter now presents an optimisation approach for the design of magnetic components which is based on an expert system with integrated FEA tools. Unlike more conventional approaches that require some initial assumptions to start a design, the approach presented begins with a set of available core sizes, and then sweeps through them to determine the optimum design. The chapter proposes a marriage between CAE and Inductor/Transformer design as an approach to achieve an optimization goal, by introducing a novel CAE system interface and integrating it to ANSYS® FEA; a popular development tool focusing primarily on FEA simulation, with a knowledge-based design advisory system. A knowledge-based design advisor is then included to generate design solutions to speed up the overall process. One advantage of the proposed system is that it can interpret and evaluate the FEA results to check whether the generated design satisfies

1Materials in this chapter were first published as:
the design requirements during the design process, and then offer ideas about design modification.

3.1. Magnetic Component Design Strategy

The design of a magnetic component is a comprehensive process that involves many iterations in order to find an optimised design which satisfies all specified requirements. The design process is usually divided into four general stages, and the proposed system follows these stages:

- **Core selection:**
  
  The first and one of the most important parts of the design is the selection of an appropriate core size for the component.

- **Conductor selection:**
  
  The conductor type is chosen to have a minimum ac resistance at the converter switching frequency while taking into account second order effects such as skin depth and proximity.

- **Pre-FEA evaluation:**
  
  Once the specifications for both the core and the windings are determined, some of the important performance variables, such as the core and winding losses and consequential temperature rise can be theoretically assessed through a set of electromagnetic formulas without simulation. If the design objectives cannot be met at this stage, the first iterative design processes must be repeated until a satisfactory outcome is achieved.

- **Post-FEA evaluation:**
  
  After the design passes the Pre-FEA assessment, FEA can then be used to achieve a more accurate prediction of the component performance parameters. Similar to the Pre-FEA evaluation, a design which fails in Post-FEA evaluation again requires a restart of the whole design process, until a successful design is finally achieved and returned to the user.
3.2. The System Framework

According to [123], a general design activity should at least contain three engineering tasks, viz:

1) synthesis: generate design solutions that satisfy the constraints composed by the parameters associated with the required behaviour and the external environment;

2) analysis: analyse the generated design solution to obtain predicted behaviour, where simulation is usually required;

3) evaluation: evaluate the predicted behaviour to check whether it satisfies the design requirements.

Based on this concept, in order to assist all the engineering tasks within the Inductor/Transformer design process, multiple technologies are integrated into the proposed system, including data/knowledge base management, expert system, parametric design and FEA. A systematic framework has then been developed to accommodate the interaction and collaboration between these technologies, so that a complete service of Inductor/Transformer design assistance can be delivered.

As shown in Figure 3.1, the framework of the system contains four major components:

1) Knowledge-based Design Advisor
2) Design-APDL Interface
3) ANSYS Program Shell
4) Solution Interpreter.

The general operating process of the system is as follows:

- For a given a set of design specifications, the design advisor first generates a proposed design solution and conducts the pre-FEA evaluation;

- If the solution passes the pre-FEA evaluation, it is then transferred to the Design-APDL interface which parameterises the design into APDL scripts. APDL stands for the ANSYS Parametric Design Language [159], which is essentially a scripting language that automates the FEA tasks.
After the scripts are generated, ANSYS [160], which is a popular product development tool focusing on FEA simulation, is called by a program shell to run the scripts and simulate the design.

The returned FEA results are analysed by a solution interpreter, and are then transferred back to the design advisor to check whether the simulated performance satisfies the design requirements.

The whole system is built in Java. Other techniques, such as Jess [161] and APDL, are also employed. The code detail for the system is shown in Appendix A.
3.2.1. Knowledge-based Design Advisor

The design advisor is a key element in the proposed system. It is essentially an expert system that encodes design rules and knowledge to:

1) generate design solutions (e.g. select a specific core and wire from database);
2) recommend FEA settings;
3) generate modification hints for unsuccessful iterations;

The implementation of the expert system is based on Jess, a Java-based rule engine which supports the development of rules and functions, as well as the reasoning process upon them. In Jess, the knowledge captured can be classified into three categories: facts, rules and functions.

Facts are the actual objects on which the system operates. Similar to the concepts of Object-Oriented Design (OOD), each fact has a template (also called a class in OOD) which defines the structure of facts. For example, a set of design specifications is one Jess fact, with this template, the design specification for a particular case can be asserted as another Jess fact.

Besides facts, mathematical design formulas are encoded using Jess functions, for example, inductance calculation is written as a function.

Moreover, the expert design knowledge, namely the design rules, are formulated as Jess rules. For instance, for a particular design specification, if the window size of the magnetic core does not fit the winding structure, the system should choose another core which has larger window size.

Within the design rule, phase is a fact that indicates the current design phase, and design is a fact that records all the design parameters that have been generated. In addition, windingAreaCal is a Jess function which calculates the winding area based on the number of turns, the cross-section area of the wire, and Magnetic.core.get.window.area is a Java function which checks the integrated datasheet to get the window area of a particular magnetic core. The intended meaning of this rule can be read as: “if the current design phase is ‘window size check’ ” and the calculated winding area is larger than the window area of the selected core, then set the design phase back to select core with an additional condition of choose larger window area, and remove the related design parameters that have been generated.
The general reasoning mechanism can be described as follows. Once the reasoning engine is started, the system initializes three facts:

1) \textit{design spec}, which captures all the input specifications;
2) \textit{design}, which captures the design solutions;
3) \textit{phase}, which records the current design phase.

The engine functions by actively firing all the rules related to the current phase, and updating the value of phase based on the firing rules. This then subsequently fires some other rules. The process will then automatically be repeated until the whole design is completed.

3.2.2. Design-ANSYS Parametric Design Language (APDL) Interface

The task of the design-APDL interface is to build a FEA model based on the design and FEA settings suggested by the design advisor, and translate the FEA model into APDL scripts to automate the FEA tasks. As shown in Figure 3.2, the translation can be generally divided into four parts:

- Setting up the physical environment: The translation interface generates scripts for defining element types and specifying material properties;
- Building and meshing the model: The interface outputs the scripts that draw the geometry and defines the meshing properties based on the geometry. When drawing the geometry, this interface has the capability to choose the right strategy for a specific core shape, e.g. for toroidal cores the strategy is to draw a cylinder then dig a hole in the middle, whereas for U-shape cores, the system first draws several blocks then combines them into a U-shape;
- Applying boundary conditions and loads: After the model is built, scripts for applying boundary conditions and loads are developed based on the input specifications as well as the circuit typology;
- Solving the analysis: Scripts for solving the problem are generated and the solving method is automatically selected according to the recommendation from the design advisor;

APDL Technical details can be found in its online manual [159].
3.2.3. ANSYS Program Shell

A program shell has been developed to call ANSYS to run the generated scripts to execute the FEA task. This call sequence is realized using Java runtime command to create a process which runs ANSYS in batch mode, where the file “design.log” is the APDL script file that is output from the parametric design interface. The final FEA solution is recorded in the file “result.txt”.

Figure 3.2. Design-APDL Interface.
3.2.4. Solution Interpreter

The solution interpreter is responsible for interpreting the returned FEA results. The capabilities of this model are twofold, viz:

1) generate APDL scripts to extract plots and text results from ANSYS based on the user selection;
2) calculate key parameters such as total loss or maximum temperature, and pass them to the design advisor for post-FEA evaluation.

3.3. Case Study

The design advisor has been used to design an Intermediate Frequency (IF) inductor for a Dual Active Bridge (DAB) bi-directional battery charger, which is then compared with a conventional design process for validation purposes.

The user interfaces to the proposed system are shown in Figure 3.3. In Figure 3.4(a), the first window, the user can input design specifications, circuit topology and
Figure 3.4. User Interface Screens: (a) Design input window, (b) Design recommendation window.
some environmental constants, and start to run the system. The design that is recommended by the design advisor is then shown in Figure 3.4(b), second window, where a UI shape ferrite core is selected (N87), with 40 turns of copper winding in this case. The user is also able to change the parameters manually at this stage if the recommendation is not satisfied. After ANSYS is called, the third window, Figure 3.5, allows the user to view the FEA results, where contour plots and vector plots of multiple electromagnetic variables are available. Detailed text results are then calculated and pass to the post-FEA evaluation.

3.4. Optimiser Outcome Verification

Table 3.1 shows the design outcome of the optimiser and the design specifications. This comparison shows that the optimiser produces an inductor with characteristics that are very close to the required design specification.
3.5. Summary

This chapter has presented an expert system which integrates a knowledge-based design advisor and ANSYS-based FEA. By incorporating FEA within the design process the system is consistent with the synthesis-analysis-evaluation design methodology, so that a more complete design assistant service is delivered. However, in the developed optimiser, a range of certain initial points (core dimensions) based on the available cores is given to the system, and the optimisation process is then based on sweeping through all these cores with massive level of iteration, selecting all possible alternatives, narrowing down the selected points according to the objective function (e.g. cost, volume, loss...), returning the optimum solution. This is a relatively time consuming process that does not give any particular insight into the final decision and why this is the optimum design point.

The main limitation with this expert system approach is that the design process is essentially dependant on the correct selection of the magnetic component core as a starting point. Typically, core selection is treated as an expert input variable, and requires considerable expertise and background to make a good selection as a starting point.

Hence this thesis now explores an alternative analytical way to make this initial core selection, by revisiting the fundamental mathematics that underpin the design of a magnetic component. These mathematics are then reformulated into a form that allows a definitive core selection to be made for a particular design requirement, instead of beginning with initial heuristic design assumptions as is conventionally done.
Chapter 4

A NEW STRATEGY FOR MAGNETIC CORE SELECTION AND INDUCTOR DESIGN

Chapter 3 has presented the development of an expert system for optimisation of the design of a magnetic component. The system is based on the integration of a knowledge-based design advisor and ANSYS-based FEA to provide an optimum solution. However, since the optimiser does not give any particular insight about its output, the validation of a design outcome can only be done either by building the prototype or by more advanced analytical analysis.

There are many design methodologies published in the literature so far which present design guidelines to design an inductive component based on fundamental concepts [54, 55, 79, 122, 162]. Generally, these guidelines propose a sequential series of steps compromising

- Core selection
- Conductor selection
- Loss evaluation
- Multiple iterations of the previous steps to achieve a final optimised design.

The first part of this chapter reviews the theoretical principles on which this magnetic component design process is based, and then proceeds to illustrate how these principles can be used to proceed with a design based on the above series of steps. From this review, it is shown how the core selection is typically made by determining an appropriate Area-Product \( (A_p) \) parameter for a particular application, which in turn is
Chapter 4  A New Strategy for Magnetic Core Selection and Inductor Design

usually defined by experience using heuristic values for the two underlying parameters of *winding utilisation factor* \((k_u)\) and *core to dc copper loss ratio* \((\gamma)\). This identification underpins the fundamental difficulty when designing magnetic components for higher power intermediate frequency applications – no clear guidelines have been found in the literature as to how to make appropriate initial assumptions for these two parameters in this application region, and values proposed in the literature for higher frequency lower power applications have been found in this research to be inappropriate for the intermediate frequency region. Proceeding with a design using such initial values can lead to a sub-optimum or even an unfeasible design, and often requires many design iterations using heuristic revisions of these parameters to achieve a satisfactory outcome.

The chapter\(^1\) then proceeds to present a new 3D graphical deterministic strategy for selecting an appropriate core for any particular application, which does not require heuristically defined values for \(k_u\) and \(\gamma\). The approach graphically identifies all theoretically possible \(A_P\) values for a given inductor specification across a broad range of \(k_u\) and \(\gamma\) values, and then constrains these values by considering physical limits such as the allowable current density, achievable inductance values, and distribution between the core and winding losses. The outcome is the identification of a greatly reduced subspace of \(A_P\) values that are physically feasible for the inductor design. This allows for a considered selection of cores that suit the design to be made with confidence that the final inductor design is physically realisable, and will meet the required design and performance specifications. Furthermore, once the core selection is made in this way, the inductor design is fully completed at the same time, without requiring the further iteration cycles that are commonly required by more conventional design approach.

4.1. Fundamental Equations for Magnetic Component Design [55]

This section is summarised from the comprehensive reference on theoretical principles of high frequency magnetic components design, published recently by Hurley *et al.* [55].

---

\(^1\)This material was first published as:

4.1.1. Inductance

Inductor design starts with the fundamental specification of inductance which defines the relationship between number of turns $N$ and total reluctance as:

$$L_m = \frac{N^2}{\mathcal{R}_{\text{total}}}$$  \hspace{1cm} (4.1)

Inductance can be also expressed in terms of permeability of the core and free space and the core dimensions as

$$L_m = \frac{\mu_{\text{eff}} \mu_0 A_c N^2}{l_c}$$  \hspace{1cm} (4.2)

where $\mu_{\text{eff}}$ is the effective permeability of the core including gap, $\mu_0 = 4\pi \times 10^{-7} \, \text{H/m}$ is free space permeability, $A_c$ is the core cross-sectional area and $l_c$ is the mean magnetic length of the core.

4.1.2. Maximum Flux Density

The relationship between the magnetic field intensity ($H_{\text{max}}$) and the current in the core with $N$ turns can be expressed by applying Ampere’s law

$$H_{\text{max}} = \frac{NI_{\text{max}}}{l_c}$$  \hspace{1cm} (4.3)

where $I_{\text{max}}$ is the peak value of the current.

The rms value of the current ($I_{\text{rms}}$) is related to the $I_{\text{max}}$ by $K_i$, which is the current waveform crest factor viz:

$$I_{\text{rms}} = K_i I_{\text{max}}$$  \hspace{1cm} (4.4)

The maximum flux density and $H_{\text{max}}$ are then related by:

$$B_{\text{max}} = \mu_{\text{eff}} \mu_0 H_{\text{max}} = \frac{\mu_{\text{eff}} \mu_0 NI_{\text{max}}}{l_c}$$  \hspace{1cm} (4.5)

where $B_{\text{max}}$ has a limit, which is determined by the saturation flux density ($B_{\text{sat}}$) of the core material. From (4.3) and (4.5), $I_{\text{max}}$ can be expressed in terms of $B_{\text{max}}$ as

$$I_{\text{max}} = \frac{B_{\text{max}} l_c}{\mu_{\text{eff}} \mu_0 N}$$  \hspace{1cm} (4.6)
4.1.3. Winding Loss

The ohmic loss or dc loss within a winding can be calculated using:

\[ P_{\text{winding,dc}} = \rho_w \frac{l_w}{A_w} I_{\text{rms}}^2 \]  
\hspace{2cm} (4.7)

where \( \rho_w \) is conductor resistivity, \( l_w \) is conductor length and \( A_w \) is conductor cross sectional area. The conductor length is given by:

\[ l_w = N \times \text{MLT} \]  
\hspace{2cm} (4.8)

where MLT is the Mean Length of each winding Turn.

\( P_{\text{cu}} \) can be revised by substituting (4.8) in (4.7), to give:

\[ P_{\text{winding,dc}} = \rho_w \frac{N \text{MLT}}{A_w} I_{\text{rms}}^2 \]  
\hspace{2cm} (4.9)

4.1.4. Core Loss

The core loss per unit volume using Steinmetz equation is given by:

\[ P_{\text{core}} = K_c f^\alpha \left( \frac{\Delta B}{2} \right)^\beta \]  
\hspace{2cm} (4.10)

where \( \Delta B \) is the peak-to-peak flux density ripple. \( K_c, \alpha \) and \( \beta \) are Steinmetz equation material parameters.

Finally, Hurley et al. [55] introduced parameter \( \gamma \) as the ratio of the core loss to the dc winding loss, given by

\[ P_{\text{core}} = \gamma P_{\text{winding,dc}} \]  
\hspace{2cm} (4.11)

4.1.5. Thermal Equation

The total inductor loss, which is the summation of the core and winding losses, usually needs to be dissipated through the surface area of the core and windings by convection. This heat transfer can be expressed by:

\[ Q = h_c A_c \Delta T \]  
\hspace{2cm} (4.12)
where \( Q \) is the total power loss (\( P_{\text{core}} + P_{\text{winding}} \)), \( h_c \) characterizes the heat transfer coefficient, \( A_t \) is the surface area of the magnetic component and \( \Delta T \) is the temperature rise. \( \Delta T \) relates to \( Q \) by the thermal resistance \( (R_\theta) \) as shown by:

\[
\Delta T = R_\theta Q = \frac{1}{h_c A_t} Q
\]  

(4.13)

This thermal resistance can typically be found in the core manufacturer’s data sheet.

### 4.1.6. Current Density

The current density in the winding is defined as the ratio of the rms value of the current \( (I_{\text{rms}}) \) to the total conductor cross-sectional area, as defined by:

\[
J_0 = \frac{I_{\text{rms}}}{A_w}
\]  

(4.14)

### 4.2. Generic Magnetic Component Design Procedure

#### 4.2.1. Core Selection

There are a number of factors determining the physical core selection including:

- The amount of stored energy in the inductor (power handling capability)
- The maximum flux density in the core which therefore requires sufficient cross-sectional area to be provided for a given required flux.
- The current rating of the inductor, which requires sufficient window space to fit the required winding conductors
- The maximum allowable temperature rise in the inductor

The first step in the design process is to select an appropriate core size that meets the inductor specifications. This selection can be done through an iterative process which is based on some initial assumptions, as follows:

1) **Calculation of \( A_p \)**

A major challenge for the design process is to select an appropriately sized core that suits the inductor requirements. An initial core size selection is often made using the core
area parameter $A_p$ [55, 122, 163]. $A_p$ is the product of the core window area ($W_a$) and core cross-sectional area ($A_c$), and specifies the energy handling capability of the core. It is defined by [122].

$$A_p = W_a \times A_c = \frac{2(Energy)(10^4)}{B_m k_u} \quad (4.15)$$

Figure 4.1 illustrates the $A_p$ definition by showing the winding window area and core cross-sectional area for a UI core. $W_a$ is connected to the current rating capability and $A_c$ is linked to the maximum allowed magnetic flux density of the core [164]. An appropriate value of $A_p$ should meet both the mechanical (physical) and electromagnetic requirements of the magnetic component within a given design simultaneously [164]. However, inductor specifications alone are generally insufficient to allow for a precise determination of the required $A_p$ and additional assumed parameters are typically required to satisfy both electromagnetic and core window area criteria [164].

Two common parameters are the Window Utilisation Factor (WUF $- k_u$), i.e. the ratio of the total conduction area ($W_c$) to the total winding window area of the core ($W_a$) [54, 55, 122], and the Core to DC Copper Loss Ratio (CCLR $- \gamma$) [55]. Equations (4.16) and (4.17) define these two parameters as [55]:

Figure 4.1. Area Product ($A_p$) illustration.
Chapter 4  
A New Strategy for Magnetic Core Selection and Inductor Design

\[ \text{WUF: } k_u = \frac{W_c}{W_a} = \frac{N A_w}{W_a} \quad (4.16) \]

\[ \text{CCLR: } \gamma = \frac{P_{\text{core}}}{P_{\text{winding,dc}}} \quad (4.17) \]

However, as noted above, the initial values that are usually chosen for these parameters relate primarily to lower power higher frequency (HF) magnetic components [54, 55], and can result in physically un-constructable designs if the same values are unilaterally applied to higher power intermediate frequency applications.

For example, a typical value of \( k_u \) is proposed to vary from 0.2 to 0.8 depending on how tightly the coil is wound, bobbin factor, the wire size and type of insulation [54, 55, 122]. Hence for a low power high frequency inductor, an initial value of \( k_u \) that is often assumed is 0.4 [54, 122, 162]. However, this research has found that a more appropriate value for a higher power inductor is typically around 0.1. Similarly, while a typical initial assumed value for \( \gamma \) is 2 [55], a more realistic value for a higher power inductor has been found to be around 5. Appropriate initial assumptions for these two parameters for higher power magnetic components are even less clear if multi-strand wire or Litz wire is used to reduce winding losses due to skin effect [58]. Note however that the initial value of \( \gamma \) in an inductor with negligible flux ripple can be taken as 0 [55].

From [55], \( A_p \) can be expressed in terms of WUF and CCLR as:

\[ A_p = \left[ \frac{\sqrt{1 + \gamma K_t L I_{\text{max}}^2}}{B_{\text{max}} K_t \sqrt{k_u \Delta T}} \right]^\frac{8}{7} \quad (4.18) \]

where \( K_t \) is the current waveform factor, \( L \) is the desired inductance, \( I_{\text{max}} \) is the inductor maximum current, \( B_{\text{max}} \) is the maximum allowable flux density for the core material to be used, \( K_t \) is a dimensional constant (\( K_t = 48.2 \times 10^3 \)), and \( \Delta T \) is the maximum allowed temperature rise.

Once the appropriate core size is selected from initial assumed values for \( k_u \) and \( \gamma \), the dimensional parameters of the core can be obtained from the manufacturer datasheet and the magnetic component design can continue as outlined in [55].
2) Calculate $\mu_{opt}$

For a given core with a known window winding area and mean turn length, there is an optimum value of $\mu_{eff}$, effective permeability, which is a balancing point between maximum flux density ($B_{sat}$) and maximum allowable heat dissipation ($P_D$). Equation (4.19) shows the expression to calculate this optimum value, viz

$$\mu_{opt} = \frac{B_{sat} l_c K_i}{\mu_0 \sqrt{\frac{P_{cu, max} N A_w}{\rho_w MLT}}} = \frac{B_{sat} l_c K_i}{\mu_0 \sqrt{\frac{P_{cu, max} k_u W_a}{\rho_w MLT}}}$$

(4.19)

where $K_i$ is the current waveform factor, $P_{cu, max}$ is the maximum copper loss, $A_w$ is the conductor cross-section area, $\rho_w$ is the conductor resistivity and $MLT$ is the mean length of a turn.

4.2.2. Winding Design

The next step after core selection is winding design which requires the following steps:

1) Maximum gap length

The maximum gap length can be found using [55]:

$$g_{max} \approx \frac{l_c}{\mu_{min}} = \frac{l_c}{\mu_{opt}}$$

(4.20)

2) Calculation of number of turns

Once the correct gap length is chosen, the number of turns in the winding, $N$, can be calculated using:

$$N = \sqrt{\frac{L}{A_L}}$$

(4.21)

where $A_L$ is the corresponding value of inductance per turn which is normally supplied by the manufacturer for different gap length [55].

3) Calculation of Current density

The current density ($J_0$) within the winding needs to meet the maximum allowable temperature rise ($\Delta T$) and can be expressed in terms of $k_u$, $\gamma$, $\Delta T$ and $A_L$ as [55]
\[ J_0 = K_t \frac{\sqrt{\Delta T}}{\sqrt{k_u(1 + \gamma)^n} \sqrt{A_p}} \]  
(4.22)

**4) Conductor size**

The required conductor cross-sectional area can be calculated using

\[ A_w = \frac{I_{rms}}{J_0} \]  
(4.23)

where \( I_{rms} \) is the rms value of inductor current.

### 4.2.3. Loss Calculation

The final step in the design process is the calculation of core and winding losses taking into account the high frequency effect on the winding loss. The required equations to calculate these losses have been explained earlier in this chapter.

### 4.2.4. Summary of Generic Magnetic Component Design:

Sections 4.1 and 4.2 have presented an overview of the generic design steps of a high frequency inductor. The mathematical equations that underpin each of the design steps have been provided and the whole design process has been illustrated as a flowchart. Figure 4.2 illustrates the flow chart of this overall generic design methodology.

From this overview, it has been shown that selecting an appropriately sized core to suit the inductor requirements is an iterative process that starts with two initial parameters, \( k_u \) and \( \gamma \). The assumed values of these parameters are highly dependent on the given application and starting with inappropriate values may lead to a non-satisfactory outcome (an unfeasible design) and require many further iterations to achieve a useful outcome.

### 4.3. Improved Core Selection and Inductor Design Methodology

In this section, an improved methodology is now presented for selecting the core for higher power inductors using a more deterministic approach to select appropriate values for \( k_u \) and \( \gamma \) and thus to determine the required core \( A_p \). Figure 4.3 shows the
overview of this improved core selection process and consequential inductor design methodology, which was coded using MATLAB m-file script with graphical results displayed for each step of the analysis algorithm execution. Details of the MATLAB m-file script are presented in Appendix B.
Inductor Design Specifications

Calculate all the possible $A_p$ values (sweeping $k_u$ and $\gamma$ over a wide range)

Current Density Constraint
($1 \text{ A/mm}^2 \leq J \leq 2 \text{ A/mm}^2$)

Inductance Constraint
($L_{\text{calc}} \leq L_{\text{target}} \pm 3\%$)

Loss Constraint
($P_{\text{core}}, P_{\text{winding}}$)

Final $A_p$

Feasible Core selected And Design Completed

Figure 4.3. Flow chart of improved core selection and inductor design methodology
4.3.1. DAB Specification

The starting point for any inductor is its design specification, which defines the desired inductance \( L \), the inductor maximum current \( I_{\text{max}} \), the maximum allowable flux density \( B_{\text{max}} \) for the core material to be used, and the maximum allowed temperature rise \( \Delta T \). These parameters are primarily determined by the converter application within which the inductor is to be used. For this chapter, the series inductance of the Dual Active Bridge (DAB) converter topology [2] shown in Figure 4.4, with the design performance requirements as detailed in Table 4.1, is used to confirm the core selection and inductor design process. Comparing the previously elucidated topologies in section 1.2, VSI and DAB, in terms of their current waveforms and associated harmonics, clearly the DAB topology is the more challenging one. The excitation waveform is a square wave, full of harmonics producing full magnitude harmonic at the fundamental frequency which is

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>( P )</td>
<td>2.5 kW</td>
</tr>
<tr>
<td>High voltage side</td>
<td>( v_{\text{dcHV}} )</td>
<td>650 V</td>
</tr>
<tr>
<td>Low voltage side</td>
<td>( v_{\text{dcLV}} )</td>
<td>50 V</td>
</tr>
<tr>
<td>Battery current</td>
<td>( i_{\text{dcLV}} )</td>
<td>50 A</td>
</tr>
<tr>
<td>Switching Frequency</td>
<td>( f )</td>
<td>20 kHz</td>
</tr>
<tr>
<td>Inductance</td>
<td>( L )</td>
<td>550 ( \mu )H</td>
</tr>
<tr>
<td>Maximum inductor current</td>
<td>( I_{\text{max}} )</td>
<td>8.5 A peak</td>
</tr>
<tr>
<td>Maximum flux density</td>
<td>( B_{\text{max}} )</td>
<td>0.24T</td>
</tr>
<tr>
<td>Temperature Rise</td>
<td>( \Delta T )</td>
<td>20°C</td>
</tr>
<tr>
<td>Ambient temperature</td>
<td>( T_{\text{amb}} )</td>
<td>50°C</td>
</tr>
</tbody>
</table>

Figure 4.4. Vanadium flow battery charger comprising three-phase voltage-source converter and 2.5-kW dual-active-bridge bidirectional DC-DC converter.
right at the switching range. This leads to more losses and consequently has more potential for poor inductor design. Since a fundamental presumption for this thesis is that the proposed core selection method can be used for any topology, and since DAB is the more challenging topology, it is chosen as the exemplar reference for the procedure.

4.3.2. General $A_P$ calculation:

Once the basic inductor specification is established, $A_P$ can be calculated using (55):

$$A_P = W_a \times A_c = \frac{\sqrt{1 + \gamma K_i L_i L_{\text{max}}^2}}{B_{\text{max}} K_t \sqrt{k_u \Delta T}}$$  \hspace{1cm} (4.24)

where $K_t$ is a dimensional constant ($K_t = 48.2 \times 10^3$), $K_i$ is the current waveform crest factor ($I_{\text{rms}} = K_i L_{\text{max}}$) and $k_u$ and $\gamma$ are the two design parameters WUF and CCLR identified earlier. Values for all possible $A_P$ are calculated as $k_u$ and $\gamma$ are swept over a wide range with fine grain increments (i.e. $0.1 < k_u < 0.8$ at 0.01 intervals and $0 < \gamma < 40$ at 0.1 increments), to ensure that a practical and realistic value for $A_P$ has been included within the investigation space. Figure 4.5 shows the 3D surface for $A_P$ created.

![Figure 4.5. Maximum possible $A_P$ values for DAB inductor.](image-url)
by this process, which includes values that may or may not be viable, depending on the physical constraints that will now be considered.

### 4.3.3. Current density constraint:

The first physical constraint is to limit the winding current density \( J_0 \) between 1 A/mm\(^2\) and 2 A/mm\(^2\), i.e. a conservative range based on prior experience that suits larger inductors, to minimize the eddy current contribution to loss profile [58]. The current density is defined by the function of \( k_u, \gamma, \Delta T \) and \( A_P \) as shown in [55]

\[
1 \leq \frac{A}{mm^2} \leq J_0(k_u, \gamma) = K_t \frac{\sqrt{\Delta T}}{\sqrt{k_u(1 + \gamma)^s}} \leq 2 \frac{A}{mm^2} \quad (4.25)
\]

Constraining the current density to within the above limits, \( 1 \leq \frac{A}{mm^2} \leq J_0 \leq 2 \frac{A}{mm^2} \) reduces the allowable values of \( k_u \) and \( \gamma \) to \( k_{u,sel} \) and \( \gamma_{sel} \), which in turn constrains the allowable values of \( A_P \) as defined by:

\[
A_P(k_{u,sel}, \gamma_{sel}) = \left[ \frac{\sqrt{1 + \gamma_{sel}}K_t LL_{\max}}{B_{\max}K_t\sqrt{k_{u,sel}\Delta T}} \right]^{\frac{8}{7}} \quad (4.26)
\]

Figure 4.6 shows the 3D representation of this constraint, where only \( A_P \) values within the red area of the previous overall possible \( A_P \) surface can now be considered for the inductor core selection.

### 4.3.4. Inductance Constraint:

With the \( A_P \) values constrained by the allowable current density, the feasible inductance that can be achieved for each value of \( A_P \) now further limits the viable \( A_P \) range. Calculation of this inductance requires a number of steps:

Firstly, the conductor cross section \( A_w \) is calculated with respect to the limited range of \( k_{u,sel} \) and \( \gamma_{sel} \) using:

\[
A_w(k_{u,sel}, \gamma_{sel}) = \frac{I_{rms}}{J_0(k_{u,sel}, \gamma_{sel})} \quad (4.27)
\]

All major parameters and physical dimensions of the core, such as \( W_a, A_c \), the core volume \( (V_c) \), the mean length of a turn (MLT), magnetic path length of the core \( (l_c) \), winding volume \( (V_w) \) and total surface area \( (A_t) \) can be determined from its \( A_P \) values.
using dimensional analysis as described in [55, 78]. For this chapter, this was done by defining a range of $W_a$ to $A_c$ ratios to decrease the number of unknown quantities. By comparing several core options, it was found that this ratio typically varies between 2 and 4. Hence the corresponding $W_a$, $A_c$, $V_c$ and $l_c$ for every $A_p$ value can be obtained using the following equations:

$$A_p = W_a \times A_c, \quad 2 \leq W_a/A_c = x \leq 4 \quad \rightarrow \quad W_a = x \times A_c$$ (4.28)

$$\rightarrow A_p = x \times A_c^2 \quad \rightarrow \quad A_c(k_{u,sel}, y_{sel}) = \sqrt{\frac{A_p(k_{u,sel}, y_{sel})}{x \in [2,4]}}$$ (4.29)

$$\rightarrow W_a(k_{u,sel}, y_{sel}) = x \in [2,4] \times \sqrt{\frac{A_p(k_{u,sel}, y_{sel})}{x \in [2,4]}}$$ (4.30)

From [55], the relationship between $V_c$ and $A_p$ is given by (4.31) and the $l_c$ can be obtained using:

$$V_c(k_{u,sel}, y_{sel}) = 5.6 \ A_p(k_{u,sel}, y_{sel})^{3/4} = l_c \times A_c$$ (4.31)
→ \( l_c(k_{u\text{,sel}}, \gamma_{\text{set}}) = \frac{V_c}{A_c} = \frac{5.6 A_p(k_{u\text{,sel}}, \gamma_{\text{set}})^{3/4}}{\sqrt{A_p(k_{u\text{,sel}}, \gamma_{\text{set}})}} \quad (4.32) \)

Next, the maximum possible number of turns can be calculated using:

\[ N(k_{u\text{,set}}, \gamma_{\text{set}}) = \frac{k_u W_0(k_{u\text{,set}}, \gamma_{\text{set}})}{A_w(k_{u\text{,set}}, \gamma_{\text{set}})} \quad (4.33) \]

Then, knowing the number of turns and by sweeping the air-gap length from 0 to 3 mm at 0.01 mm increments, a range of possible inductance values can be calculated using:

\[ L(k_{u\text{,set}}, \gamma_{\text{set}}) = \frac{N^2(k_{u\text{,set}}, \gamma_{\text{set}})}{\gamma_\text{core} + \gamma_\text{gap}} \]

\[ = \frac{N^2(k_{u\text{,set}}, \gamma_{\text{set}})}{\mu_0 H_r A_c(k_{u\text{,set}}, \gamma_{\text{set}}) + \mu_0 A_c(k_{u\text{,set}}, \gamma_{\text{set}}) l_{\text{gap}}} \quad (4.34) \]

Finally, the allowable range of \( k_{u\text{,set}} \) and \( \gamma_{\text{set}} \) can be limited by constraining the air-gap variation and the number of turns to keep the obtained inductance within an acceptable range of ± 3% of the design target, i.e.

\[ L(k_{u\text{,set}}, \gamma_{\text{set}}) \leq L \pm 3\% \quad (4.35) \]

The red surface lines in Figure 4.7 show the range of inductances that can be achieved using (4.35), while the blue asterisk points on the figure identify where the inductances are within the limits defined by (4.35). Figure 4.8 shows the consequence of this constraint on \( A_p \) values, where the green sub-surface identifies the only region of \( A_p, k_u \) and \( \gamma \) that can achieve the required inductances within the current density constraint defined by (4.25)
Figure 4.7. Possible L values (red) and selected L values (blue) for DAB inductor

Figure 4.8. Possible $A_P$ values (twisted surface), selected $A_P$ values constrained by current density (red) and Inductance (green) for DAB inductor
4.3.5. Loss Constraint:

The final constraint considered is to determine where the core and winding losses crossover on the constrained (green) \(A_p\) sub-surface shown in Figure 4.8. Determination of this crossover point requires a number of steps.

First for each \(A_p\) value, the core loss can be calculated using [55]

\[
P_{\text{core}} = 5.6 A_p (k_{u, \text{set}}, \gamma_{\text{set}})^{3/4} \times K_c \times f^a \times B_{\text{max}}^b
\]  

(4.36)

In order to proceed with the calculation of the dc winding loss for each \(A_p\) value, the corresponding \(MLT\) values for the refined range of \(k_u\) and \(\gamma\) needs to be obtained. This can be done by using

\[
MLT(k_{u, \text{set}}, \gamma_{\text{set}}) = 4 \times \sqrt{A_c(k_{u, \text{set}}, \gamma_{\text{set}})}
\]

\[
= 4 \times \left[ A_p(k_{u, \text{set}}, \gamma_{\text{set}}) \right]^{1/4}
\]

(4.37)

Then, knowing the \(MLT(k_{u, \text{set}}, \gamma_{\text{set}})\), the required conductor length \((l_w)\) for each \(A_p\) value can be found using

\[
l_w = N \times MLT(k_{u, \text{set}}, \gamma_{\text{set}})
\]  

(4.38)

Now, the dc winding loss for each \(A_p\) value can be calculated using

\[
P_{\text{winding, dc}} = \rho_w \frac{N \times MLT(k_{u, \text{set}}, \gamma_{\text{set}})}{N_w} l_{\text{rms}}^2
\]

(4.39)

\[
l_{\text{rms}} = J_0(k_{u, \text{set}}, \gamma_{\text{set}}) A_w(k_{u, \text{set}}, \gamma_{\text{set}})
\]

(4.40)

Substituting (4.39) into (4.40) gives:

\[
P_{\text{winding, dc}} = \rho_w \frac{N^2 \times MLT(k_{u, \text{set}}, \gamma_{\text{set}})}{N A_w} (J_0(k_{u, \text{set}}, \gamma_{\text{set}}) A_w(k_{u, \text{set}}, \gamma_{\text{set}}))^2
\]

\[
= \rho_w N A_w MLT J_0(k_{u, \text{set}}, \gamma_{\text{set}})^2
\]

\[
= \rho_w k_{u, \text{set}} W_a MLT J_0(k_{u, \text{set}}, \gamma_{\text{set}})^2
\]

(4.41)

where the winding volume \((V_w)\) is defined as:

\[
V_w = W_a MLT(k_{u, \text{set}}, \gamma_{\text{set}})
\]  

(4.42)

Combining (4.42) into (4.41) gives
\[ P_{\text{winding,dc}} = \rho_w k_{u,sel} V_w J_0(k_{u,sel}, \gamma_{sel})^2 \]  

Using dimensional analysis [55], \( P_{\text{winding,dc}} \) can be expressed as:

\[ P_{\text{winding,dc}} = 10 A_p (k_{u,sel}, \gamma_{sel})^{3/4} \times \rho_w \times k_{u,sel} \times J_0(k_{u,sel}, \gamma_{sel}) \]  

The total winding loss includes both dc and ac winding losses. The ac winding loss is denoted by \( P_{\text{winding,ac}} \). In this exemplar design, litz-wire is selected for the inductor winding to reduce eddy current effects [9]. By adapting Dowell’s equation to litz-wire as described in Section 2.2, the ac-to-dc winding resistance ratio of the inductor winding with litz wire can be written as [81]:

\[
F_R^{\text{litz}} = \frac{R_{\text{dc}}^{\text{litz}}}{R_{\text{ac}}^{\text{litz}}} = A_{\text{str}} \left( \frac{\sinh(2A_{\text{str}}) + \sin(2A_{\text{str}})}{\cosh(2A_{\text{str}}) - \cos(2A_{\text{str}})} \right) \\
+ \frac{2(k. N_i^2 - 1) \sinh(A_{\text{str}}) - \sin(A_{\text{str}})}{3 \cosh(A_{\text{str}}) + \cos(A_{\text{str}})}
\]  

where \( k \) is the number of strands of each litz bundle and

\[ A_{\text{str}} = \left( \frac{\pi}{4} \right)^{0.75} \frac{d_{\text{str}}}{\delta} \sqrt{\eta} \]  

where \( d_{\text{str}} \) is the litz-wire bare strand diameter.

Finally, the total power loss (\( P_{\text{Total}} \)) of the inductor which is the summation of core and total winding losses can be calculated using

\[ P_{\text{Total}} = P_{\text{core}} + P_{\text{winding,dc}} + P_{\text{winding,ac}} \]  

Figure 4.9 shows the variation in core loss and total winding loss (\( P_{\text{winding,dc}} + P_{\text{winding,ac}} \)) within the constrained value range of \( k_u \) and \( \gamma \) identified by the green surface of Figure 4.8. The intersection between these two losses defines a line where the losses are matched. The point of minimum absolute loss can then be identified on this line as shown in Figure 4.9. Since the two curves are monotonic, there cannot be any local minima and so the solution must be a global optimum. Translating this point back into the matching \( k_u \) and \( \gamma \) values defines the final optimal \( A_p \) value for the core, and thus completes the core selection process as shown in Figure 4.10.
Figure 4.9. Variation in Core and Total winding loss versus the final refined range of $k_u$ and $\gamma$

Figure 4.10. Possible $A_P$ values (twisted surface), selected $A_P$ values constrained by current density (red), Inductance (green) and Loss (blue)
For the exemplar DAB inductor used in this thesis, an appropriate core was determined to require an $A_p$ value of $A_p > 72$, with $k_u = 0.1$ and $\gamma = 4.8$, using the above methodology. Verification of the inductor design and matching experimental results are presented in the next chapter.

4.4. Summary

This chapter has presented a novel 3D graph based methodology for the core selection and design of inductors operating in the higher power and/or lower frequency range. Unlike other approaches, the methodology overcomes the problem of sensitivity of core selection and inductor design to initial value estimates of Window Utilization Factor and Core to DC Copper Loss Ratio. The result significantly improves the inductor design outcome in terms of number of design iterations, size (volume), achievement of target inductance and overall design robustness. The technique is adaptive, does not require any preconceptions of parameter initialization, is applicable to the design of a range of higher power inductors, and also can be used for transformer design with only minor modifications.
Chapter 5
Magnetic Design Verification

High-power bidirectional battery chargers are an essential building block of renewable energy systems, performing the conversion of a primary higher voltage level provided by an incoming AC-DC conversion system into a secondary lower level that suits a battery voltage [25, 33, 165, 166]. Intermediate-frequency switch-mode battery chargers offer substantial advantages over lower frequency systems, because of the savings in volume and weight that can be achieved. The bidirectional dual-active-bridge (DAB) DC-DC converter shown in Figure 5.1 is an attractive solution that also inherently offers the galvanic isolation required on these applications. A DAB converter comprises two single-phase H-bridges connected back-to-back across an AC link formed by an isolating/scaling transformer \( T \) and an intermediate link inductor \( L \). These two magnetic components are crucial for the overall system performance, and a substantial engineering effort is usually required to successfully design and manufacture them because of the large number of interrelated electrical, physical and geometrical parameters that must be considered. These parameters include operating frequency, core material, excitation waveform, temperature, and mechanical dimensions \(^1\).

In Chapter 4 the minimum \( A_p \) value for the core of the exemplar DAB series inductance used in this thesis was determined as \( A_p = 72 \) using the 3D graphical analysis strategy presented in that chapter. An inductor for the DAB was then constructed from this design, using a slightly larger (available) core with \( A_p = 93 \) with the winding design as determined in chapter 4. The airgap of the core was set at 1.58mm (slightly above the theoretical value of 1.1mm) to exactly set the inductance to the required value of 550\( \mu \)H.

\(^1\)Materials in this chapter were first published as:
The DAB transformer was designed using the conventional iterative design process shown in Figure 5.3. The design outcome was then verified using simulation and experimental confirmation in a pre-production converter that was used for a 2.5 kW vanadium flow battery bidirectional charging system.

### 5.1. Bidirectional DAB Battery Charger System

The dual-active-bridge bidirectional converter shown in Figure 5.1 comprises two back-to-back connected single-phase voltage-source converters, with individual phase legs formed by complementary switch pairs $(S_1, S_2)$ and $(S_3, S_4)$ for the primary high-side H-bridge, and $(S_5, S_6)$ and $(S_7, S_8)$ for the low-side secondary H-bridge. The intermediate-frequency switching transformer $T$ provides galvanic isolation between the two bridges and scales the primary voltage so as to achieve a lower secondary side voltage. The LV dc-link capacitor $C_L$ provides the bulk energy storage on the battery side, while the HV dc-link capacitor $C_{hv}$ establishes a high frequency decoupled link to the AC-DC conversion stage, which is typically a three-phase voltage-source converter.

The DAB operates by using the two single-phase converters to synthesise a pair of square waveforms $v_{ab}$ and $v_{cd}$, which define the voltage $v_L = v_{ab} - n_v v_{cd}$ applied across the inductor $L$ positioned between the two H-bridges. By varying the phase displacement $\phi$ of these two square waveforms $v_{ab}$ and $v_{cd}$, the resulting inductor voltage $v_L$ is varied, which controls the energy flow between the grid-connected AC-DC converter and the vanadium flow battery.
Figure 5.2 shows the modulation strategy used for this system. Figure 5.2(a) illustrates that for the HV bridge, the two modulation commands $m_a$ and $m_b$ are compared against a triangular carrier to produce the switch logic gate signals $s_a$ and $s_b$ (and also their complementary signals $s_a'$ and $s_b'$). These four gate signals are applied to semiconductor switches $S_i$ through to $S_4$, resulting in the switched phase leg to phase leg output voltage $v_{ab}$. Note also the small phase displacement $\theta$ which is inserted between the switching of the individual H-bridge phase legs to reduce the detrimental effects of practical parasitic components on the converter switching process, as discussed later in this chapter. Similarly, as shown in Figure 5.2(b), the two LV bridge modulation commands $m_c$ and
$m_d$ are compared against the same triangular carrier to produce the switch logic gate signals $s_c$ and $s_d$ (and their complementary signals $s_c'$ and $s_d'$). These signals are applied to power switches $S_5$ through to $S_8$, resulting in the switched output voltage $v_{cd}$. Once again, the same small phase displacement $\theta$ is introduced between the switching of the individual H-bridge phase legs to reduce the effect of parasitic components. Figure 5.2(c) shows the resulting voltage across and current through the inductor $L$, which primarily depend on the phase displacement $\phi$ between the two H-bridge voltage outputs. The simulation results of the topology shown in Figure 5.1 operating under the modulation scheme shown in Figure 5.2 and for the system parameters provided in set the target design specifications for the transformer $T$ and inductor $L$.

## 5.2. Magnetic Component Design

Figure 5.3 shows the (generic) design flowchart used for developing the transformer $T$. The parameters in Table 5.1, which are obtained from the VFBC converter system analysis, define the target specifications for the magnetic components including transformer turns ratio, inductance value, core cross sectional area and conductor type.

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>$P$</td>
<td>2.5 kW</td>
</tr>
<tr>
<td>High voltage side</td>
<td>$v_{dchv}$</td>
<td>650 V</td>
</tr>
<tr>
<td>Low voltage side</td>
<td>$v_{dclv}$</td>
<td>50 V</td>
</tr>
<tr>
<td>Battery current</td>
<td>$i_{dclv}$</td>
<td>50 A</td>
</tr>
<tr>
<td>Carrier Frequency</td>
<td>$f_c$</td>
<td>20 kHz</td>
</tr>
<tr>
<td>Transformer turns ratio</td>
<td>$n$</td>
<td>10</td>
</tr>
<tr>
<td>Inductance</td>
<td>$L$</td>
<td>550 $\mu$H</td>
</tr>
<tr>
<td>Inductor current</td>
<td>$i_L$</td>
<td>8.5 A peak</td>
</tr>
<tr>
<td>Maximum temperature</td>
<td>$T_{max}$</td>
<td>70°C</td>
</tr>
<tr>
<td>Temperature Rise</td>
<td>$\Delta T$</td>
<td>20°C</td>
</tr>
<tr>
<td>Ambient temperature</td>
<td>$T_{amb}$</td>
<td>50°C</td>
</tr>
</tbody>
</table>

Table 5.1. Converter System Parameters and Magnetics Target Specifications
Figure 5.3. Flowchart applied for designing the transformer T.
Using these specifications, the transformer design process begins by selecting an appropriate core material based on the operating switching frequency and by choosing initial core geometry. Then transformer’s primary and secondary winding turns ($N_{pri}$ and $N_{sec}$) are calculated and the maximum flux density $B_{max}$ is found based on the primary side voltage $V_{pri}$ and core cross sectional area. The value for $B_{max}$ is then compared against $B_{sat}$ and if $B_{max} > B_{sat}$ the design needs to be modified to avoid reaching the saturation threshold (accounting for a safety margin of 10% to 20%). This is initially done for a single core stack. However, if the calculated peak flux density $B_{max}$ still exceeds the saturation flux density $B_{sat}$ for a realistic number of turns, then the number of stacked cores (i.e. parallel cores) must be increased to maintain the flux below its saturation threshold.

The conductor type is chosen based on the ac resistance value at the operating switching frequency, taking into account second-order effects such as skin depth and proximity. Typically, the conductor current density needs to be below a given threshold (selected as 2 $\text{A/mm}^2$ for this work) to minimize skin effect. Finally, the core and winding losses and consequential temperature rise are assessed, and if the thermal design objectives are not met, the iterative design process is then repeated until a satisfactory outcome is achieved.

For this battery charger application, the transformer was designed using two stacks of UI93/104/20 ferrite core (details given in Appendix C) as shown in Figure 5.4. Figure 5.4(b) illustrates the winding arrangement on each leg, made up of two individual layers of copper foils that sandwich the HV winding, which in turn comprises two individual layers of 10-turns of litz-wire.

The inductor core was selected as a single stack of UI93/104/20 ferrite core as shown in Figure 5.5, which has an $A_p$ value of 93. This is about 25-30% in excess of the minimum value of 72 as calculated in Chapter 4, which is consistent with commonly recommended design margins for core selection [55].

Each leg winding comprises four layers of 5-turns of litz-wire for each layer, which are intentionally kept away from the air gap by using a spacer wall on the bobbin. This reduces the winding temperature rise because of fringing effect. For both the transformer and inductor design, Kapton tape was used as primary insulator between layers.
Table 5.2 shows the measured parameters of the transformer and inductor built using these design principles.

5.3. Loss Estimation

The main sources of loss in the transformer are the core material and the conductors (both LV and HV windings). For the transformer, the losses are estimated as follows.

5.3.1. Core Loss

Most approaches reported in the literature estimate transformer core loss under non-sinusoidal excitation using Steinmetz coefficients [44, 48]. The method of [46], which uses a modified version of the original Steinmetz approach to account for the
peak-to-peak flux density instead of instantaneous flux density, is used in this design process.

### 5.3.2. Winding Loss

Winding loss depends on both skin and proximity effects (eddy current effects), which cause the current density to be non-uniform in the conductor cross-sections at higher frequencies, leading to increased winding losses. Because of the complexity of winding geometries and interactions between conductors in windings, it is difficult to find a general overall analytical solution for the eddy current losses in windings.

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transformer $T$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of turns of primary winding</td>
<td>$N_{pri}$</td>
<td>40</td>
</tr>
<tr>
<td>Number of turns of secondary winding</td>
<td>$N_{sec}$</td>
<td>4</td>
</tr>
<tr>
<td>Magnetising inductance</td>
<td>$L_{mag}$</td>
<td>12.7 mH</td>
</tr>
<tr>
<td>Leakage inductance</td>
<td>$L_{leak}$</td>
<td>30.8 $\mu$H</td>
</tr>
<tr>
<td>Inductor $L$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Inductance @ 20 kHz</td>
<td>$L_{20k}$</td>
<td>542 $\mu$H</td>
</tr>
</tbody>
</table>

Figure 5.5. Inductor $L$: (a) three-dimensional sketch and (b) photograph of the prototype.
Consequently, since each conductor type behaves differently w.r.t. skin and proximity effects, they must be considered individually for calculating winding loss.

1) Transformer LV winding

Copper foil was used as the conductor for the transformer LV winding. Applying Dowell’s equation \[ 58 \] to calculate the ac-to-dc resistance ratio, also known as the skin and proximity effect factor, yields:

\[
F_R^{foil} = \frac{R_{ac}^{foil}}{R_{dc}^{foil}} = A \left[ \frac{\sinh(2A) + \sin(2A)}{\cosh(2A) - \cos(2A)} \right. \\
\left. + \frac{2(N_t^2 - 1) \sinh(A) - \sin(A)}{3} \cosh(A) + \cos(A) \right]
\]  \( (5.1) \)

where \( N_t \) is the number of layers, \( A \) is the ratio of the foil thickness to the skin depth, given by \( A = \sqrt{\eta h / \delta_0} \), with \( h \) being the foil thickness, \( \delta_0 \) being the skin depth, and \( \eta \) being the porosity factor.

2) Transformer HV Winding

Twisted litz conductors are used for the transformer HV winding to reduce eddy current effects \[ 63 \]. By adapting Dowell’s equation to litz-wire \[ 81 \], the ac-to-dc winding resistance ratio of the HV winding can be written as

\[
F_R^{litz} = \frac{R_{ac}^{litz}}{R_{dc}^{litz}} = A_{str} \left[ \frac{\sinh(2A_{str}) + \sin(2A_{str})}{\cosh(2A_{str}) - \cos(2A_{str})} \right. \\
\left. + \frac{2(k \cdot N_t^2 - 1) \sinh(A_{str}) - \sin(A_{str})}{3} \cosh(A_{str}) + \cos(A_{str}) \right]
\]  \( (5.2) \)

where \( k \) is the number of strands of each litz bundle and \( A_{str} = \left( \frac{\pi}{4} \right)^{0.75} \frac{d_{str}}{\delta} \sqrt{\eta} \) where \( d_{str} \) is the litz-wire bare strand diameter.

A similar approach was used to estimate the losses on the inductor. As the focus of this thesis is on the inductor design a detailed calculation of the inductor loss is now shown in the next section.
5.3.3. Inductor Loss Calculation

For the DAB inductor, core loss was calculated using the iGSE method, (2.6), as 14.2 W.

The winding dc loss was calculated as:

\[
R_{dc} = \rho_w \frac{l_w}{A_w} = 1.72 \times 10^{-8} \times \frac{7.06}{3 \times 1.82 \times 10^{-8}} = 0.02 \Omega
\]

\[
P_{winding, dc} = R_{dc}I_{\text{max}}^2 = 0.02 \times 8.5^2 = 1.445 W
\]

Using this approach, the winding ac loss of the designed inductor comprising of litz-wire conductor with 19 strands of 0.355 mm diameter, \( \delta_0 \) using (5.2) was calculated as:

\[
\delta_0 = \frac{1}{\sqrt{\pi \mu_0 \sigma f}} = \frac{0.066}{\sqrt{20000}} = 0.466 mm
\]

\[
A_{str} = \left( \frac{\pi}{4} \right)^{0.75} \frac{d_{str}}{\delta} = \left( \frac{\pi}{4} \right)^{0.75} \frac{0.355}{0.461} = 0.6424
\]

\[
F_R^{\text{litz}} = \frac{R_{ac}}{R_{dc}} = 6.711
\]

The \( F_R^{\text{litz}} \) factor is for one of the winding sets. Since a split configuration has been used for this inductor design, the total winding ac loss is given by:

\[
P_{winding, ac} = 2F_R^{\text{litz}} \times P_{winding, dc} = 2 \times 6.711 \times 1.445 W = 19.4 W
\]

**Error! Reference source not found.** summarises the calculated dc and ac resistance values for the designed inductor using this methodology.

Summing the calculated core and resistance losses gives a total theoretical power loss for the inductor of 14.2W + 19.4 W = 33.6 W. This loss was then validated experimentally by measuring the voltage and current waveforms across the inductor.

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>dc winding Resistance</td>
<td>( R_{dc} )</td>
<td>0.02 ( \Omega )</td>
</tr>
<tr>
<td>Ratio of ac to dc resistance for litz wire</td>
<td>( F_R^{\text{litz}} )</td>
<td>6.711</td>
</tr>
<tr>
<td>Calculated ac winding Resistance</td>
<td>( R_{ac,cal} )</td>
<td>0.268 ( \Omega )</td>
</tr>
</tbody>
</table>
(waveforms shown in Figure 5.9), separating these waveforms into harmonic components using FFT analysis, and summing the power loss at each harmonic frequency to give a total power loss of \( P_i = \sum V_i I_i \cos \theta_i = 33 \text{W} \). Table 5.4 compares the calculated and measured losses determined in this way, where it can be seen that they match very closely.

Equation (4.13) defines the maximum allowable power loss that will ensure the inductor temperature rise remains within the design specification, linked through the heat transfer coefficient \( h_c \) and the inductor surface area \( A_t \). From [55], the surface area can be approximately determined from the core \( A_p \) using \( A_t = k_a \sqrt{A_p} \) where \( k_a \) typically has a value of 40. The thermal resistance of the inductor is then determined by \( R_\theta = \frac{1}{h_c A_t} \).

For this inductor design, the core \( A_p \) was \( 9.3 \times 10^5 \text{mm}^4 \), which gives a surface area of \( A_t = 0.039 \text{m}^2 \). From [55], an appropriate value of \( h_c \) for a forced convection cooling inverter such as this application, would be in the range \( 10 - 30 \text{W}/\text{°C m}^2 \), and so a value of \( 25 \text{ W}/\text{°C m}^2 \) was used. This gives a thermal resistance of

\[
R_\theta = \frac{1}{h_c A_t} = \frac{1}{25 \times 0.039} = 1.037 \text{°C/W}
\]

Consequently, for the design requirement of an inductor temperature rise of \( 20 \text{°C} \), the anticipated maximum allowable power loss using (4.13) should be no more than

\[
P_{\text{max}} = \frac{\Delta T}{R_\theta} = \frac{20}{1.037} = 19.28 \text{ W}
\]

which as shown in Table 5.4 is significantly less than the actual loss of 33W determined by both calculation and measurement.

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>dc Winding Loss</td>
<td>( P_{\text{winding,dc}} )</td>
<td>1.445 W</td>
</tr>
<tr>
<td>Total ac Winding Loss</td>
<td>( P_{\text{winding,ac}} )</td>
<td>19.4 W</td>
</tr>
<tr>
<td>Core Loss</td>
<td>( P_{\text{core}} )</td>
<td>14.2 W</td>
</tr>
<tr>
<td>Total Loss (Calculated)</td>
<td>( P_{\text{total,cal}} )</td>
<td>33.6 W</td>
</tr>
<tr>
<td>Total Loss (Measured)</td>
<td>( P_{\text{total,meas}} )</td>
<td>33 W</td>
</tr>
<tr>
<td>Maximum Allowable Power Loss</td>
<td>( P_{\text{max}} )</td>
<td>19.28 W</td>
</tr>
</tbody>
</table>
Essentially, this discrepancy is most likely because the equations given in [55] to calculate \((h_c \text{ and } R_\theta)\) are primarily intended for small core dimensions and are not so applicable for bigger core sizes such as the one used in this design. In particular, either the upper limit for \(h_c\) with forced cooling is too low (a value of 43 would be required to match the theoretical and measured losses for this design), or the relatively simplistic relationship between surface area and thermal resistance defined in (5.3) from [55] is not so appropriate for a larger core. Resolving these issues would require more advanced thermodynamic analysis to derive a more accurate relationship between power loss and temperature rise under forced cooling conditions, and is left to future work beyond the scope of this thesis.

5.4. Simulation and Experimental Validation

Table 5.2 provides the key measured data for the physical transformer \(T\) and inductor \(L\) that were constructed based on this design process. These parameters were then incorporated into the converter simulation model to compare the converter performance (under nominal operating conditions) with experimental results measured on the full-scale system.

Figure 5.6 and Figure 5.7 show the transformer primary and secondary voltages for simulation and experiment. From these plots the anticipated turns ratio \(a = 10\) has been clearly achieved. However, there are some discrepancies between simulated and experimentally measured plots because of practical second-order effects that are not taken into account in the simulation model. More specifically, the high frequency oscillations not present in the simulation are caused by the combined effect of parasitic inter-winding capacitances and cabling inductances.
Figure 5.6. Simulation waveforms: (a) transformer primary-side voltage $v_{pri}$ and (b) transformer secondary-side voltage $v_{sec}$.

Figure 5.7. Experimental waveforms: (a) transformer primary-side voltage $v_{pri}$ and (b) transformer secondary-side voltage $v_{sec}$.

Figure 5.8 and Figure 5.9 present simulation and experimental waveforms for the HV and LV switched voltages $v_{ab}$ and $v_{cd}$, and inductor current $i_L$. Again, for these plots, aside from minor mismatches caused by practical second-order effects, there is an excellent agreement between theory and practice. Particularly, the experimental LV switched voltage $v_{cd}$ contains high-frequency oscillations at the beginning of each level change, which quickly damp out after less than 500 ns. These spikes are a result of the inherently large $dI/dt$ through the semiconductor switches and their speed limited diode conduction mechanism. The use of the additional phase shift delay $\theta$ identified in Figure 5.2 was used to reduce the magnitude of these spikes as shown in Figure 5.9(b), since this delay effectively reduces the $dI/dt$ by a factor of two.
Figure 5.8. Simulation waveforms: (a) HV bridge switched voltage $v_{ab}$, (b) LV bridge switched voltage $v_{cd}$ and (c) inductor current $i_L$. 
Figure 5.9. Experimental waveforms: (a) HV bridge switched voltage \( v_{ab} \), (b) LV bridge switched voltage \( v_{cd} \), and (c) inductor current \( i_L \).
5.5. Thermal Management

The core and winding losses produce substantial heat, creating a significant temperature rise for the individual magnetic components. A split winding arrangement for each component was therefore used to reduce the number of layers on each winding and improve their heat extraction.

Figure 5.10(a) illustrates the converter enclosure, showing the relative position of the power stage heatsink, magnetics, box openings, and fans. The unit has three openings located at the front, side and back of the box. Two brushless DC fans were positioned force air through the heatsink fins, and to facilitate airflow across inductor and transformer. This cooling arrangement was determined as part of the overall converter enclosure design, based essentially on prior experience with cooling a converter of this power rating. Hence the precise airflow over the inductor was not quantified, as is often the case with such a system design.

Figure 5.10(b) shows several experimental temperature measurements for the converter running at nominal operating conditions for over 2.5 hours, where it can be seen that temperature rise of the inductor stabilised within the specified requirement of 20°C. This ensured that the absolute temperature of both magnetic components remained below 70°C in an ambient temperature of 50°C, satisfying the requirement of \( T < 75^\circ C \) set for this design because of the ABS plastic used to construct the bobbins.

5.6. Summary

This chapter has presented a combined theoretical/heuristic development approach for the magnetics design for a commercial 2.5-kW bidirectional battery charger. The target parameters of the magnetic components are determined from comprehensive switched simulations of the converter topology, and the detailed design then proceeds to meet these specifications. It has also been shown how the minimum core size for the DAB series inductor using developed 3D graph based method in previous chapter matches quite well against the core size that was selected using the iterative design methodology described in this chapter (allowing for an appropriate margin for the core \( A_p \) value actually used compared to the minimum possible value determined by the 3D graphical analysis). The inductor and transformer design were then verified by experimental measurements and converter operational tests.
Figure 5.10. Experimental thermal results: (a) converter enclosure (top view) indicating measurement points and (b) temperature values.
Chapter 6

DESCRIPTION OF SIMULATION AND EXPERIMENTAL SYSTEMS

This chapter consists of two main sections. The first section introduces the simulation systems that have been developed to analyse the performance of the converter and magnetic components to fine-tune the final design before prototyping. The second section provides descriptions of both the exemplar inductor prototype construction and the experimental system that has been used for the validation of the design outcome as the final stage of the work. This system allows the results of the proposed design method to be evaluated under real experimental conditions.

6.1. Simulation Systems

Simulation is an important tool that is commonly used to validate theoretical developments. In this work the Powersim simulation package (PSIM) [167] and MATLAB software have been used extensively. The PSIM package allows detailed simulation of the converter system and magnetic components modelling to determine their practical values and specifications. To begin the work, a PSIM simulation system is used to provide an initial determination of the magnetic component values and the converter performance. For a more accurate representation of the physical system and to achieve a performance that better matches the anticipated experimental conditions, second-order effects such as semiconductor device voltage-drops, device parasitic capacitances and further optimizations are also integrated into the simulation system.

The MATLAB software package was then used for developing the improved core selection method and the design optimization studies.
6.1.1. PSIM

PSIM is a time-based simulation package that is particularly designed for the analysis of power electronic circuits and their controllers. The developed PSIM simulation system in this work is described in two major sections, viz:

- **Power stage**: including the primary and the secondary H-bridges, magnetic components, voltage and current measurement sensors, battery and supply.
- **Controller stage**: comprising the DSP controller with two sub blocks including analogue conditioning and PWM modulator blocks.

**Power stage**

Figure 6.1 shows the simulation schematic diagram of the power stage for the Bi-directional Battery Charger (VFBC), which connects between two dc sources $V_{hi}$ and $V_{bat}$. $V_{bat}$ represents the dc battery source that sinks and sources energy from one side of the DAB. $V_{hi}$ is the dc link voltage that connects between the primary grid connection and the DAB, typically through an active rectifier system. For this work, since it is not necessary to consider the operation of this active rectifier, the dc link is simulated as a simple voltage source.

The DAB primary high-side H-bridge, and the low-side secondary H-bridge which are formed by complementary switch pairs, which connect through the intermediate-frequency switching inductor and transformer.

As shown in Fig. 6.1, voltage and current sensors are placed at various positions for acquiring measurements. The voltage sensors measure the primary side dc link voltage and the secondary side battery voltage. The current sensors monitor the primary and secondary dc currents and the high-side and low-side ac link currents.

**Controller stage**

Figure 6.2 shows the simulation schematic diagram of the controller stage which includes all the C blocks, current reference calculations, analogue conditioning and scaling, and the PWM structure used to create the modulation commands for the converter switches.
Figure 6.1. PSIM schematic diagram of power stage of Bi-directional Battery Charger (VFBC).
Figure 6.2. PSIM schematic diagram of controller stage of VFBC.
Figure 6.3. PSIM schematic diagram of Pulse Width Modulator.

Figure 6.4. PSIM schematic diagram of analog conditioning.

Figure 6.3 shows more detail of the PWM modulator which generates the switching gate signals needed by the power stage. The modulator comprises a comparator which compares the input modulation reference waveform against a triangular carrier waveform to create the on-off commands for the IGBT switches. It also includes a deadtime generation block to account for the effects of phase leg deadtime on the converter performance.

Figure 6.4 shows more detail of the analogue conditioning circuit which is made up of first-order low-pass filters (LPF) and limiters for analogue to digital converter scaling. This analogue conditioning circuit breaks down into four sub blocks to measure both AC and DC voltages and currents.
Figure 6.5 and Figure 6.6 show the PSIM schematic diagram of the high-side dc link voltage and current measurements, respectively. Figure 6.7 and Figure 6.8 show the PSIM schematic diagram of the low-side dc voltage and current measurements, respectively. Note that these simulations are arranged to replicate the experimental controller circuitry, to achieve as close a match between simulation and experiment as is possible. In particular, the physical analogue circuit used for measuring voltage from a sensor is functionally represented using low-pass filters (LPF), limiters (clamps) and offset to shift the signal from negative to positive and to have only positive values. The current measurement block is made up of the LEM turns ratio, burden resistor, the first-order low-pass filters (LPF) representing differential amplifier of op-amp, limiting blocks to clamp the outputs of the op-amp and DSP supply rails, and offset and DSP clamp circuitry.

![Figure 6.5. PSIM schematic diagram of high side dc voltage measurement.](image1)

![Figure 6.6. PSIM schematic diagram of AC link current measurement.](image2)
6.1.2. MATLAB

All the algorithms developed for the 3D graph based core selection method in this thesis were implemented in MATLAB® (MathWorks, Inc.) using MATLAB scripts. MATLAB is a powerful numerical computing platform which allows for solving sophisticated engineering and scientific problems. It provides a Graphical User Interface (GUI) with vast libraries of pre-built toolboxes.

Using the “meshgrid” function from such toolboxes, a grid of vectors γ and $k_u$ were created to obtain a full grid of high resolution X and Y coordinate arrays. This grid was then given as an input to the “mesh” function to form a 3D surface plot corresponding to Figure 6.7. PSIM schematic diagram of low-side dc voltage measurement.

Figure 6.7. PSIM schematic diagram of low-side dc voltage measurement.

Figure 6.8. PSIM schematic diagram of low-side dc current measurement.

6.1.2. MATLAB

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Using the “meshgrid” function from such toolboxes, a grid of vectors γ and $k_u$ were created to obtain a full grid of high resolution X and Y coordinate arrays. This grid was then given as an input to the “mesh” function to form a 3D surface plot corresponding to
variation in $A_P$ with respect to different $\gamma$ and $k_u$ values. The values of $A_P$ were then color-coded for better visualization so that each colour represented a specific $A_P$ value. The final $A_P$ was selected through an iterative process to narrow-down the output values until the design conditions were satisfied and the algorithm converged to an optimum $A_P$ value. This final $A_P$ value was displayed by an astrix which superimposed on the 3D surface plot.

The MATLAB script for the core selection algorithm is listed in Appendix B.

6.2. Experimental Setup

The theoretical developments and simulation results of the new core selection method for the inductor design are required to be validated in a real experimental system. This is a necessary step for the final verification of the developed method as various practical effects may limit and degrade the performance of the magnetic component and consequently the converter performance. The implementation of the experimental system identifies these practical limitations and allows any required correction and optimisation, to ensure a fully practical outcome. The experimental results obtained using this experimental setup have been presented in Chapter 5.

This section describes the experimental systems in two sections. First, the exemplar inductor prototyping is presented. Secondly, the converter hardware is presented for both the power stage of the converter and the controller stage. The basic boards for this converter were provided by Creative Power Technologies (CPT) [168] and the Power Electronics Group at RMIT University.

6.2.1. Overview of the Experimental Setup

A block diagram of the experimental system is shown in Figure 6.9. the experimental test bench is a Vanadium flow battery charger comprising three-phase voltage-source inverter (VSI) and 2.5-kW dual-active-bridge (DAB) bidirectional DC-DC converter which are supplied by CPT [168].

The power converter systems used in this thesis are made up of a power stage structure which is operated by a DSP controller unit. From Figure 6.9, it can been seen that the experimental setup comprises five main hardware blocks, as follows:
Chapter 6  Description of Simulation and Experimental Systems

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- Inductor under test:
  The exemplar inductor is the prototype inductor that was constructed based on the DAB series inductor specifications.

- Power stage:
  The power stage is made up of the incoming DC voltage supply and the two switching converters (the VSI and the DAB).

- Controller stage:
  The DSP based controller board.

- Loads:
  The vanadium battery equivalent was made up using 8 off 12-V car batteries connected in a series parallel arrangement.

- Measurement instruments:

Figure 6.9. Block diagram of experimental setup.
Chapter 6  Description of Simulation and Experimental Systems

The measurement instruments include the differential voltage probes, current probes, digital CROs and Agilent data logger.

6.2.2. Inductor Prototyping

The inductor was built using a UI ferrite core (EPCOS Ferrite N87, UI 93/104/20) with an air gap length of \( l_c = 1.58 \text{ mm} \). The windings were made using litz-wire with 19 strands of 0.355 mm diameter for each individual strand. The number of turns was \( N = 40 \). Using both limbs of the core, each winding comprised four layers of 5-turns of trifilar litz-wire. Kapton tape was used as the primary insulator between each layer. The bobbins used for this prototype were designed specifically for this project to improve thermal performance, with nudges included inside the bobbin wall to allow better air flow around the core and barrier spacers on the bobbin frame to keep the windings away from the air gap, as shown in Figure 6.10.

![Figure 6.10. Custom designed bobbins: (a) inclusion of nudges inside bobbin walls (b) spacer barrier next to the air gap side.](image-url)
A frequency sweep impedance adapter was used to measure the inductance value at the desired switching frequency of 20 kHz. The inductor winding, the final manufactured prototype inductor and its frequency sweep measurement are shown in Figure 6.11.

Figure 6.11. Prototype inductor: (a) inductor winding with litz-wire and modified bobbin (b) final manufactured inductor with the holding frame (c) frequency sweep measurement.
6.2.3. Power Stage

**Input Supply**

A grid connected three phase VSI operating as an active rectifier was used as the high voltage input supply to the Battery charger.

**Power Converter**

The DAB bi-directional DC-DC converter is made up of two 1200V, 50 A IGBT pairs to form two converter phase legs as shown in Figure 6.12. These are linked together with the DC bus which is made of copper bars and then connected to four bulk capacitors of 4mF capacitance, shown in Figure 6.13. Four film capacitors are attached directly to the underside of the DC busbars to provide high frequency switching support. Since the converter used in this experimental system was a general-purpose system, the power devices were not selected for the work and were significantly overrated. However, this was not important, since the project aim was to focus on inductor design and evaluation.

**Load**

The converter load is 8 12-V DC car batteries connected in pairs with the pairs connected in series as the setup is a 2.5 kW bidirectional battery charger. Figure 6.14 shows the battery load.

![Figure 6.12. Experimental DAB with IGBTs and DC busbars.](image-url)
Figure 6.13. Experimental DAB bulk capacitors.

Figure 6.14. Experimental battery loads.
6.2.4. Controller Stage

The power stage converters, VSI and DAB, were each switched by controller boards supplied by CPT. These controller boards are based on the Texas Instruments TMS320F2810 Digital Signal Processor (DSP), a powerful 150 MHz 32-bit fixed-point microprocessor capable of handling all converter control tasks. It interfaces to the power stage via the CPT-DA2810 and CPT-E13 support PCB assemblies.

**DSP Controller**

The TMS320F2810 DSP (also known as 2810 DSP) is a well-known microprocessor by Texas Instruments which is explicitly designed for motor and power electronic control applications. It has many peripheral features including two independent event manager modules (EVA and EVB), each containing general purpose (GP) timers, full-compare/PWM units and capture units to generate PWM signals, the Analog-to-Digital Converter (ADC) module, Central Processing Unit (CPU) and Serial Communication Interface (SCI) module.

**CPT-DA2810 Controller Board**

The CPT-DA2810 [169] is a standardised DSP controller board designed to provide a fully flexible interface between the 2810 DSP microprocessor and the lower level motherboards. The functional block diagram of this controller board is shown in Figure 6.15. This controller board offers several functionalities including; a JTAG interface for chip programming, RS-232 for serial communications, analogue input protection circuitry and low voltage power supply.

**CPT-E13 Controller Board**

The CPT-E13 [170] board is a controller motherboard and the DA2810 DSP is directly plugged into it. This controller offers many features including; isolated digital inputs and outputs, analogue inputs for conditioning and protection circuitry, gate drive PWM interface, isolated power supply and serial ports for communication interface. Figure 6.16 shows the functional block diagram of this controller board.
Chapter 6  Description of Simulation and Experimental Systems

Figure 6.15. Functional block diagram of the CPT-DA2810 controller card [168]

Figure 6.16. Functional block diagram of the CPT-E13 controller board [169]
Chapter 6  Description of Simulation and Experimental Systems

Figure 6.17. CPT controller boards (CPT-E13 and CPT-DA2810)

Figure 6.18. CAD image of the DC-DC converter enclosure showing temperature measurement point and the air flow direction
The controller board and the CAD image of the DC-DC converter are shown in Figure 6.17 and Figure 6.18, respectively.

6.2.5. Software

The DSP code used to run the control unit of the experimental system was adapted from previous code structures by Power and Energy Group (PEG) postgraduate students and the CPT engineers. The code contains serial communication, analogue-to-digital converter (ADC) readings, closed-loop modulation and a user interface functionality.

6.3. Summary

This chapter has presented an overview of the PSIM and MATLAB based simulation systems that were used for this study. The power stage and control stage were represented using standard power electronic elements and control blocks in PSIM, while the MATLAB simulation system was used for developing the 3D graph based core selection method using MATLAB scripts and m.files. The chapter then presents the experimental prototype inductor and Vanadium Flow Battery Charger system used to validate the work presented in this thesis. The inductor was built using a UI 93/104/20 ferrite core (EPCOS, N87), split windings using trifilar litz-wire (19 strands of 0.355mm diameter strand), custom designed bobbins and an air gap length of 1.58mm. The charger is based on a 2.5-kW dual-active-bridge (DAB) bidirectional DC-DC converter. The CPT-DA2810 and CPT-E13 controller boards together with the TI TMS320F2810 DSP were used as the DSP based control unit to control this converter.
Chapter 7

SUMMARY

The work carried out in this thesis has addressed some of the shortcomings regarding the design and optimisation of higher power inductors operating in the intermediate switching frequency range, i.e.:

- Lack of knowledge and limited methodologies for the design of higher power inductors operating at the intermediate frequency range.
- Challenges with design parameter initialization and core selection for this frequency range.
- Iterative and time consuming design techniques that require significant expertise to achieve an effective design outcome.
- The results of generic design processes rely heavily on the knowledge and background of the designer.

The work has also proposed and developed a semi-automatic and user-friendly design advisory system to help engineers designing high-performance high-power inductors, with ANSYS based Finite Element Analysis (FEA) evaluation used to refine the parameters determined by the 3D graph-based design technique presented in this thesis. This system has been designed to be not only beneficial to expert designers but also to be used by less experienced engineers who have more limited experience and background about the design of higher power magnetic components.

7.1. Summary of work

The contributions of this work can be classified into three major areas, as follows:
Chapter 7  Summary

7.1.1. Develop an expert system to assist with the magnetic component design process

The research work has presented a Graphical User Interface (GUI) that links to a knowledge based advisory system for the design of magnetic components. The overall system creates a novel inductor/transformer knowledge-based advisor and support system which integrates the outcome of a generic design process with ANSYS® Maxwell as an FEA tool. One major benefit of the system is its capability to recommend design alternatives to a user to meet a set of target specifications. The system then evaluates the design to provide a multi-objective optimization outcome in terms of volume, temperature rise and localized hot spots.

7.1.2. Propose a new strategy for the selection of magnetic cores

From the research work, a significant weakness was identified within the existing body of knowledge concerning the design of inductors operating at intermediate switching frequencies, i.e. the problem of selecting a proper initial starting point for the fundamental parameters WUF ($k_u$) and CCLR ($\gamma$). The work then identified that typical initial assumptions that are made for values of WUF and CCLR for designs in the higher frequency range, are not applicable to intermediate frequencies. From this understanding, a novel 3D graphical strategy was proposed to determine values for these two design parameters that does not require an initial estimate, thus significantly reducing the number of design iterations required and also achieving an improved design outcome. The proposed strategy is an automated 3D method that graphically identifies the theoretical relationship between $A_P$ values and $k_u$ and $\gamma$ across a broad range. Second-order effects such as skin and proximity effects are included into the design procedure, and constraints such as limitation of current density, inductance values, and core and winding loss distributions are used to narrow the feasible region of core selection. The results led to a greatly reduced subspace of $A_P$ values that are physically viable for the inductor design – often limiting the core selection to essentially one alternative (a single pass design selection) and allowing a single pass core selection to be made with confidence.
7.1.3. Verification of the proposed Design process

The last part of the work verified the feasibility and performance of the new core selection methodology using a prototype inductor for the series inductor of the Dual Active Bridge (DAB) converter, chosen because of its more challenging High Frequency (HF) fundamental current. The design success was confirmed by making a prototype inductor for a real converter, operating it under full load conditions, and comparing the resulting performance between simulation and experimental measurements. As well as confirming the electrical performance of the inductor, the temperature rise for both core and windings was monitored over a long period of time (i.e. approximately 2.5 hours) to confirm the anticipated thermal performance of the design.

7.2. Future work

While this thesis has proposed a general core selection methodology for the design of higher power magnetic components operating at intermediate switching frequencies, the approach has only been validated and tested for a single exemplar inductor operating in this frequency range. However, from this work, a number of possible further research avenues can be identified:

- For the exemplar inductor design case presented in this thesis, the outcome of the core selection process was a single point. However, over constraining the core selection process could lead to no viable core selection points on the 3D graphical outcome, while under constraining the core selection could result in a series of selection points as the result. Further work is required to determine if and when such less satisfactory selection outcomes may occur, and how the selection constraints should then be adapted to accommodate this situation.
- What other factors should be considered in the core selection constraint process? Some possible additional factors are core material degradation dependency and variable thermal resistance within the core and winding structures.
- A fan forced environment makes the thermal calculation more complicated and more advanced thermodynamic equations and calculations would be required to accurately predict the maximum allowable power loss for fan forced design
condition. Evaluation of $R_p$ for larger inductors in a fan forced environment is an area of future research.

- The core selection strategy could readily be adapted to create a tool for designing magnetic components across a much broader frequency range (i.e. intermediate and high frequency range). However, its applicability and suitability would need to be verified with the design and verification of a further series of physical magnetic elements.

- While in principle the same core selection process could be applied to the transformer core selection process, it is probable that different constraints will need to be applied to reduce the feasible region of core selection for this alternative magnetic component. Such an investigation is identified here as future possible work arising from this project.

7.3. Closure

A typical inductor design procedure involves several iterative steps including core selection, winding selection, power handling capacity and thermal management. A major challenge for the design process is to select an appropriate core as the first stage of the design process. This thesis has provided a novel methodology to facilitate this process, which is particularly suited to the design of higher power inductors for converters operating at intermediate switching frequencies.
Appendix A
CODE FOR MULTI OBJECTIVE OPTIMISATION

This appendix contains the developed code for the multi objective optimiser used for the knowledge based design advisory system in Chapter 3.

```java
import java.io.IOException;
import java.io.InputStream;
import java.util.*;
import org.moeaframework.Executor;
import org.moeaframework.core.NondominatedPopulation;
import org.moeaframework.core.Solution;
import org.moeaframework.core.variable.EncodingUtils;
import org.moeaframework.examples.ga.knapsack.Knapsack;
import org.moeaframework.util.Vector;

public class testExample {
    public static void main(String[] args) throws IOException {
        // solve using NSGA-II
        NondominatedPopulation result = new Executor()
            .withProblemClass(testProblem.class)
            .withAlgorithm("NSGAII")
            .withMaxEvaluations(300000)
            .withProperty("populationSize", 1000)
            .run();

        // print the results
        List<double[]> perfs = new ArrayList<double[]>();
        List<double[]> designs = new ArrayList<double[]>();
        for (int i = 0; i < result.size(); i++) {
            Solution solution = result.get(i);
            if (!solution.violatesConstraints()) {
                double[] objectives = solution.getObjectives();
                int pwwID = EncodingUtils.getInt(solution.getVariable(1));
                System.out.println("Solution "+ (i+1) + ":");
                System.out.println("totalloss: " + objectives[0]);
                System.out.println("coreweight: " + objectives[1]);
                System.out.println("ldiff: " + objectives[2]);
                System.out.println("air_gap: " +
                    (double)EncodingUtils.getInt(solution.getVariable(0))/10.0);
                System.out.println("pww: " + Math.pow(92, (double)(36 - pwwID)/39));
            }
        }
    }
}
```

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System.out.println("Sp C P: " + (double)EncodingUtils.getInt(solution.getVariable(2))/10.0);
System.out.println("M10: " + (double)EncodingUtils.getInt(solution.getVariable(3))/10.0);
System.out.println("M21: " + EncodingUtils.getInt(solution.getVariable(4)));
System.out.println("M22: " + EncodingUtils.getInt(solution.getVariable(5)));
System.out.println("N: " + EncodingUtils.getInt(solution.getVariable(6)));
System.out.println("Nl: " + EncodingUtils.getInt(solution.getVariable(7)));
System.out.println(solution.violatesConstraints());
double[] perf = Arrays.copyOfRange(objectives, 0, 2);
boolean overperform = false;
int j = 0;
for (j = 0; j < perfs.size(); j++) {
double[] pre_perf = perfs.get(j);
if ((perf[0] >= pre_perf[0]) & (perf[1] >= pre_perf[1])) {
  overperform = true;
  break;
} else if ((perf[0] <= pre_perf[0]) & (perf[1] <= pre_perf[1])) {
  break;
}
}
if (!overperform) {
  if (j < perfs.size()) {
    perfs.remove(j);
    designs.remove(j);
  }
  perfs.add(perf);
double[] design = new double[8];
design[0] = (double)EncodingUtils.getInt(solution.getVariable(0))/10.0;
design[1] = (double)EncodingUtils.getInt(solution.getVariable(1));
design[2] = (double)EncodingUtils.getInt(solution.getVariable(2))/10.0;
design[3] = (double)EncodingUtils.getInt(solution.getVariable(3))/10.0;
design[4] = (double)EncodingUtils.getInt(solution.getVariable(4));
design[5] = (double)EncodingUtils.getInt(solution.getVariable(5));
design[6] = (double)EncodingUtils.getInt(solution.getVariable(6));
design[7] = (double)EncodingUtils.getInt(solution.getVariable(7));
designs.add(design);
}
System.out.println("-------------- Wire type: " + testProblem.wire_type + "; Winding type: " + testProblem.winding_type + " ----------------------");
for (int i = 0; i < perfs.size(); i++) {
double Air_gap = designs.get(i)[0];
double Pww = 0.127 * Math.pow(92, (double)(36 - designs.get(i)[1])/39);
double Sp_C_P = designs.get(i)[2];
double M10 = designs.get(i)[3];
double M21 = (int)designs.get(i)[4];
double M22 = (int)designs.get(i)[5];
int N = (int)designs.get(i)[6];
int Nl = (int)designs.get(i)[7];
double ae = testProblem.Ae(Nl);
double Rtotal = testProblem.Rtotal(ae, Air_gap);
double coreloss = testProblem.coreloss(Rtotal, ae, N, Nl);
double conductorloss = testProblem.conductorLoss(Pww, Sp_C_P, M10, M21, M22, N, Nl);
double coreweight = testProblem.coreweight(Nl);
double windowconst = testProblem.windowPercentage(Pww, Sp_C_P, M10, M21, M22, N);
double temperconst = testProblem.heatDissipation(Nl);
double bmaxconst = testProblem.bmax(N, Rtotal, ae);
double Lconst = testProblem.L(Rtotal, N);
System.out.println("Solution " + (i+1) + ":");
System.out.println("Performance ----------------------");
System.out.println("core loss: " + coreloss);
Appendix A  Code for Multi Objective Optimisation

```
System.out.println("conductor loss: "+ conductorLoss);
System.out.println("total loss: "+ perfs.get(i)[0]);
System.out.println("core weight: "+ coreWeight);
System.out.println("window percentage: "+ windowConst);
System.out.println("temperature: "+ temperConst);
System.out.println("Bmax: "+ bmaxConst);
System.out.println("J: "+ Jconst);
System.out.println("L: "+ Lconst);
System.out.println("Design parameters =============");
System.out.println("Air gap: "+ Air_gap);
System.out.println("PwwID: "+ designs.get(i)[1]);
System.out.println("Pww: "+ Pww);
System.out.println("Sp-C_P: "+ Sp_C_P);
System.out.println("M10: "+ M10);
System.out.println("M21: "+ M21);
System.out.println("M22: "+ M22);
System.out.println("N: "+ N);
System.out.println("Nl: "+ Nl));
```
Appendix A  Code for Multi Objective Optimisation

```java
public static double Fe = 1.1;
public static double Wh = 48;
public static double Ww = 34.6;
public static double Ve = 144000;
public static double Ch = 105.05;
public static double Cv = 90.60;
public static double Kc = 16.9;
public static double alpha = 1.25;
public static double beta = 2.35;
public static String wire_type = "litz";
public static String winding_type = "double";
public static int strands = 19;
public static double Pwg = 1;
public static double hc = 25;
public static double Tmin = 50;
public static double Tmax = 500;

public testProblem () {
    super(8,3,5);
}

@Override
public void evaluate(Solution sol) {
    double coreloss = 0;
    double conductorloss = 0;
    double Ldiff = 0;
    double coreweight = 0;
    double windowconst = 0;
    double temperconst = 0;
    double bmaxconst = 0;
    double Jconst = 0;
    double Lconst = 0;
    double Ae = Ae(Nl);
    double Rtotal = Rtotal(ae, Air_gap);
    coreloss = coreloss(Rtotal, ae, N, Nl);
    conductorloss = conductorLoss(Pww, Sp_C_P, M10, M21, M22, N, Nl);
    Ldiff = Math.abs(L(Rtotal, N) - L_req);
    coreweight = coreweight(Nl);
    windowconst = windowPercentage(Pww, Sp_C_P, M10, M21, M22, N) - 0.8;
    temperconst = heatDissipation(Nl) - Tmax;
    bmaxconst = bmax(N, Rtotal, ae) - Bmax;
    Jconst = J(Pww, M21, M22, Nl) - Jmax;
    Lconst = Ldiff - 5;
    if (windowconst < 0) {
        windowconst = 0.0;
    }
    if (temperconst < 0) {
        temperconst = 0.0;
    }
    if (bmaxconst < 0) {
        bmaxconst = 0.0;
    }
    if (Jconst < 0) {
        Jconst = 0.0;
    }
    if (Lconst < 0) {
        Lconst = 0.0;
    }
```

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Appendix A  Code for Multi Objective Optimisation

```java
sol.setObjective(0, coreloss + conductorloss);
sol.setObjective(1, coreweight);
sol.setObjective(2, Ldiff);
sol.setConstraint(0, windowconst);
sol.setConstraint(1, temperconst);
sol.setConstraint(2, bmaxconst);
sol.setConstraint(3, Jconst);
sol.setConstraint(4, Iconst);

@Override
public Solution newSolution() {
    Solution sol = new Solution(numberOfVariables, numberOfObjectives, numberOfConstraints);
    sol.setVariable(0, EncodingUtils.newInt(1, 20));
    sol.setVariable(1, EncodingUtils.newInt(1, 40));
    sol.setVariable(2, EncodingUtils.newInt(1, 50));
    sol.setVariable(3, EncodingUtils.newInt(1, 50));
    sol.setVariable(4, EncodingUtils.newInt(1, 10));
    sol.setVariable(5, EncodingUtils.newInt(1, 4));
    sol.setVariable(6, EncodingUtils.newInt(1, 500));
    sol.setVariable(7, EncodingUtils.newInt(1, 3));
    return sol;
}

public static double Ae (int Nl) {
    double Cd = Lam_t * Nl * Lff;
    double Ae = Lam_w * Cd;
    return (Ae);
}

public static double Rtotal (double Ae, double Air_gap) {
    double Rl = 2 * (Cw - Lam_w) + 2 * (Ch - Lam_w);
    double Rcore = Rl/(uair * ur_ferrite)/Ae*1000;
    double Rair = 2 * Air_gap/uair/(Ae*Fe)*1000;
    double Rtotal = Rcore + Rair;
    return (Rtotal);
}

public static double coreloss(double Rtotal, double Ae, int N, int Nl) {
    double Bmax_calc = N * Ip /( Ae / 1000000 ) / Rtotal;
    double Cv = (Ve/1000000000)*Nl;
    double Pc = Cv * Kc * Math.pow(Bmax_calc, beta) * Math.pow(fs, alpha);
    return (Pc);
}

public static double bmax(int N, double Rtotal, double Ae) {
    double Bmax_calc = N * Ip / (Ae / 10000) / Rtotal;
    return (Bmax_calc);
}

public static double L(double Rtotal, int N) {
    double L_calc = Math.pow(N, 2) / Rtotal * 1000000;
    return (L_calc);
}

public static double J(double Pww, int M21, int M22, int Nl) {
    double Pwa = 0;
    if (wire_type.equals("round")) {
        Pwa = Math.PI * Math.pow((Pww/2), 2);
    } else if (wire_type.equals("litz")) {
        Pwa = strands * Math.PI * Math.pow((Pww/2), 2);
    }
    double J = (Ip/Pwa) / M21 / M22;
    return (J);
}

public static double coreweight(int Nl) {
    double Cv = (Ve/1000000000)*Nl;
    double Cm = Cv * dens_core;
    return (Cm);
}

public static double conductorLoss (double Pww, double Sp_C_P, double M10, int M21, int M22, int N, int Nl) {
    double Pwa = 0;
```
double M12 = 0;
double M13 = 0;
double Cd = Lam_t * Nl * Lff;
double S36 = 0.1;
double M38 = N * 1;
double M39 = M38 / 1000;
double M45 = Pwa * 0.000001;
double M52 = r_cu_m / M45;
double M53 = M52 * M39;
double M54 = M53 / (M21 * M22);
P_cu = (Math.pow(Ip, 2) * M54);
}
else if (winding_type.equals("double")) {

double S21 = M21;
double S27 = Math.floor((Ww - 2 * M10) / M12 / M22);
double S28 = Math.ceil(N / S27 / 2);
double S22 = M22;
double S35 = 2 * (Lam_w + 2 * Sp_C_P + (2 * (Pwg + M13) * (S28 * S21)) / 2 + Cd + 2 * Sp_C_P);
double S37 = S35 * (1 + S36);
double S45 = Pwa * 0.000001;
double S52 = r_cu_m / S45;
double S53 = S52 * S39 * 2;
double S54 = S53 / (S21 * S22);
P_cu = (Math.pow(Ip, 2) * S54);
}
return (P_cu);

public static double windowPercentage(double Pww, double Sp_C_P, double M10, int M21, int M22, int N) {

double Pwa = 0;
double P_lay = 0;
double M12 = 0;
double M13 = 0;
double perc = 0;
if (wire_type.equals("round")) {
    Pwa = Math.PI * Math.pow((Pww / 2), 2);
    M12 = Pww;
    M13 = Pww;
}
else if (wire_type.equals("litz")) {
    Pwa = strands * Math.PI * Math.pow((Pww / 2), 2);
    M12 = 2 * Math.sqrt(Pwa / Math.PI);
    M13 = M12;
}
if (winding_type.equals("single")) {
    double M27 = Math.floor((Ww - 2 * M10) / M12 / M22);
    double P_lay = Math.ceil(N / M27);
    double P_thick = (Pwg + M13) * (P_lay * M21) + Sp_C_P;
    perc = P_thick / Ww;
}
else if (winding_type.equals("double")) {
    return (P_cu);
}
Appendix A  
Code for Multi Objective Optimisation

double S27 = Math.floor((Wh - 2 * M10) / M12 / M22);
double S28 = Math.ceil(N / S27 / 2);
double S29 = (Pwg + M13) * (S28 * S21) + Sp_C_P;
perc = S29 / (Ww / 2);
}
return (perc);
}

public static double heatDissipation (int Nl) {
double Cd = Lam_t * Nl * Lff;
double At = ((Cw * Ch - Wh * Ww * 2) + Cw * Cd + Ch * Cd) * 2 / 1000000;
double R_thermal = 1 / (hc * At);
double P_D = T_rise / R_thermal;
return (P_D);
}
Appendix B

MATLAB Code for 3D Graph-Based Core Selection

This appendix presents the developed code in MATLAB for the 3D graph-based core selection method in Chapter 4.

```matlab
% NY inductor with a gapped core in DAB converter Wa to Ac Ratio for any Ap and AC copper loss calculation

clear all
close all
clc
fs=20e3;
T_ambient=50;
T_rise=20;
u0=4*pi*1e-7;
ur=1950;
sigma=5.814e7;%Copper conductivity
p20=1.72e-8;%Copper resistivity

% specifications of the core material
Kc=16.9;
alpha=1.25;
beta=2.35;

% specifications of the inductor
L=550e-6;
I_max=8.5;
I_min=-8.5;
B_max=0.2;
D0=0.5;
ku=0.125;%Window utilization factor

% core selection
I_rms=round(sqrt(((I_max+I_min)/2)^2+1/3*((I_max-I_min)/2)^2));
Ki=round(I_rms/I_max,1); % Waveform factor
Kt=4.822*10^-4;%Dimensional constant
Gamma_range=0:0.1:40;
ku_range=0.1:0.01:0.8;
counter_ku=1;
counter_gamma=1;
counter_J_c_density=1;
cwnt=1;
for Gamma=0:0.1:40
for ku=0.1:0.01:0.8
```

123
Ap(counter_gamma,counter_ku)=((sqrt(1+Gamma)*Ki*L*I_max^2)/(B_max*Kt*sqrt(ku*T_rise)))^(8/7)*10^8;  % in cm^4
J_c_density(counter_gamma,counter_ku)=Kt*sqrt(T_rise)*10^-4/(sqrt(ku*(1+Gamma))*(Ap(counter_gamma,counter_ku)*10^-8)^(1/8));  % current density in A/cm2
Aw(counter_gamma,counter_ku)=I_rms/J_c_density(counter_gamma,counter_ku);  % wire cross-sectional area in cm2
WA(counter_gamma,counter_ku)=Aw(counter_gamma,counter_ku)/ku;  % in cm2
MLT_calc(counter_gamma,counter_ku)=sqrt(Ap(counter_gamma,counter_ku)/WA(counter_gamma,counter_ku))*4;
P_copper(counter_gamma,counter_ku)=(p20*MLT_calc(counter_gamma,counter_ku)/Aw(counter_gamma,counter_ku))*I_rms^2;
P_total(counter_gamma,counter_ku)= P_copper(counter_gamma,counter_ku)*(1+Gamma);
T(cvnt,:)= [ku ; Gamma ; Aw(counter_gamma,counter_ku)];
cvnt=cvnt+1;
if (J_c_density(counter_gamma,counter_ku)<=200 && J_c_density(counter_gamma,counter_ku)>=100)
gamma_sv_J_c_density(counter_J_c_density)=Gamma;
ku_sv_J_c_density(counter_J_c_density)=ku;
J_c_density_selected(counter_J_c_density)=J_c_density(counter_gamma,counter_ku);
AP_selected2(counter_J_c_density)=Ap(counter_gamma,counter_ku);
AW_selected2(counter_J_c_density)=Aw(counter_gamma,counter_ku);
counter_J_c_density=counter_J_c_density+1;
end;
counter_ku=counter_ku+1;
end;
counter_ku=1;
counter_gamma=counter_gamma+1;
end;
figure(1);
Ap=Ap';
[gamma_2,ku_2] = meshgrid(Gamma_range,ku_range);
mesh(gamma_2,ku_2,Ap);
xlabel('Gamma');
ylabel('Ku');
zlabel('Ap cm^4');
title ('Optimum Design Parameter');
hold on;
plot3(gamma_2,ku_2,gamma_sv_J_c_density,kappa,ku_sv_J_c_density,AP_selected2,'r.');
figure(3);
mesh(gamma_2,ku_2,Ap);
xlabel('Gamma');
ylabel('Ku');
zlabel('Ap cm^4');
title ('Optimum Design Parameter');
hold on;
plot3(gamma_2,ku_2,gamma_sv_J_c_density,kappa,ku_sv_J_c_density,AP_selected2,'r.');

delta_skin=0.066*sqrt(1/fs);  % skin depth in m
porosity_factor=0.9;
NL=4;  % number of layer
k=19;  % number of strands in each litz bundle
D_str=0.355e-3;  % strand diameter in m
figure;
for i=1:length(AP_selected2)
for x=2:0.1:4
  Ac(i)=sqrt(AP_selected2(i)/x); % in cm2
  lc(i)=5.6*(AP_selected2(i)^(3/4))/Ac(i); % in cm
  Wa(i)=x*Ac(i); % window area in cm2
  Aw_wire(i)=1_m_us/J_c_density_selected(i); % wire cross-sectional area in cm2
  d(i)=sqrt((pi/4)*Aw_wire(i)); % wire diameter in cm
  A_str=((pi/4)^0.75)*sqrt(1/orosity_factor)/delta_skin*d_str;
  F_ac_dc=A_str*((sinh(2*A_str)+sin(2*A_str))/(cosh(2*A_str)-cos(2*A_str))+(2*(k*NL^2-1))/3)*((sinh(A_str)-sin(A_str))/((cosh(A_str)+cos(A_str)))); % Rac to Rd0 ratio

for l_gap=0.1:0.1:3 % air gap in mm
  ap_sel(counter2)=AP_selected2(i);
  gamma_sel(counter2)=gamma_sv_J_c_density(i);
  ku_sel(counter2)=ku_sv_J_c_density(i);
  jc_density_sel(counter2)=J_c_density_selected(i);
  Number_of_turn1(counter2)=round(ku_sv_J_c_density(i)*(x*Ac(i))/AW_selected2(i));
  L(counter2)=((Number_of_turn1(counter2)^2)/((lc(i)*0.01/(u0*ur*Ac(i)*0.0001))+(l_gap*0.001/(u0*Ac(i)*0.0001)))); % in H
  gap(counter2)=l_gap;
  p_cop_test2=10*p20*ku_sel(counter2)*((ap_sel(counter2)*10^-8)^(3/4))*(jc_density_sel(counter2)*(10^4))^2;
  plot(ap_sel(counter2),p_cop_test2,'r.');
  hold on;
  pause(0.01);
end end
end
%
¶---------------------
Test---------------------
AP_test=93;
ku_sel_test=0.12;
gamma_sel_test=26.8;
P_copper_test=0;
P_core_test=0;

for x=2:1:4
  Ac_test(x)=sqrt((AP_test)/x); % in cm2
  %Ac_test(x)=sqrt(1./(x'/(AP_test)));
  lc_test(x)=5.6*(AP_test^(3/4))/Ac_test(x); % in cm
  Wa_test(x)=x*Ac_test(x); % window area in cm2
  Vc_test(x)=lc_test(x)*Ac_test(x); % core volume in cm3
  MLT_test(x)=4*sqrt(Ac_test(x)); % in cm
  J_c_density_test=Kt*sqrt(T_rise)*10^-4/(sqrt(ku_sel_test*(1+gamma_sel_test))*(AP_test*10^-8)^(1/8)); % current density in A/cm2
  Aw_test=x*Ac_test(x); % wire cross-sectional area in cm2
  Number_of_turn1_test(x)=round(ku_sel_test*(x*Ac_test(x))/Aw_test);
  P_copper_test(x)=(p20*Number_of_turn1_test(x)*MLT_test(x)*10^-2)*(I_rms^2)/
  (Aw_test*10^-4); % copper loss in W
  P_core_test(x)=Kc*(fs^alpha)*(B_max^beta)*((Vc_test(x)*(10^-6))); % core loss in W
  P_total_test(x)=P_core_test(x)+P_copper_test(x);
end
%
figure (2);
plot3(Number_of_turn1,gap,L,'r.');
xlabel('Number of Turn');
ylabel('Air Gap Length (mm)');
zlabel('Inductance L(H)');
grid on;
%
Limiting inductance to narrow down ku and gamma range L=550uH
induct_selected=find(L<0.000551 & L>0.000549);
Appendix B  MATLAB Code for 3D Graph-Based Core Selection

```matlab
% hold on;
% plot3(Number_of_turn1(induct_selectected),gap(induct_selectected),L(induct_selectected),'b* ');
%%
figure(3);
hold on;
plot3(gammaSel(induct_selected),kuSel(induct_selected),apSel(induct_selected),'g*'); %Ap surface constrained by Inductance
%% Loss
P_copper=0;
P_core=0;
for i=1:length(kuSel)
P_copper(i)=10*p20*kuSel(i)*((apSel(i)*10^-8)^(3/4))*(jc_densitySel(i)*(10^4))^2;dc
copper loss in W
P_copper_ac(i)=P_ac_dc*P_copper(i); %ac copper loss in W
P_core(i)=5.6*Kc*(fs^alpha)*(B_max^beta)*((apSel(i)*(10^-8))^(3/4));%core loss in W
end
P_Copper_total=2* P_copper_ac;
P_total=P_Copper_total+P_core;

% loss_index=find(abs(P_core-P_Copper_total)<=0.0025);
% figure(3);
% plot3(gammaSel(loss_index),kuSel(loss_index),apSel(loss_index),'b*','linewidth',4);
% disp(['AP = ' num2str(min(apSel(loss_index)))])
% disp(['Gamma = ' num2str(unique(gammaSel(loss_index)))])
% disp(['Ku = ' num2str(unique(kuSel(loss_index)))]);
%% volume=5.6*(apSel.^3/4);
% figure(4);
% subplot(3,1,1);
% plot(apSel,P_Copper_total,'g.'); % Copper loss
% xlabel('AP');
% title('Copper Loss');
% subplot(3,1,2);
% plot(apSel,P_core,'k.'); % Copper loss
% xlabel('AP');
% title('Total Loss');
% subplot(3,1,3);
% plot(apSel,volume,'r.'); % Volume
% xlabel('AP');
% title('Volume');

figure(5);
plot3(gammaSel(induct_selected),kuSel(induct_selected),P_Copper_total(induct_selected),'r.');
hold on;
plot3(gammaSel(induct_selected),kuSel(induct_selected),P_core(induct_selected),'b.');
hold on;
grid on;
xlabel('Gamma');
ylabel('Ku');
zlabel('Loss');
legend('Total Winding Loss','Core Loss');
hold on;
plot3(gammaSel(loss_index),kuSel(loss_index),P_core(loss_index),'k*','linewidth',4);
```

Appendix C
UI 93/104/20 Core

This appendix provides the dimensions and magnetic characteristics of the core used in this work. The commercial core manufactured by EPCOS [171], UI 93/104/20, was used for prototyping and all the dimensions are given per set. Figure C.1 shows the dimensions of the UI core. Material specifications and magnetic characteristics of the core are listed in Table C.1 and Table C.2, respectively.

![Dimensions of the UI 93/104/20 by EPCOS](image)

Figure C.1. Dimensions of the UI 93/104/20 by EPCOS
### Table C.1. Material Specifications

**Ferrite N87 (EPCOS)**

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Steinmetz parameter</td>
<td>$K_c$</td>
<td>16.9</td>
</tr>
<tr>
<td>Steinmetz parameter</td>
<td>$\alpha$</td>
<td>1.25</td>
</tr>
<tr>
<td>Steinmetz parameter</td>
<td>$\beta$</td>
<td>2.35</td>
</tr>
<tr>
<td>Relative permeability</td>
<td>$\mu_r$</td>
<td>1950</td>
</tr>
<tr>
<td>Saturation flux density</td>
<td>$B_{sat}$</td>
<td>0.3T</td>
</tr>
</tbody>
</table>

### Table C.2. Magnetic Characteristics (per set)

**UI 93/104/20 (EPCOS)**

<table>
<thead>
<tr>
<th>Description</th>
<th>Label</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnetic path length</td>
<td>$l_e$</td>
<td>258</td>
<td>mm</td>
</tr>
<tr>
<td>Cross-sectional area</td>
<td>$A_e$</td>
<td>560</td>
<td>mm$^2$</td>
</tr>
<tr>
<td>Cross-sectional area</td>
<td>$A_{min}$</td>
<td>560</td>
<td>mm$^2$</td>
</tr>
<tr>
<td>Volume of the core</td>
<td>$V_e$</td>
<td>144000</td>
<td>mm$^3$</td>
</tr>
<tr>
<td>Mass</td>
<td>$m$</td>
<td>750</td>
<td>g/set</td>
</tr>
</tbody>
</table>
REFERENCES


References


References


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References


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