Modelling and Characterisation of Losses in Nanocrystalline Cores

A thesis submitted to the University of Manchester for the degree of
Doctor of Philosophy
in the Faculty of Engineering and Physical Sciences

2015

By
Yiren Wang

School of Electrical and Electronic Engineering
LIST OF CONTENTS

LIST OF CONTENTS ........................................................................................................... 2
LIST OF FIGURES ............................................................................................................. 6
LIST OF TABLES ................................................................................................................ 10
LIST OF ABBREVIATIONS .............................................................................................. 11
LIST OF SYMBOLS ........................................................................................................... 13
ABSTRACT .......................................................................................................................... 18
DECLARATION .................................................................................................................... 19
COPYRIGHT STATEMENT .................................................................................................. 20
ACKNOWLEDGEMENTS ..................................................................................................... 21

1 INTRODUCTION ............................................................................................................. 22
  1.1 Background .................................................................................................................. 22
  1.2 Objectives ................................................................................................................... 25
  1.3 Scope of This Thesis .................................................................................................... 26

2 LITERATURE REVIEW .................................................................................................. 28
  2.1 Introduction .................................................................................................................. 28
  2.2 DC-DC Converter Topologies ..................................................................................... 28
  2.3 Magnetic Components for DC-DC converters .......................................................... 33
    2.3.1 Magnetic Core Materials ...................................................................................... 35
    2.3.2 Winding Techniques ............................................................................................ 38
  2.4 Loss Mechanisms in Magnetic Components ............................................................ 40
    2.4.1 Core loss ................................................................................................................ 40
    2.4.2 Winding Loss ......................................................................................................... 51
    2.4.3 Gap Loss ................................................................................................................. 52
2.5 Modelling of Magnetic Components ................................................................. 55
  2.5.1 Electromagnetic Modelling ............................................................................. 55
  2.5.2 Thermal Modelling ......................................................................................... 57
2.6 Magnetic Component Design and Optimisation ............................................... 59
2.7 Summary of Literature Review ........................................................................... 62

3 CORE LOSS CHARACTERISATION OF NANOCRYSTALLINE CORES ........ 64
  3.1 Introduction ....................................................................................................... 64
  3.2 Experimental Set Up .......................................................................................... 64
  3.3 Measurement Results ......................................................................................... 68
    3.3.1 Measured Magnetisation Curves ................................................................. 69
    3.3.2 Effect of Duty Cycle on Core Loss .............................................................. 71
    3.3.3 Effect of DC Bias on Core Loss ................................................................... 75
  3.4 Conclusion ......................................................................................................... 82

4 FINITE ELEMENT MODELLING OF LAMINATED CORES FOR GAP LOSS CALCULATION ................................................................. 83
  4.1 Introduction ....................................................................................................... 83
  4.2 Modelling Method for Laminated Cores ......................................................... 83
  4.3 FEA Software .................................................................................................... 86
  4.4 Initial Inductor Modelling .................................................................................. 87
    4.4.1 Core Modelling ............................................................................................. 88
    4.4.2 Coil Modelling ............................................................................................. 91
    4.4.3 Background Region ..................................................................................... 93
    4.4.4 Model Symmetry and Boundary Conditions ............................................... 93
  4.5 Finite Element Mesh ......................................................................................... 96
    4.5.1 Mesh Control ............................................................................................... 97
4.5.2 Model Optimisations for Mesh Generation ............................................... 100
4.5.3 Error Analysis .................................................................................. 102
4.6 Conclusion ......................................................................................... 106

5 FEA GAP LOSS ANALYSIS ................................................................... 108
5.1 Introduction ....................................................................................... 108
5.2 Inductor Model Configurations ............................................................ 108
5.3 Gap Loss Analysis .............................................................................. 110
5.3.1 Flux Distribution ........................................................................... 111
5.3.2 Eddy Current Distribution ............................................................... 118
5.3.3 Gap Loss Distribution .................................................................... 121
5.3.4 Temperature Distribution ................................................................. 126
5.4 Gap Loss Prediction ............................................................................ 127
5.4.1 Sensitivity Analysis ........................................................................ 128
5.4.2 Gap Loss Prediction ....................................................................... 132
5.5 Conclusion ......................................................................................... 135

6 EXPERIMENTAL VALIDATION OF FEA GAP LOSS MODELLING ......... 137
6.1 Introduction ....................................................................................... 137
6.2 Experimental Set Up .......................................................................... 138
6.2.1 Converter ....................................................................................... 138
6.2.2 Test Inductors ............................................................................... 138
6.2.3 Calibration of Thermal Sensors ....................................................... 141
6.2.4 Calibration of Heat Sink with Fan .................................................. 141
6.2.5 Operating Conditions ..................................................................... 143
6.3 Total Loss Measurements ................................................................. 145
6.3.1 Loss Calculations .......................................................................... 145
LIST OF FIGURES

Figure 1-1. Power train architecture examples for EVs ................................................................. 23
Figure 2-1. Conventional non-isolated boost converter topologies ................................................. 29
Figure 2-2. Bi-directional dual interleaved boost converter with independent inductors. 30
Figure 2-3. Bi-directional dual-interleaved boost converters with IPT ........................................ 32
Figure 2-4. Conventional full bridge isolated DC-DC converter ................................................. 32
Figure 2-5. Component weights in bi-directional dual-interleaved DC-DC boost converter
with IPT [18, 23] .......................................................................................................................... 34
Figure 2-6. Planar inductor [88] ................................................................................................. 39
Figure 2-7. Measured core loss for different duty cycles (LTCC ferrite core, 1.5 MHz, 
100 ºC) [92] ................................................................................................................................ 41
Figure 2-8. Core losses under DC bias conditions (Vitroperm 500F core) [96] ..................... 41
Figure 2-9. Closed type calorimetric method [98] ....................................................................... 42
Figure 2-10. Two-winding core loss measurement method .......................................................... 43
Figure 2-11. Impedance-based core loss measurement [106] ................................................... 45
Figure 2-12. Schematic diagram of gap loss due to in-plane eddy currents ......................... 53
Figure 2-13. Schematic diagrams of modelling of laminated core ........................................... 57
Figure 2-14. Gap loss allocation for thermal FEA [160] ............................................................. 59
Figure 2-15. Reducing the in-plane eddy current loss by attachment of thin ferrite plates 
[136] ............................................................................................................................................. 61
Figure 3-1. Test set up .................................................................................................................... 65
Figure 3-2. Test core with primary and secondary windings .................................................... 66
Figure 3-3. Inductor voltage and current waveforms ................................................................. 67
Figure 3-4. Determination of DC flux density using the magnetisation curve ....................... 69
Figure 3-5. Measured full magnetisation curves at room temperature .................................... 70
Figure 3-6. Measured waveforms at $D_{sw} = 0.75, D = 0.5$ (core: Finemet F3CC0050, $N_p:N_s = 2:1, B_{DC} = 0.64 \text{ T}$) ......................................................................................................................... 72
Figure 3-7. Measured waveforms at $D_{sw} = 0.6, D = 0.2$ (core: Finemet F3CC0050, $N_p:N_s = 2:1, B_{DC} = 0.46 \text{ T}$) ......................................................................................................................... 73
Figure 3-8. Measured B-H loops (Finemet F3CC0050, $B_{DC} = 0.95 \text{ T}$) ............................... 74
Figure 3-9. Measured B-H loops (Vitroperm 500F W156-01, B_{DC} = 0.81 T) ................. 74
Figure 3-10. Measured B-H loops (Vitroperm 500F W156-03, B_{DC} = 0.95 T) ................. 74
Figure 3-11. Measured B-H loops (Vitroperm 500F 3397, B_{DC} = 0.81 T) ...................... 75
Figure 3-12. Finemet F3CC0050 core losses under different operating conditions .......... 76
Figure 3-13. Vitroperm 500F W156-01 core losses under different operating conditions .... 77
Figure 3-14. Vitroperm 500F W156-03 core losses under different operating conditions .... 78
Figure 3-15. Vitroperm 500F 3397 core losses under different operating conditions ...... 79
Figure 3-16. Core loss variation with DC bias conditions ............................................. 81
Figure 4-1. Schematic diagrams of modelling of laminated core ..................................... 84
Figure 4-2. Flow chart of FEA in Opera 3D ................................................................. 87
Figure 4-3. Laminated core types .................................................................................. 88
Figure 4-4. Finemet F3CC0032 core dimensions ............................................................ 89
Figure 4-5. Wound C-core model with volume orientation vectors .................................. 90
Figure 4-6. Inductor model with winding ....................................................................... 92
Figure 4-7. Core symmetry with symmetry planes (background air hidden) ................. 94
Figure 4-8. Boundary conditions for the symmetry planes ............................................. 95
Figure 4-9. 1/8\textsuperscript{th} model to solve .................................................................. 96
Figure 4-10. Finite element types [177] ......................................................................... 97
Figure 4-11. Mesh size transition ............................................................................... 99
Figure 4-12. Core model with additional air volumes filling the gaps and window regions .................................................................................................................. 101
Figure 4-13. Core model coated with a thin layer of air ............................................... 101
Figure 4-14. Meshed models (expanded to show around core regions only) .............. 103
Figure 4-15. Amplitude of flux density in the core at 60 kHz using the two meshing methods (minimum element size: 0.5 mm, maximum element size: 5 mm) .......... 105
Figure 5-1. Legend used in Chapter 5 ......................................................................... 110
Figure 5-2. Flux distributions in the core without air gaps .......................................... 112
Figure 5-3. Flux distributions in the core with air gaps ............................................... 113
Figure 5-4. Cutaway view of perpendicular flux distribution ...................................... 114
Figure 5-5. Perpendicular flux distribution away from the gap edge, \( l_g = 4.4 \) mm .... 114
Figure 5-6. Surface views of perpendicular flux distributions ..................................... 115
Figure 5-7. Perpendicular flux distributions along the central lines on outer and inner core surfaces

Figure 5-8. Perpendicular flux distributions along surface lines parallel to the gap edge, \( dg = 5 \text{mm} \)

Figure 5-9. Cutaway view of eddy current distributions

Figure 5-10. Surface views of eddy current distributions

Figure 5-11. Eddy current distributions along surface lines parallel to the gap edge, \( dg = 5 \text{mm} \)

Figure 5-12. Eddy current distributions along centre lines on core surfaces

Figure 5-13. Gap loss distributions, core view

Figure 5-14. Surface views of gap loss distributions

Figure 5-15. Effect of frequency on gap loss distributions along centre lines on the outer core surface (AC current excitation unchanged, \( lg = 4.4 \text{mm} \))

Figure 5-16. Effect of excitation level on gap loss distributions along centre lines on the outer core surface (frequency: 60 kHz, \( lg = 4.4 \text{mm} \))

Figure 5-17. Effect of gap length on gap loss distributions along centre lines on the core outer surface (frequency: 60 kHz, AC current excitation unchanged)

Figure 5-18. Temperature distribution in the core in free air (Ambient: 30 ºC)

Figure 5-19. Gap loss variation with gap length (\( W_{core} = 30 \text{ mm}, B_m = 0.16 \text{T} \))

Figure 5-20. Gap loss variation with core strip width (\( l_g = 4.4\text{mm}, B_m = 0.17 \text{T} \))

Figure 5-21. Gap loss variation with frequency (\( l_g = 4.4\text{mm}, W_{core} = 30 \text{ mm} \))

Figure 5-22. Gap loss variation with peak flux density (\( f = 60 \text{ kHz}, W_{core} = 30 \text{ mm} \))

Figure 5-23. Gap losses calculated from published and proposed equations (\( l_g = 4.4\text{mm}, B_m = 0.17 \text{T} \))

Figure 5-24. Cutaway views of flux distributions with only one coil (average flux density: 0.14 T)

Figure 5-25. Gap loss distribution with only one coil around the upper core leg

Figure 6-1. Potted inductor on the heat sink cooled by a fan

Figure 6-2. Core with AlN heat spreaders

Figure 6-3. Top view of the heat sink characterisation set up
Figure 6-4. Calculated and measured total inductor losses (gap loss calculated using Approximation 1)................................................................................................................................. 149
Figure 6-5. Calculated and measured total inductor losses (gap loss calculated using Approximation 2)................................................................................................................................. 150
Figure 6-6. Calculated and measured total inductor losses (gap loss calculated using Approximation 3)................................................................................................................................. 151
Figure 6-7. FE thermal model.................................................................................................................. 155
Figure 6-8. Gap loss allocation for thermal FEA .................................................................................. 156
Figure 6-9. Coil model for thermal FEA.............................................................................................. 158
Figure 6-10. Locations of thermal sensors (The other coil not shown) ............................................. 160
Figure 6-11. Measured temperatures in the inductor using plastic spacers, Condition 5, $T_{HS} = 38 ^\circ C$.............................................................................................................................................. 161
Figure 6-12. Measured temperatures in the inductor with AlN heat spreaders, Condition 5, $T_{HS} = 38 ^\circ C$.............................................................................................................................................. 162
Figure 6-13. Thermal FEA results showing the potted inductors, Condition 5, $T_{HS} = 38 ^\circ C$ .............................................................................................................................................. 163
Figure 6-14. Thermal FEA results showing the cores only, Condition 5, $T_{HS} = 38 ^\circ C$... 163
Figure 6-15. FEA and measured steady-state temperature rises in the inductor without heat spreaders (No AlN) and with heat spreaders (AlN), Condition 5, $T_{HS} = 38 ^\circ C$...... 165
Figure 6-16. FEA and measured steady-state temperature rises in the inductor without heat spreaders at 50 kHz and 60 kHz, $T_{HS} = 38 ^\circ C$ ................................................................................................................................. 168
Figure 6-17. FEA and measured steady-state temperature rises at 50 % and 20 % duty ratio, Condition 3 and Condition 8, $T_{HS} = 36 ^\circ C$ ................................................................................................................................. 169
LIST OF TABLES

Table 2-1. Magnetic material properties ..................................................................................... 36
Table 3-1. Nanocrystalline test cores ......................................................................................... 65
Table 3-2. Core permeabilities and inductances at 50 kHz.......................................................... 71
Table 3-3. Test core comparisons............................................................................................... 82
Table 4-1. Equivalent material properties in homogenised Finemet Core............................... 85
Table 4-2. Finemet F3CC0032 core dimensions ....................................................................... 89
Table 4-3. Comparison of tetrahedral and mosaic meshing (frequency: 60 kHz, minimum element size: 0.5 mm, maximum element size: 5 mm)......................................................... 104
Table 4-4. Comparison of field calculations with different mesh size (mosaic meshing, frequency: 60 kHz)................................................................................................................... 106
Table 6-1. Measured temperatures of iced water using different thermal sensors ................. 141
Table 6-2. Characterisation of the thermal resistance of the heat sink assembly at the measurement point ........................................................................................................................................... 143
Table 6-3. Experimental operating conditions .......................................................................... 144
Table 6-4. Average percentage errors in total loss estimations ............................................... 152
Table 6-5. Gap loss allocation ..................................................................................................... 157
Table 6-6. Material thermal conductivities used in the thermal model .................................... 159
Table 6-7. Summary of FEA and measured steady-state temperature rises above the heat sink, Condition 5, $T_{HS} = 38 \, ^\circ C$ ................................................................................................................................. 164
# LIST OF ABBREVIATIONS

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>AlN</td>
<td>Aluminium nitride</td>
</tr>
<tr>
<td>BEVs</td>
<td>Battery electric vehicles</td>
</tr>
<tr>
<td>CUT</td>
<td>Core under test</td>
</tr>
<tr>
<td>DNSE</td>
<td>Double natural Steinmetz extension</td>
</tr>
<tr>
<td>DPF</td>
<td>Displacement factor</td>
</tr>
<tr>
<td>DUT</td>
<td>Device under test</td>
</tr>
<tr>
<td>EMI</td>
<td>Electromagnetic interference</td>
</tr>
<tr>
<td>EVs</td>
<td>Electric vehicles</td>
</tr>
<tr>
<td>FCEVs</td>
<td>Fuel cell electric vehicles</td>
</tr>
<tr>
<td>FE</td>
<td>Finite element</td>
</tr>
<tr>
<td>FEA</td>
<td>Finite element analysis</td>
</tr>
<tr>
<td>FWC</td>
<td>Flux waveform coefficient</td>
</tr>
<tr>
<td>GRP</td>
<td>Glass reinforced plastic</td>
</tr>
<tr>
<td>GSE</td>
<td>General Steinmetz equation</td>
</tr>
<tr>
<td>HEVs</td>
<td>Hybrid electric vehicles</td>
</tr>
<tr>
<td>ICE</td>
<td>Internal combustion engine</td>
</tr>
<tr>
<td>iGSE</td>
<td>Improved general Steinmetz equation</td>
</tr>
<tr>
<td>i²GSE</td>
<td>Improved-improved general Steinmetz equation</td>
</tr>
<tr>
<td>IPT</td>
<td>Interphase transformer</td>
</tr>
<tr>
<td>LTCC</td>
<td>Low temperature Co-fired ceramic ferrite core</td>
</tr>
<tr>
<td>MSE</td>
<td>Modified Steinmetz equation</td>
</tr>
<tr>
<td>NSE</td>
<td>Natural Steinmetz extension</td>
</tr>
<tr>
<td>PCBs</td>
<td>Printed circuit boards</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse-width modulated</td>
</tr>
<tr>
<td>SE</td>
<td>Steinmetz equation</td>
</tr>
<tr>
<td>Si</td>
<td>Silicon</td>
</tr>
<tr>
<td>SiC</td>
<td>Silicon carbide</td>
</tr>
<tr>
<td>SPG</td>
<td>Steinmetz premagnetization graph</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>--------------------------------------------</td>
</tr>
<tr>
<td>THS-II</td>
<td>Toyota Hybrid System II</td>
</tr>
<tr>
<td>VAC</td>
<td>Vaccumschmelze</td>
</tr>
<tr>
<td>WcSE</td>
<td>Waveform-coefficient Steinmetz equation</td>
</tr>
</tbody>
</table>
**LIST OF SYMBOLS**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$A_c$</td>
<td>Effective core cross sectional area</td>
</tr>
<tr>
<td>$A_{coil}$</td>
<td>Conductor cross sectional area</td>
</tr>
<tr>
<td>$B$</td>
<td>Magnetic flux density</td>
</tr>
<tr>
<td>$B_{DC}$</td>
<td>DC magnetic flux density</td>
</tr>
<tr>
<td>$B_m$</td>
<td>Peak AC flux density</td>
</tr>
<tr>
<td>$B_n$</td>
<td>Flux density normal to the lamination plane</td>
</tr>
<tr>
<td>$B_{pk-pk}$</td>
<td>Peak to peak flux density</td>
</tr>
<tr>
<td>$B_{sat}$</td>
<td>Saturation flux density</td>
</tr>
<tr>
<td>$B_t$</td>
<td>Flux density tangential to the lamination plane</td>
</tr>
<tr>
<td>$CG1$</td>
<td>Thermal sensor position, middle of the core leg near the air gap</td>
</tr>
<tr>
<td>$CG2$</td>
<td>Thermal sensor position, middle of the core leg underneath the end of foil winding</td>
</tr>
<tr>
<td>$CoG1$</td>
<td>Thermal sensor position, core inner corner near the air gap</td>
</tr>
<tr>
<td>$CoG2$</td>
<td>Thermal sensor position, core outer corner near the air gap</td>
</tr>
<tr>
<td>$D$</td>
<td>Duty ratio of inductor voltage and current waveforms</td>
</tr>
<tr>
<td>$D_{sw}$</td>
<td>Duty ratio of converter switching waveforms</td>
</tr>
<tr>
<td>$E$</td>
<td>Electrical field strength</td>
</tr>
<tr>
<td>$F$</td>
<td>Core packing factor</td>
</tr>
<tr>
<td>$G$</td>
<td>Gap loss constant</td>
</tr>
<tr>
<td>$G_{in}$</td>
<td>Gap loss allocation block, core gap edges on inner surfaces</td>
</tr>
<tr>
<td>$G_{out}$</td>
<td>Gap loss allocation block, core gap edges on outer surfaces</td>
</tr>
<tr>
<td>$H$</td>
<td>Magnetic field strength</td>
</tr>
<tr>
<td>$H_{core}$</td>
<td>Core leg height</td>
</tr>
<tr>
<td>$H_{DC}$</td>
<td>DC magnetic field strength</td>
</tr>
<tr>
<td>$H_{window}$</td>
<td>Core window height</td>
</tr>
<tr>
<td>$I_{DC}$</td>
<td>DC current</td>
</tr>
<tr>
<td>$I_m$</td>
<td>Peak AC current</td>
</tr>
<tr>
<td>$I_n$</td>
<td>Amplitude of the n-th harmonic component of the inductor current</td>
</tr>
</tbody>
</table>
**In 1**  
Gap loss allocation block, core inner surfaces closer to the air gap

**In 2**  
Gap loss allocation block, core inner surfaces closer to the core corners

**I_{pk-pk}**  
Ripple current in inductor

**J**  
current density

**L**  
Inductance

**L_{core}**  
Core leg length in a core half

**L_{Gin}**  
Gap loss allocation dimension for block Gin

**L_{Gout}**  
Gap loss allocation dimension for block Gout

**L_{in}**  
Input inductor

**L_{SideIn}**  
Gap loss allocation dimension for block Side_In

**L_{SideOut}**  
Gap loss allocation dimension for block Side_Out

**MidCore**  
Gap loss allocation block, middle of core legs

**MLT**  
Mean length per turn of the winding

**N**  
Total number of turns in inductor excitation winding

**N_p**  
Number of turns in primary/excitation winding

**N_s**  
Number of turns in secondary/sense winding

**Out 1**  
Gap loss allocation block, core outer surfaces closer to the air gap

**Out 2**  
Gap loss allocation block, core outer surfaces closer to the core corners

**P_{cu}**  
Copper loss

**P_{cu,AC}**  
AC copper loss

**P_{cu,DC}**  
DC copper loss

**P_e**  
Classical eddy current loss

**P_{ex}**  
Excess loss

**P_g**  
Gap loss

**P_{hy}**  
Hysteresis loss

**P_{t,calculated}**  
Calculated total inductor loss

**P_{t,measured}**  
Measured total inductor loss

**P_v**  
Core loss per unit volume

**R_{AC,n}**  
AC winding resistance at the n-th harmonic
\( R_{DC} \) DC winding resistance
\( R_{\text{HS}} \) Thermal resistance of the heat sink with fan assembly
\( \text{Side\_In} \) Gap loss allocation block, core side edges on inner surfaces
\( \text{Side\_Out} \) Gap loss allocation block, core side edges on outer surfaces
\( T \) Period
\( T_{\text{coil}} \) Thermal sensor position, middle of the top of the coil
\( TG1 \) Thermal sensor position, centre of the outer core strip near the air gap
\( TG2 \) Thermal sensor position, centre of the outer core strip half way down the core leg length
\( T_{\text{HS}} \) Steady-state heat sink temperature (°C)
\( V_{in} \) Converter input voltage
\( V_{i}(t) \) Instantaneous inductor voltage
\( V_{o} \) Converter output voltage
\( W_{\text{coil}} \) Width of foil winding
\( W_{\text{core}} \) Width of core lamination strip

\( d \) Depth of gap loss allocation on core surfaces
\( d_g \) Distance to the core gap edge
\( f \) Frequency of the input inductor waveform
\( f_{eq} \) Equivalent frequency in MSE
\( f_r \) Fundamental frequency in MSE
\( f_{sw} \) Converter switching frequency
\( i_{L}(t) \) Instantaneous inductor current
\( i_p(t) \) Measured instantaneous current in primary/excitation winding
\( k \) Steinmetz parameter
\( k_{Bm} \) Gap loss coefficient, dependency on peak AC flux density
\( k_{c\text{cu}} \) Thermal conductivity of copper
\( k_e \) Classical eddy current loss parameter
\( k_{ex} \) Excess loss parameter
\( k_f \) Gap loss coefficient, dependency on frequency
\( k_g \) Gap loss coefficient
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k_{gap}$</td>
<td>Thermal conductivity of air/potting material between copper turns in the coil</td>
</tr>
<tr>
<td>$k_h$</td>
<td>Hysteresis loss parameter</td>
</tr>
<tr>
<td>$k_i$</td>
<td>Steinmetz parameter in GSE, iGSE and $i^2$GSE</td>
</tr>
<tr>
<td>$k_{lg}$</td>
<td>Gap loss coefficient, dependency on total air gap length</td>
</tr>
<tr>
<td>$k_n$</td>
<td>Equivalent coil thermal conductivity in the direction through the turns</td>
</tr>
<tr>
<td>$k_{NSE}$</td>
<td>Steinmetz parameter in NSE</td>
</tr>
<tr>
<td>$k_t$</td>
<td>Equivalent coil thermal conductivity in the direction along the turns</td>
</tr>
<tr>
<td>$k_{tape}$</td>
<td>Thermal conductivity of Kapton tape</td>
</tr>
<tr>
<td>$k_{wcore}$</td>
<td>Gap loss coefficient, dependency on core strip width</td>
</tr>
<tr>
<td>$l_m$</td>
<td>Mean magnetic path length</td>
</tr>
<tr>
<td>$l_g$</td>
<td>Total air gap length</td>
</tr>
<tr>
<td>$t_{coil}$</td>
<td>Thickness of coil</td>
</tr>
<tr>
<td>$t_{cu}$</td>
<td>Thickness of copper foil</td>
</tr>
<tr>
<td>$t_{cu,coil}$</td>
<td>Total thickness of copper strips in a coil</td>
</tr>
<tr>
<td>$t_{gap}$</td>
<td>Total thickness of gaps between copper turns in a coil</td>
</tr>
<tr>
<td>$t_l$</td>
<td>Thickness of lamination</td>
</tr>
<tr>
<td>$t_{tape}$</td>
<td>Total thickness of Kapton tape in a coil</td>
</tr>
<tr>
<td>$v_s(t)$</td>
<td>Measured instantaneous voltage in secondary/sense winding</td>
</tr>
<tr>
<td>$v_{sw}(t)$</td>
<td>Measured instantaneous transistor gate voltage</td>
</tr>
<tr>
<td>$\Delta B$</td>
<td>AC magnetic flux density</td>
</tr>
<tr>
<td>$\Delta H$</td>
<td>AC magnetic field strength</td>
</tr>
<tr>
<td>$\Delta T_{HS}$</td>
<td>Measured steady-state temperature rise in heat sink with fan (°C)</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Steinmetz parameter</td>
</tr>
<tr>
<td>$\beta$</td>
<td>Steinmetz parameter</td>
</tr>
<tr>
<td>$\delta$</td>
<td>Skin depth of core lamination material</td>
</tr>
<tr>
<td>$\delta_{cu}(n)$</td>
<td>Skin depth of copper at the n-th harmonic</td>
</tr>
<tr>
<td>$\delta_e$</td>
<td>Equivalent skin depth of homogeneous core model</td>
</tr>
</tbody>
</table>
ε  Percentage error in total inductor loss estimation
η  Stretch factor of winding
θ  Phase shift between inductor voltage and current waveforms
ϑ  Phase angle of sinusoidal waveform
μ₀  Permeability in vacuum
μₘ  Relative permeability of lamination material in rolling direction
μₙ  Equivalent permeability in the direction through the lamination planes
μᵣ  Relative permeability
μᵣ  Equivalent permeability in the direction along the lamination planes
ξ(ₙ)  Porosity factor of winding at the n-th harmonic
ρₜˌcu  Resistivity of copper
σₘ  Electrical conductivity of lamination material in rolling direction
σₙ  Equivalent electrical conductivity in the direction through the lamination planes
σᵣ  Equivalent electrical conductivity in the direction along the lamination planes
φ  Phase shift error
ABSTRACT

The University of Manchester

Yiren Wang

A thesis submitted for the degree of Doctor of Philosophy

Modelling and Characterisation of Losses in Nanocrystalline Cores

September 2015

Increasing the power density of the DC-DC converters requires the size and weight of the magnetic components, such as inductors and transformers, to be reduced. In this thesis, the losses in nanocrystalline inductor cores are characterised and analysed, including the traditional core loss and the gap loss caused by the air gap fringing flux. The loss calculations will form a basis for the design and optimisation of high power inductors for DC-DC converters for EV applications.

This thesis first characterises experimentally the core losses in four nanocrystalline cores over a range of operating conditions that are representative of those encountered in typical high power converter applications, including non-sinusoidal waveforms and DC bias conditions. The core losses are assessed by the measured B-H loops and are characterised as a function of DC flux density, showing that for a fixed AC induction level, the losses can vary by almost an order of magnitude as the DC bias increases and the duty ratio moves away from 0.5. The results provide a more complete picture of the core loss variations with both DC and AC magnetisations than is available in manufacturers’ data sheets.

An electromagnetic finite element (FE) model is used to examine the gap loss that occurs in finely laminated nanocrystalline cores under high frequency operation. The loss is significant in the design example, contributing to almost half of the total inductor loss, and the gap loss is highly concentrated in the region of the air gap. The dependence of the gap loss on key inductor design parameters and operating conditions is also explored. An empirical equation is derived to provide a design-oriented basis for estimating gap losses.

Thermal finite element analysis is used to estimate the temperature rise and identify the hot spot in a nanocrystalline inductor encapsulated in an aluminium case. The temperature distribution in the core largely corresponds to the non-uniform distribution of the gap loss. The thermal FEA can also be used to evaluate different thermal management methods to optimise the design for a more compact component.

The FE modelling of gap loss and the thermal predictions are validated experimentally on a foil-wound Finemet inductor, showing good agreement between the predictions and measurements under various operating conditions.
DECLARATION

No portion of the work referred to in this thesis has been submitted in support of an application for another degree of qualification of this or any other university or other institution of learning.
COPYRIGHT STATEMENT

i. The author of this thesis (including any appendices and/or schedules to this thesis) owns certain copyright or related rights in it (the “Copyright”) and she has given The University of Manchester certain rights to use such Copyright, including for administrative purposes.

ii. Copies of this thesis, either in full or in extracts and whether in hard or electronic copy, may be made only in accordance with the Copyright, Designs and Patents Act 1988 (as amended) and regulations issued under it or, where appropriate, in accordance with licensing agreements which the University has from time to time. This page must form part of any such copies made.

iii. The ownership of certain Copyright, patents, designs, trade marks and other intellectual property (the “Intellectual Property”) and any reproductions of copyright works in the thesis, for example graphs and tables (“Reproductions”), which may be described in this thesis, may not be owned by the author and may be owned by third parties. Such Intellectual Property and Reproductions cannot and must not be made available for use without the prior written permission of the owner(s) of the relevant Intellectual Property and/or Reproductions.

iv. Further information on the conditions under which disclosure, publication and commercialisation of this thesis, the Copyright and any Intellectual Property and/or Reproductions described in it may take place is available in the University IP Policy (see http://documents.manchester.ac.uk/DocuInfo.aspx?DocID=487), in any relevant Thesis restriction declarations deposited in the University Library, The University Library’s regulations (see http://www.manchester.ac.uk/library/aboutus/regulations) and in The University’s policy on Presentation of Theses.
ACKNOWLEDGEMENTS

I would like to express my sincere gratitude to my supervisor, Professor Andrew J. Forsyth, for his constant support, guidance and encouragement throughout my PhD study, and the invaluable suggestions during the writing of this thesis.

I am deeply grateful to the EPSRC VESI project for the financial support during this research project.

Special thanks to Dr Gerardo Calderon-Lopez for his help, experience and advice. Thank you for spending countless hours to help me in the labs and discuss the results with me.

I would like to thank all my fellow students in the Power Conversion Group for the useful discussions and encouragement. I also would like to thank my friends from North China Electrical Power University for the friendship and all the fun we had since we came to the UK.

To my parents, thank you for your unconditional love and support. I love you so much and hope I’ve made you proud. I also want to express my appreciation to my husband, Linwei Chen, for the love, patience and all the moments we had together. I couldn’t have done this without you.
CHAPTER 1

INTRODUCTION

1.1 Background

In the last few decades, there has been growing concern over air pollution and global warming due to greenhouse gas emissions. The EU has been working to cut greenhouse gases substantially, aiming to achieve a 20% reduction in the emissions compared to 1990 levels by 2020 and a reduction of at least 80% by 2050 [1]. For over a century, internal combustion engine (ICE) vehicles have dominated road transport, which alone contributes about one-fifth of the EU’s total greenhouse gas emissions [1, 2]. The emission reduction targets along with the depletion of fossil fuels have accelerated the development of clean and energy-efficient vehicular technologies [3].

Electric vehicles (EVs), which have very low, or zero, emissions, are one of the solutions to tackle the climate change concerns and the energy crisis. The development of EVs has been in continuous progress in the past few decades. In addition to the main traction drive, traditional mechanical, hydraulic and pneumatic systems on board vehicles are gradually being converted to electrical systems [4, 5] to improve efficiency, reliability and controllability, resulting in a rapid growth in vehicle electrical systems including batteries, supercapacitors, electric motors, power electronic converters and controllers.

The steady increase in vehicle electrical load is driving the electrical system voltages and power to higher levels. The voltage was from 6 V to 12 V several decades ago mainly for auxiliary loads [5, 6]. High power traction drives tend to operate at several hundred volts whilst intermediate voltage levels [7, 8], such as 48 V, are currently being proposed for some auxiliary loads [9]. The power capability of an automotive electrical power train is typically tens of kW. To accelerate the development of EVs, step-change advances are needed in the size, weight and cost of on-board power electronic devices to bring cost-effective products to the market.
Electric vehicles broadly fall in to three categories: battery EVs (BEVs), hybrid EVs (HEVs) and fuel cell EVs (FCEVs). BEVs are powered by high-capacity battery packs and use purely electric power. HEVs combine the conventional ICE with electrical propulsion and have achieved a continuous growth in the worldwide market in the past decade [10]. FCEVs use the chemical energy stored in the fuel for example hydrogen to generate electricity directly. The energy systems used in EVs are required to store high energy to extend the driving range, and to meet the sudden high power demand of the vehicles during starting or acceleration, supercapacitors or flywheels that have high specific power capability are often used in conjunction with other energy systems.

There are a large number of configurations established for EV power trains [11-16]. Figure 1-1 illustrates two examples of the on-board electrical power trains. Alternative topologies use different combinations of energy devices and power electronic circuits.

![Power train topology for FCEVs](image)

![Power train topology for BEVs](image)

Figure 1-1. Power train architecture examples for EVs

The typical voltage for fuel cells is between 70 V to 120 V and supercapacitor output voltages must usually vary over a 2:1 range to use the capacity of the capacitors...
efficiently. Depending in their size, supercapacitor banks could be rated up to 200 V [17]. In contrast, battery banks are normally rated around 200 V to 400 V [18, 19]. DC-DC converters are therefore required to interface the various elements and control the power flow in the EV power train by boosting the low voltage levels at the energy source and storage devices to higher levels, typically around 600 V, at the DC-link of the traction drives for high-efficiency operation [7, 18]. The power levels are typically in the range 30 kW to 100 kW [7, 12, 20]. In Figure 1-1a, the fuel cell is interfaced via a uni-directional DC-DC converter, whilst bi-directional operation is required for the converters interfacing with the batteries and supercapacitors, Figure 1-1b, so that the excess energy during braking can flow back through the converter to recharge the batteries and supercapacitors. Other design requirements for the DC-DC converters include high power, high efficiency, high reliability, reduced size and weight, low cost, low electromagnetic interference (EMI) and straightforward control [13, 21].

One of the obstacles to increase the power density of the DC-DC converters is the size of the magnetic components, inductors and transformers, which are responsible for about 30-50% of the overall converter mass [18]. One of the main factors that limits the reduction in size of these components is the maximum permissible temperature in the core and windings, and this is aggravated by the high temperature environment in many automotive applications [22].

One obvious way to reduce the size of the magnetic components is to increase the switching frequency since this leads to a requirement for lower inductance values, reduced core area and fewer turns, but this must be traded against the potential increase in semiconductor switching losses. Typically, with silicon devices, the converters are operated at up to 20 kHz, but with the use of new power devices such as high power silicon carbide (SiC) semiconductors a switching frequency above 100 kHz is feasible [18, 23, 24]. However, at higher frequencies the AC core and winding losses in magnetic components tend to increase, potentially resulting in a larger temperature rise due to the smaller surface area for heat transfer from a reduced component size. Therefore, a key design objective is to limit the hot spot temperature in the magnetic components, which
requires high flux cores with low loss characteristics, low winding losses, a capability to predict losses accurately and also an effective thermal management system to remove the heat.

Furthermore, to satisfy the high power levels at the traction drives, the low voltage energy sources need to deliver large DC currents. Air gaps are required in the converter’s input inductors to avoid saturating the cores, but the gaps will introduce additional losses in the core and winding [25, 26]. Accurate estimation of losses is essential for the design and optimisation of the magnetic components and knowing the distribution of losses will help to identify the component’s hot spot.

1.2 Objectives

The overall aim of the research in this thesis is to understand and analyse the losses in nanocrystalline cores. These cores are of particular interest since they combine a high saturation flux density of around 1.2 T (compare with 0.3 T in ferrite), with hysteresis losses that are comparable with those in ferrite. However, they have the disadvantage of a high electrical conductivity which can result in high eddy current losses around the air gap due to the fringe field under high frequency operation, which is not well understood at the moment. The loss calculations will form a basis for the design and optimisation of high power inductors for DC-DC converters for EV applications.

The key objectives are summarised below:

- To characterise experimentally the core loss in nanocrystalline cores under typical converter operating conditions, including DC bias, non-sinusoidal and asymmetric waveform shapes.
- To develop a finite element (FE) electromagnetic model to investigate the gap loss caused by the air gap fringing field and its distribution in the core.
- To identify and predict the hotspot temperature in the core using FE thermal analysis.
- To validate the loss calculations and thermal predictions by temperature measurements in an inductor.
1.3 Scope of This Thesis

The thesis consists of seven chapters. The present Chapter 1 briefly introduces EV applications and the requirements of the on-board DC-DC converters. The key objectives of the research are then identified. The remaining contents of the thesis are organised as follows.

Chapter 2 provides a broad literature survey of magnetic components for DC-DC converters, focusing on the loss mechanisms in inductor cores. The DC-DC converter topologies and the requirements placed on the magnetic components are discussed. The different magnetic materials and component types are discussed. The modelling techniques for magnetic components are summarised, followed by a review of the design and optimisation techniques for magnetic components.

Chapter 3 presents the characterisation of core losses in four nanocrystalline cores. The core loss variations under representative converter operating conditions are investigated.

Chapter 4 develops a 3D model of a gapped inductor to investigate the gap loss using FE electromagnetic analysis. The main challenges are the modelling of the finely laminated core structure and the control of the 3D finite element mesh.

Chapter 5 provides the FEA results of gap loss calculations using the model developed in Chapter 4. The distribution of magnetic field, in particular the perpendicular components of the air gap fringing flux, and the distribution of induced eddy currents and the associated losses are investigated. A sensitivity analysis is performed to study the dependency of gap loss on some key inductor design parameters. An empirical equation is derived from the results to provide a design-oriented estimation of the gap loss in nanocrystalline inductor cores.

Chapter 6 validates the loss characterisation in Chapter 3 and the gap loss calculations in Chapter 5 by comparing the estimated total inductor losses with measurements. A FE
thermal analysis using the estimated losses is undertaken to calculate the steady-state temperature rise in an inductor encapsulated in an aluminium heat sink. The estimated temperature distribution in the component is also validated via experimental measurements.

Chapter 7 provides the conclusions of the research work, identifies the key contributions, and describes potential areas for future research.
CHAPTER 2

LITERATURE REVIEW

2.1 Introduction

A large number of papers have been published on magnetic components for DC-DC converters for automotive applications, aiming to reduce the size and weight of these components. This chapter presents a review of the literature relating to the research areas of this thesis, particularly focusing on the losses in inductor cores.

Section 2.2 reviews the DC-DC converter topologies for EV applications. In Section 2.3, the magnetic components for DC-DC converters are discussed, including the core materials and winding techniques. Section 2.4 presents a review of loss mechanisms in magnetic components, the gap loss caused by air gap fringing flux being the main research topic of this thesis. In Section 2.5, the modelling techniques for magnetic components are described. Finally, the published magnetic component design and optimisation methods are discussed in Section 2.6.

2.2 DC-DC Converter Topologies

A large number of topologies and techniques to improve the power density and efficiency of DC-DC converters for EV applications have been published in the literature. The most well-known converter configuration, shown in Figure 2-1a, is a basic non-isolated boost topology to step up the output voltage from the energy source to feed the high voltage DC-link of the traction system [19-22, 27-29]. By changing the switch configurations, the boost converter can be given a bi-directional capability to interface with rechargeable energy storage devices, Figure 2-1b.
In [28], the conventional bi-directional boost converter is compared with other non-isolated bi-directional converter topologies, such as the cascade buck-boost converter, Cuk, SEPIC and Luo converters. The authors show that the conventional bi-directional boost converter is more efficient than the alternatives because it requires fewer components and the device voltage and current stresses tend to be lower. The major disadvantage of this simple topology is the high current ripple in the high voltage capacitor C2 that increases the filtering requirements.

Many converter designs use a multiphase structure with interleaved operation to improve the performance at high power levels [19, 26, 30-41]. Interleaved converters comprise several converter units in parallel but operated with a phase-shifted pattern. There are several advantages of the interleaved topology. First, the input and output current ripple is reduced. Second, the interleaving operation effectively increases the ripple frequency at the input and output ports, downsizing the filter capacitors and reducing the potential EMI problems. What’s more, the multiple phases share the current at the low voltage port, which reduces the stresses on the windings and semiconductors in each phase, allowing better thermal management and a higher switching frequency to be used [26]. For example, Figure 2-2 shows a bi-directional dual interleaved boost converter with individual inductors and the current waveforms. The two converter cells are switched with a 180° phase shift and the ripple frequency at the input and output is doubled and reduced in amplitude.
Integrated magnetic components have been considered in the literature [19, 26, 31, 34, 38, 39, 42-47] to increase the power density of interleaved converters. The inductors of the interleaved phases may be integrated through magnetic coupling to reduce the total number of components and their size, weight and losses, and they can be close-coupled [42, 48], loose-coupled [49] or integrated winding coupled [45].

In [34], the authors show that by using coupled inductors the total inductor size can be reduced by almost half compared to a design with uncoupled inductors in a 1 kW, four-phase interleaved, bi-directional, 14 V to 42 V DC-DC converter. Reference [50] compares the performance of 56 kW, four-channel interleaved converters using
independent phases and inductively coupled phases, showing that the design with coupled magnetics almost doubles the power density of that of the independent phase design (87 kW/l versus 44.2 kW/l).

A volume comparison of four 1 kW, 50 kHz, non-isolated DC-DC converter topologies is carried out in [29], including a conventional single-phase boost converter, a two-phase interleaved boost converter with independent inductors, a two-phase interleaved boost converter with loose-coupled inductors and a two-phase interleaved boost converter with integrated winding coupled inductors. The authors conclude that interleaving phases and magnetic coupling can effectively increase the efficiency and power density of converters and it is a suitable option for EV applications.

Integrating the inductors into a single component sometimes requires custom core shapes and may include multiple core materials. For example, an EƗƎ-shape core structure is proposed in [26] for a dual interleaving converter. The authors in [44] uses a IM-shape composite core structure made of powdered iron and ferrite core pieces to integrate the inductors of a three phase interleaved converter.

Instead of integrating all the inductors onto a single core, some authors use an interphase transformer (IPT) between phases along with an input inductor [17, 19, 42, 48, 51]. A bi-directional dual interleaved boost converter with IPT is shown in Figure 2-3. The IPT is a transformer-like component with inversely coupled phases to cancel the DC flux. The frequency doubling effect allows the size of the input inductor, $L_{\text{in}}$, to be reduced. This provides more flexibility in the design and optimisation where the input inductor and IPT can be designed separately using different cores based on their individual requirements [52]. Also, this topology can be efficient and very easy to control, making it attractive for power management in EV power trains. Efficiencies of over 97 % have been reported in several converter prototypes operated at 40-60 kW based on this configuration [18, 23, 51].
The main drawback of non-isolated topologies is the limited voltage conversion ratio caused by the inductors’ stray resistance and device losses at high duty ratios. The maximum step-up ratio is approximately four [30]. When a high step-up ratio is required, isolated DC-DC converters are preferred which utilise high frequency transformers for electrical isolation [6, 30]. The full bridge DC-DC converter, shown in Figure 2-4, is one of the most frequently used isolated converter topologies for multi-kW applications [33, 53].
2.3 Magnetic Components for DC-DC converters

Magnetic components are one of the major concerns in the design of DC-DC converter as they are significant loss and volume contributors [54]. Therefore, a common design objective is to reduce their size, weight and losses. There are several magnetic devices that are commonly used in DC-DC converters, for example DC inductors, AC inductors, transformers and coupled inductors [55]. Since they are operated in different ways they must therefore be designed accordingly.

One obvious way to reduce the size of the magnetic components is to increase the switching frequency, but this is usually limited by the capability of the semiconductor devices as the switching losses tend to be higher under high frequency operation. Therefore, reducing the volume of the magnetic components requires advanced converter topologies with fast switching semiconductors and careful selections of core materials, core and winding structures as well as effective cooling methods [44].

The aforementioned dual-interleaved boost converter with IPT is a good candidate topology for high power density magnetic designs. In [18] and [23], the weights of the three 60 kW dual-interleaved converters with IPT using silicon (Si) and silicon carbide (SiC) switching devices are presented and the total converter weights are broken down to show the contribution of each component to the overall mass, as shown in Figure 2-5. The SiC converters, shown in Figure 2-5 b and c, are about 2 kg lighter than the Si prototype, shown in Figure 2-5a, where 1.2 kg of reduction is mainly due to the increased operating frequency, 75 kHz instead of 25 kHz. The magnetic devices, a DC inductor and the IPT, account for about 50% of the total converter weight. The importance of making further weight reductions in these components is highlighted if the overall converter size is to be made even smaller.

The input inductor in this converter topology normally carries a large DC current under high power operations. The DC bias increases the maximum peak flux density experienced by the inductor core and therefore air gaps are required to avoid saturation.
The following review of magnetic components will be focused on the DC inductor for high power high frequency applications.

a. Si IGBT converter, 25 kHz, 60 kW, 6.9 kg
b. SiC BJT converter, 75 kHz, 60 kW, 5 kg
c. SiC MOSFET converter, 75 kHz, 60 kW, 5.2 kg

Figure 2-5. Component weights in bi-directional dual-interleaved DC-DC boost converter with IPT [18, 23]
2.3.1 Magnetic Core Materials

Ideal magnetic materials for optimal magnetic designs should have high saturation flux density, low power loss, high resistivity, high operating temperature capability and low cost. The authors of [56-61] have presented a series of publications to compare the magnetic materials for high-current high-frequency inductors in DC-DC converters. They identify that the selection of magnetic materials for high power inductors also needs to consider the practical effects, including the frequency, saturation flux density, DC bias, current ripple, air gaps, duty cycle and cooling method. Reference [62] compares the magnetic materials at high temperatures, showing that the magnetic properties can also vary significantly with temperatures, for example, the hysteresis loss in a 6.5 % Silicon steel specimen is reduced by around 20% as the temperature increases from room temperature to 200 °C.

Table 2-1 summarises the key features of several core materials currently being employed in DC-DC converters, including ferrite, powder cores, powder iron, amorphous metals and nanocrystalline materials [63-70].

The iron-powder cores have the highest losses and their properties and loss characteristics depend on the mixes of the iron content. Ferrite materials have the lowest loss and offer other advantages such as low cost and very high resistivity to reduce eddy current losses. However, ferrites have the lowest saturation flux density, $B_{\text{sat}}$, only around 0.3-0.5 T, which limits their uses in high power applications because large air gaps are required to prevent saturation under DC bias conditions [71]. Also, ferrite magnetic properties vary with temperature [63].

Silicon steel cores have the highest saturation flux density, followed by some powder cores, for example High Flux [64] and Mega Flux [65], and amorphous metal cores. Nanocrystalline material is a recent innovation formed by annealing amorphous metal and offers a high saturation flux density of over 1 T, low losses at high frequency even comparable to those in ferrites [72, 73] and high thermal stability [74].
Table 2-1. Magnetic material properties

<table>
<thead>
<tr>
<th>Material Type</th>
<th>Manufacturer</th>
<th>Material</th>
<th>( B_{\text{sat}} ) (T) (@ 25^\circ\text{C} )</th>
<th>( \mu_r ) @20kHz</th>
<th>Continuous Operating Temperature (°C)</th>
<th>Resistivity (μΩ·m)</th>
<th>Thermal Conductivity (W/mK)</th>
<th>Density (g/cm(^3))</th>
<th>Core Loss (kW/m(^3)) @ 0.1T, 20kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ferrite</td>
<td>Ferroxcube [63]</td>
<td>3C93</td>
<td>0.52</td>
<td>1800</td>
<td>140</td>
<td>5 × 10(^6)</td>
<td>3.5-5</td>
<td>4.8</td>
<td>5</td>
</tr>
<tr>
<td>Powder Core</td>
<td>Magnetics [64]</td>
<td>Molypermalloy (MPP)</td>
<td>0.75</td>
<td>14-550</td>
<td>200</td>
<td>-</td>
<td>-</td>
<td>8.2</td>
<td>45</td>
</tr>
<tr>
<td>Powder Core</td>
<td>Magnetics [64]</td>
<td>KoolMu 26</td>
<td>1.05</td>
<td>26-125</td>
<td>200</td>
<td>-</td>
<td>-</td>
<td>6.8</td>
<td>83</td>
</tr>
<tr>
<td>Powder Core</td>
<td>Magnetics [64]</td>
<td>High Flux</td>
<td>1.5</td>
<td>14-160</td>
<td>200</td>
<td>-</td>
<td>-</td>
<td>7.7</td>
<td>116</td>
</tr>
<tr>
<td>Powder Core</td>
<td>Chang Sung Corp. [65]</td>
<td>Mega Flux</td>
<td>1.6</td>
<td>26-90</td>
<td>200</td>
<td>-</td>
<td>11.4</td>
<td>6.8</td>
<td>186</td>
</tr>
<tr>
<td>Powder Iron</td>
<td>Micrometals [66]</td>
<td>Mix-26</td>
<td>1.38</td>
<td>75</td>
<td>&lt;75</td>
<td>-</td>
<td>4.2</td>
<td>7</td>
<td>630</td>
</tr>
<tr>
<td>Powder Iron</td>
<td>Micrometals [66]</td>
<td>Mix-30</td>
<td>1.38</td>
<td>22</td>
<td>&lt;75</td>
<td>-</td>
<td>2</td>
<td>6</td>
<td>835</td>
</tr>
<tr>
<td>Amorphous</td>
<td>Metglas [67]</td>
<td>2605SA1</td>
<td>1.56</td>
<td>600</td>
<td>150*</td>
<td>1.37</td>
<td>10**</td>
<td>7.18</td>
<td>70</td>
</tr>
<tr>
<td>Amorphous</td>
<td>Metglas [67]</td>
<td>2605SA3</td>
<td>1.41</td>
<td>35000</td>
<td>150*</td>
<td>1.38</td>
<td>10**</td>
<td>7.29</td>
<td>17</td>
</tr>
<tr>
<td>Silicon Steel</td>
<td>JFE Steel [68]</td>
<td>10JNHF600</td>
<td>1.88</td>
<td>600</td>
<td>150*</td>
<td>0.82</td>
<td>18.6**</td>
<td>7.53</td>
<td>150</td>
</tr>
<tr>
<td>Silicon Steel</td>
<td>JFE Steel [68]</td>
<td>10JNEX900</td>
<td>1.8</td>
<td>900</td>
<td>150*</td>
<td>0.82</td>
<td>18.6**</td>
<td>7.49</td>
<td>180</td>
</tr>
<tr>
<td>Nanocrystalline</td>
<td>Vaccumschmelze [69]</td>
<td>Vitroperm 500F</td>
<td>1.2</td>
<td>13200</td>
<td>120*</td>
<td>1.15</td>
<td>10**</td>
<td>7.3</td>
<td>5</td>
</tr>
<tr>
<td>Nanocrystalline</td>
<td>Hitachi Metals [70]</td>
<td>Finemet FT-3M</td>
<td>1.23</td>
<td>15000</td>
<td>155*</td>
<td>1.2</td>
<td>10**</td>
<td>7.3</td>
<td>5</td>
</tr>
</tbody>
</table>

*limited by lamination epoxy; **Thermal conductivity along laminations
Powder cores have distributed air gaps and are normally manufactured into toroid shapes which can make it difficult to wind with high current windings. The permeability is not adjustable after manufacturing. Powder cores have isotropic thermal conductivity but they are not recommended for conductive cooling as the stray flux due to the distributed gap can set up eddy currents in nearby electrically conductive heat sinks [75]. One alternative cooling method is to submerge the component in dielectric oil [61].

Silicon steel cores are stamped into laminations then stacked to build up the core, while amorphous metal and nanocrystalline materials are tape wound [61], typically as C-cores, then cut to provide discrete air gaps and an adjustable permeability. This makes the placement of winding easy but the laminated structure results in anisotropic thermal properties and the maximum operating temperatures are limited by the thermal rating of the insulation between laminations. These core materials can be conduction cooled by metallic cold plates but optimal thermal paths should be chosen as the thermal transfer is poor through the laminations [58].

Reference [76] compares two core materials that have been used in the DC-DC converters for Toyota hybrid vehicles. They are the silicon steel core, JNEX, one from the Super Core range from JFE Steel [68] that is used in the second generation Toyota Hybrid System II (THS-II) and the powder core, Mega Flux, from Chang Sung Corporation [65] which is used in the third generation of THS-II. Test inductors are fabricated using the cores and encapsulated in aluminium cases, and then tested on a 15 kHz, 40 kW, DC-DC converter. The results show that the Mega Flux core has lower losses and better thermal performance than the JNEX core.

In [75], the design trade-offs between the core, winding and cooling method are discussed and a KoolMu, distributed gap, powder core and a Finemet nanocrystalline core have been tested on a 15 kHz, 30 kW, single-phase DC-DC boost converter. The authors recommend the distributed gap material such as KoolMu for low power density applications when high temperature can be tolerated and the nanocrystalline material for high power density designs with stringent thermal requirements.
Nanocrystalline materials, such as Finemet and Vitroperm, have recently found themselves increasingly used in the converter applications to achieve size reduction, for example in [18, 23, 72, 77-80]. Besides their superior soft magnetic properties, nanocrystalline cores can be custom-designed to allow optimisation of core shapes to meet specific design requirements [77].

### 2.3.2 Winding Techniques

Apart from using low loss, high flux density magnetic cores, reducing the size and volume of high-power, high-frequency inductors requires advanced winding technologies that offer high current capability, low resistances (DC and AC resistances) to reduce losses, high filling factor and good thermal performance.

The windings are normally made from copper for its high conductivity while in [81] aluminium windings are considered due to its lower weight and cost. Conventional winding types include litz wire windings, solid round or square conductors and copper foil/strip windings, and among them litz-wire and copper foil windings are most commonly used.

Litz wires have been widely used in high frequency applications to reduce the AC copper losses due to skin and proximity effects [72]. They are easy to wind and can be adapted to various core window geometries [82]. However, they may not be suitable for some high power applications where large DC currents are required in the windings, since litz wires have poor filling factors, high cost and poor heat transfer characteristics [83].

Copper foil windings are wound with multiple layers of thin copper strip. They are relatively easy to wind and can well suit the window area of C-cores. The width of the copper strips is normally chosen to fit the width of the core window and a high window utilization factor can be achieved due to the high filling factor [84]. Foil windings can have large copper cross sectional area and low DC resistance, and therefore they provide
a good solution for high current applications. Another attractive feature of foil windings is the large contact surface area that is available for heat transfer [54, 83, 85]. Due to these advantages, foil windings have been widely used in high-current, high-frequency inductor designs [18, 58, 86]. However, foil windings may suffer from eddy current losses at high frequencies, especially with the presence of air gaps. The air gap fringing flux may set up eddy currents in the copper strips so it is important that the windings are kept a sufficient distance away from the core gaps [75]. This effect is reviewed in Section 2.4.2.

In recent years, planar magnetic components have attracted increasing attention. The planar winding allows the magnetic component to be embedded into printed circuit boards (PCBs), and it can be used to achieve low profile designs when compared with conventional wire-wound windings [87]. Planar magnetic component also offers a large thermal transfer surface area, low cost and good fabrication reproducibility [88]. Planar ferrite cores are commonly used together with the planar winding. They have reduced height, but the flattened structure increases the component footprint, as can be seen in Figure 2-6 [88]. In addition, planar windings tend to have a low filling factor and a relatively low current handling capability [87].
2.4 Loss Mechanisms in Magnetic Components

The size and weight of magnetic components is normally limited by thermal considerations. Therefore, the design and optimisation of the components rely on accurate predictions of losses. There are mainly two sources of losses in magnetic devices, core loss and winding loss. However, in laminated cores another loss mechanism is often significant, that is the gap loss caused by the flux fringing around the gap leaving and re-entering the core in a direction of high loss, namely through the plane of the laminations [89].

2.4.1 Core loss

Accurate prediction of the core loss is potentially a complex task because the core loss is affected by many factors, for example the frequency and amplitude of the AC excitation, the waveform shape and the DC bias, however datasheets usually only provide information at a limited number of conditions and with sinusoidal excitation.

In modern power electronics, square voltage waveforms are commonly used. In [90], a lower core loss is observed under symmetrical square-wave voltage excitation compared with the loss induced by a sinusoidal waveform even for the same peak flux density and the same frequency. The duty ratio of the excitation waveform will also influence the core loss [91]. The core losses with sinusoidal and triangular flux waveforms are plotted against the duty ratio of the waveform in [92] for a low temperature Co-fired ceramic (LTCC) ferrite core, as shown in Figure 2-7, showing the core loss is lowest at 50 % duty ratio and the loss increases when the duty ratio is smaller or larger than 50 %.

The DC bias adds to the complexity of core loss predictions. Although DC flux does not generate loss itself, it affects the magnetisation of the core materials [93-95]. Figure 2-8 shows the measured core losses against DC bias conditions using a Vitroperm 500F core [96]. The core losses are seen to be increased with increasing DC bias.
Furthermore, the core loss can be temperature dependent. Ferrite core loss has a strong dependence on the operating temperature and ferrite cores are often designed to operate at a specific temperature at which the power loss is minimum, typically in the range 80 °C - 150 °C [61]. While nanocrystalline core loss is a relatively weak function of temperature.
The test results presented in [72] show that Finemet material has small loss variations with temperature in the range 25 °C - 150 °C.

An additional factor that may affect core losses is the preparation of the cores, for example, it is reported in [72] that the nanocrystalline cut core has about twice the loss of that in an un-cut core and this is attributed to the material being damaged during the cutting process and the insulation between the lamination layers being degraded.

### 2.4.1.1 Core loss measurement techniques

This section reviews the measurement techniques and the existing core loss models that are described in the literature. The core loss measurement techniques generally fall into two categories: thermal or electrical [97-99].

The thermal method, or calorimetric method, normally uses a thermally isolated chamber that may contain dielectric fluid to enhance thermal uniformity. The temperature difference between the inlet and outlet coolant of the chamber is measured to determine the power loss in the device under test (DUT), Figure 2-9.

![Figure 2-9. Closed type calorimetric method [98]](image)
The thermal loss measurement method is universal to all kinds of power loss measurements and can be used to measure the losses under arbitrary excitation or operating conditions, but it requires a special chamber and the measurement is time consuming [98]. The major drawback is that the thermal method measures not only the core loss but also the winding loss and other losses in the chamber. Besides, this method may not be suitable for core materials like ferrites since the loss is temperature dependent.

The electrical methods are based on the measurements of voltage and current, and are fast and easy to perform and reproduce [98]. The most popular electrical core loss measurement method is the two-winding method, also called the four-wire method, where two closely coupled windings are placed around the core, a primary or excitation winding and a secondary or sense winding, as shown in Figure 2-10. The current in the primary winding and the voltage across the secondary winding are measured and the core loss is assessed by integrating the product of voltage and current waveforms. This method has been widely used [72, 93, 96, 100-104]. Its advantages are that it is not limited to sinusoidal waveforms and also the winding loss is not included. Furthermore, with the measured voltage and current waveforms, the B-H loop can be plotted to provide additional material information such as permeability [97].

![Two-winding core loss measurement method](image-url)
The disadvantage of the two-winding method is that it is sensitive to phase errors at high frequency [98]. The percentage error in the measured loss can be quantified as [97]:

$$\text{percentage error} = \frac{100[\cos(\theta + \varphi) - \cos \theta]}{\cos \theta}$$  \hspace{1cm} (2-1)

where $\theta$ is the actual phase angle between the voltage and current waveforms and $\varphi$ is the phase error which can be introduced by the parasitic impedance in the circuit, digital sampling of the oscilloscope or mismatch between sensing probes. When $\theta$ is close to 90º, a small phase shift error will cause a large error in the core loss. Detailed error analyses in the two-winding core loss measurements have been discussed in [96, 99, 105] and it is shown that this method is not suitable to measure low permeability cores, especially gapped cores.

Another electrical core loss measurement method utilises an impedance analyser and the core loss is derived from the measured impedance characteristics [106-110]. The excitation provided by the impedance analyser is usually a very small sinusoidal signal and it may not be enough to excite power inductors or transformers in a representative manner. In [106, 109], power amplifiers are used to test the component with amplified sinusoidal excitation, as shown in Figure 2-11. In [108], the power amplifier is replaced by a static converter for square wave excitation. The impedance-based method can be an automated procedure that allows the core loss to be swept over a wide frequency range so it is convenient and fast. However, the winding loss is also included in the measurement and this method is sensitive to phase errors.

To overcome the errors due to phase measurement, several techniques have been proposed in the literature. The resonant method is used in [100, 111, 112] where the inductor under test is resonated with a capacitor at the test frequency. This requires the capacitance to be tuned to the appropriate value for each test core and condition. However, the method is limited to sinusoidal waveforms only and the winding loss cannot be separated. A mutual inductance neutralisation technique is proposed in [113] by using an air-cored mutual inductor to phase shift the voltage and current waveforms.
In [92], an additional transformer with a low loss core is used to cancel the reactive power in order to reduce the sensitivity to phase errors.

2.4.1.2 Core loss models

The most commonly used empirical equation to predict the core loss is a simple power law equation, known as the Steinmetz equation (SE) [114]:

\[ P_v = k f^\alpha B_m^\beta \]  

where \( P_v \) is the time-average core loss per unit volume, \( f \) and \( B_m \) are the frequency and the peak induction of the sinusoidal excitation, respectively, and \( k, \alpha, \beta \) are material parameters, often referred to as Steinmetz parameters.

The Steinmetz parameters are determined from linear curve fitting of the measured core loss data under sinusoidal excitation in a double logarithm plot. Therefore they are valid
only for a limited range of frequency and sinusoidal flux density with no DC bias. For
converter applications, rectangular voltage waveforms are commonly used and in
addition the inductor normally carries a large DC current, both of which will affect the
core loss.

To overcome the limitations of the original Steinmetz equation, many researchers have
been working to extend the loss prediction to arbitrary waveforms. The published core
loss models generally fall into four categories: the core loss separation approach,
hysteresis models, the steinmetz-based empirical core loss models and the loss map
approach.

**A. Loss separation approach**

The loss separation approach breaks down the core loss, $P_v$, into three loss components:
hysteresis loss, $P_{hy}$, classical eddy current loss, $P_e$, and excess loss, $P_{ex}$, also called
residual loss or anomalous loss [115]. The hysteresis loss is assumed to be proportional to
the frequency, the classical eddy current loss is assumed to be proportional to the square
of frequency whilst the excess loss is considered to be proportional to the 1.5$^{th}$ power of
frequency [103].

$$P_v = P_{hy} + P_e + P_{ex} = k_h B_m^2 f + k_e B_m^2 f^2 + k_{ex} B_m^{1.5} f^{1.5} \tag{2-3}$$

where $k_h$, $k_e$ and $k_{ex}$ are constants related to the hysteresis loss, classical eddy current loss
and excess loss, respectively, $B_m$ and $f$ are the amplitude and frequency of the sinusoidal
flux waveform. This method has been used in [94, 103, 116]. In [103], equation (2-3) is
extended to rectangular voltage magnetisations. The hysteresis and eddy current loss
calculations are modified to include the effects of a superimposed DC field in [94] but the
excess loss is not included. The loss separation methods are considered to be more
accurate [117] but they require extensive measurements and computation to extract the
coefficients, $k_h$, $k_e$ and $k_{ex}$. 
**B. Hysteresis models**

Hysteresis models, such as the Jiles-Atherton and Preisach models, investigate the core loss from a physical point of view. The Jiles-Atherton model is based on a macroscopic energy calculation and the Preisach model describes the time and space distribution of domain-wall motion [117]. These models have the disadvantage that they tend to be complicated to use in practical designs.

**C. Steinmetz-based core loss models**

The modified Steinmetz equation (MSE) was introduced in 1996 [118] in which the core loss was related to the rate of change of the magnetic induction, \( dB(t)/dt \). An equivalent frequency \( f_{eq} \) is calculated from \( dB(t)/dt \), as in (2-4), to replace the frequency term in the original SE in (2-2).

\[
\begin{align*}
\frac{B_{pk-pk}}{2} & = \frac{2}{\pi^2} \int_0^T \left( \frac{dB(t)}{dt} \right)^2 dt \\
P_v & = \left( k f_{eq} B_m^{\alpha} \right) f_r
\end{align*}
\]

where \( B_{pk-pk} \) is the peak to peak induction, \( T \) is the period of the flux waveform and \( f_r = 1/T \). The MSE uses the Steinmetz parameters from the original SE and therefore does not require additional effort for parameter extraction, but it is sometimes not consistent with the frequency dependence of the original SE. The limitations of MSE are discussed in [119].

The general Steinmetz equation (GSE) was proposed in 2001 [119] to overcome the anomalies in the MSE prediction. The idea is that the core loss depends on not only the rate of the change of magnetic induction, \( dB(t)/dt \), but also the instantaneous induction level, \( B(t) \), as in (2-6).

\[
P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB(t)}{dt} \right|^\alpha \left| B(t) \right|^\beta dt
\]
where
\[ k_i = \frac{k}{(2\pi)^{a-1} \int_0^{2\pi} |\cos \theta|^a |\sin \theta|^{b-a} d\theta} \]  

(2-7)

which is determined from an equivalent loss calculation using the original SE with a sinusoidal excitation with phase angle \( \theta \).

To also take into account the history of the material that affects its reaction to magnetic variation, the GSE was modified and the improved GSE (iGSE) was published in 2002 [101]. The iGSE replaces the instantaneous flux value with the peak to peak induction,  
\[ (2-8) \]

(2-8). The iGSE loss model is used in [117, 120] for core loss evaluations.

The natural Steinmetz extension (NSE),  
\[ (2-10) \]

(2-10), was proposed two years later after the iGSE was introduced [91]. However, the NSE and iGSE end up with the same expression after transformation.

The authors later extended the NSE to a double natural Steinmetz extension (DNSE) in 2005 [121]. The DNSE applies the Steinmetz extensions twice to estimate the loss as two parts, one hysteresis part and the other part dependent on \( dB(t)/dt \).
The improved GSE (i\textsuperscript{2}GSE) which is a further update of the iGSE was published in [102]. The i\textsuperscript{2}GSE considers the effect of magnetic relaxation on the core loss. The magnetic relaxation effect is that the loss is not necessarily zero when \(\frac{dB(t)}{dt}\) becomes zero because there is a short delay in the magnetization. An extra term is added to the iGSE in (2-8) to account for the relaxation effect, as in (2-12).

\[
P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB(t)}{dt} \right|^\alpha B_{pk-pk}^{\beta-\alpha} dt + \sum_{c_{l}=1}^{n} Q_{rl} P_{rl}
\]  

(2-12)

where \(P_{rl}\) is calculated for every voltage change, \(c_{l}\), as

\[
P_{rl} = \frac{1}{T} k_r \left| \frac{dB(t_{-})}{dt} \right|^{\alpha_r} B_{pk-pk}^{\beta_r} \left(1 - e^{-\frac{t_{l}}{\tau}} \right)
\]  

(2-13)

and \(Q_{rl}\) is to determine the change in the voltage

\[
Q_{rl} = e^{-q_r \left| \frac{dB(t_{+})/dt}{dB(t_{-})/dt} \right|}
\]  

(2-14)

The i\textsuperscript{2}GSE improves the accuracy of the core loss estimation compared with the iGSE [102]. Five new material parameters are introduced and they are \(\alpha_r, \beta_r, k_r, \tau\) and \(q_r\). The authors demonstrate the extraction of these parameters, but a large number of measurements are needed, which make this model complicated to use in practice.

The waveform-coefficient Steinmetz equation (WcSE) was proposed in 2008 to accommodate the more complicated resonant operating waveforms in some converters [72]. The flux waveform coefficient (FWC) concept is used to correlate the non-sinusoidal waveforms to the sinusoidal one with the same peak induction. The hypothesis is that the core loss is proportional to the integral of the flux density. The core loss is estimated by multiplying the flux waveform coefficient (FWC) with the original SE, (2-15).

\[
P_v = WFC \cdot k_{f} B_{m}^{\beta}
\]  

(2-15)
The FWC equals to $\pi/4$ for a square voltage waveform (triangular flux) and $\pi/3$ for a triangular voltage waveform (parabola flux). This method is easy to use however it is only limited to 50% duty ratio conditions as the integral of flux density does not alter with duty ratio.

The abovementioned Steinmetz-based core loss models have taken into account the non-sinusoidal waveforms and they are widely used in magnetic component designs. However, none of them includes the DC bias effect.

Most of the work in the literature considers the effect of DC bias on core loss by curve-fitting the experimental measurement results [96, 99, 102, 113], for example the results shown in Figure 2-8. A Steinmetz premagnetisation graph (SPG) is introduced in [96] to show the dependency of Steinmetz parameters on DC premagnetization. In [99], the authors point out that the core loss should be expressed as a function of DC flux density, $B_{DC}$, rather than a function of DC magnetic field, $H_{DC}$, because $B_{DC}$ contains information of core permeability and therefore it is more practical to use in magnetic designs.

Some publications try to estimate the core loss under DC bias conditions by multiplying the Steinmetz equation with a DC factor to account for the effect of DC bias [93, 118, 122, 123]. A displacement factor (DPF) is introduced in [122] which is the ratio between the core losses with and without DC bias. The results are provided for a ferrite material at different operating conditions, showing the DPF is a function of both AC flux density and DC magnetic field.

The measured core loss results reported in the literature show that the core loss increases with increasing DC bias. However, the dependency of the losses on the DC premagnetization for various core materials, for example the experimental curve-fitting data, the SPG or DPF, are not provided by the core manufacturers and are not readily available to use by magnetic designers.
D. Loss map method

The loss map method [104, 124, 125] develops a database that contains the loss information for various operating points, including peak to peak flux density, frequency, DC bias, $H_{DC}$, and temperature. This approach provides loss estimations over a wide range of operating conditions and it is independent of Steinmetz parameters. Details about the loss map structure and how to use the loss map are explained in [104]. However, the use of the loss map method is limited by the availability of the data.

2.4.2 Winding Loss

At low frequency, the winding loss can be calculated from the DC copper resistance and the RMS current [55]. However, the eddy currents in the inductor windings, principally due to skin and proximity effects, become significant sources of power loss at high frequencies and they must be considered in loss calculations.

The high frequency winding losses due to skin and proximity effects have been studied for various types of conductors [71, 126-130]. The AC resistance of the conductors is often calculated from Dowell’s expression [127] where the AC resistance is correlated with the DC resistance by a factor. The analysis is based on the assumption of a one dimensional leakage field passing through the winding.

A general expression for winding loss including high frequency eddy current effects is derived in [131] for inductors for non-isolated pulse-width modulated (PWM) DC-DC converters in continuous conduction mode. The non-sinusoidal inductor current waveform is decomposed into harmonics by Fourier series and the amplitudes of the current harmonics are expressed as a function of the converter’s switching duty cycle. The AC resistances are then calculated at each harmonic frequency using Dowell’s expression. This approach is adopted in this thesis to calculate the losses in the foil windings and the equations will be summarised in Chapter 6.
In the recent years, many researchers have studied the winding losses due to air gap fringing flux in gapped inductors [83, 132-135]. The fringing field around the air gaps will induce eddy currents in the nearby conductors and increase losses. This effect is normally studied by using 2D finite element simulations. To minimise the losses due to fringing flux, it is suggested that the windings must be spaced away from the core by a distance of at least twice the air gap length [83].

2.4.3 Gap Loss

Air gaps are normally used in inductors to prevent core saturation under DC bias conditions, but the air gap fringing flux can cause additional losses in the winding and core [89]. The winding loss due to the air gap fringing field has been thoroughly studied but the phenomenon of gap loss, which is the additional loss in the core, is less understood in the literature.

The gap loss occurs within the core near the air gap in the magnetic circuit and arises due to the fringing flux around the air gap. The fringing of the flux from the sides of the core creates a component of flux that is normal to the surface of the core laminations, shown as $B_n$ in Figure 2-12, and this will therefore create eddy currents and losses within the lamination planes [25]. The effect may be negligible in cores with very high resistivity and isotropic properties such as ferrites, but in some high current, high frequency DC inductors using amorphous metal or nanocrystalline cores where air gaps of 2-3 mm are used, the gap loss can be significant [136].

The core losses in amorphous cut cores with air gaps are investigated experimentally and numerically in [137]. The total loss in the gapped core was measured at 20 kHz and 40 kHz by multiplying the voltage and current in the inductor winding. To reduce the error caused by the phase shift between the current and voltage, a resonant capacitor was used to keep the power factor of the gapped inductor over 0.8. The core loss was deduced by subtracting the winding loss and the capacitor loss from the total loss and the core loss was compared with varying $1/\mu_r$, where $\mu_r$ was the relative permeability. The gap length
is nearly proportional to $1/\mu_r$ and the maximum gap length in the test conditions roughly corresponds to 0.5 mm. The experimental results show that the loss in amorphous cut cores with air gaps increases significantly with increasing air gap length. The loss is increased by approximately 2.5 times when $1/\mu_r$ is increased by 4 times, whilst this effect is not significant in ferrite cores. The increase in core loss is attributed to the in-plane eddy currents due to the fringing flux and a simple finite element model was developed to investigate this phenomenon. The model consists of infinitely laminated amorphous discs of 20 $\mu$m thickness and the fringing field is accounted for by a uniform external field perpendicular to the disc surfaces. However, the fringing flux is highly non-uniform in practice. The authors later examined a nanocrystalline Finemet core in the frequency range 20 kHz to 100 kHz, showing the Finemet core loss also increases strikingly with an increased air gap length [136].

![Schematic diagram of gap loss due to in-plane eddy currents](image)

In a recent publication [138], the increase of core loss due to orthogonal flux is reported for a transformer. The transformer utilises a nanocrystalline core for a 20 kHz dual active bridge converter. Although air gaps are not present, the authors observed high concentration of orthogonal leakage flux in the inner and outer-most core layers,
generating higher losses. They propose to reduce these losses by inserting an adapted leakage layer made of ferrite material outside the core window to reduce the orthogonal leakage flux.

The prediction of gap loss uses an empirically based equation, (2-16), that was originally proposed by Lee in 1947 for steel laminated cores operating at power line frequencies [139]:

\[ P_g = G l_g W_{\text{core}} f B_m^2 \]  \hspace{1cm} (2-16)

where \( P_g \) is the gap loss in Watts, \( l_g \) and \( W_{\text{core}} \) are the total gap length and lamination width in mm, \( B_m \) is the peak induction in the core in Tesla with frequency \( f \) in kHz, and the term \( G \) is a numerical constant.

In the second edition of this book (1955) [140] Lee proposed a change to (2-16) where the frequency term was raised to the power of 0.5 and also the lamination thickness, \( t_l \), and permeability, \( \mu_r \), were included:

\[ P_g = G l_g W_{\text{core}} \mu_r t_l \sqrt{f} B_m^2 \]  \hspace{1cm} (2-17)

However in subsequent publications in the 1970s [25, 141], Lee calculated that the original formulation (2-16) was most accurate. This was based on testing both stamped lamination cores and tape wound cores of silicon steel at frequencies of 60 Hz and 400 Hz. Lamination/Strip thicknesses of 0.36 mm, 0.15 mm and 51 \( \mu \)m were used. Different values were proposed for the constant \( G \) for various configurations of windings and core gaps. For example, for a single cut C-core with two windings, one over each core leg, \( G \) has the value of 0.39.

Equation (2-16) is generally used to predict gap loss in high frequency inductor cores. The equation appears in McLyman’s Transformer and Inductor Design Handbook [89] and it is used in the inductor design software from Metglas [67].
However, some concern has been expressed over the accuracy of equation (2-16) for high frequency inductors using amorphous metal cores. Testing results in [61] show that this equation significantly overestimates the gap loss in an amorphous metal core and the authors suggest the gap loss tends to be insignificant for these types of inductors with very thin laminations and high resistivity. Other work has also suggested that air gap fringing flux has minimal impact on the loss increase caused by cut cores for nanocrystalline materials [72].

2.5 Modelling of Magnetic Components

Design of magnetic components always involves electromagnetic and thermal analyses in order to understand eddy current losses and to predict the temperature rises within the components. Analytical solutions to these problems are usually very complicated and difficult because most of the analyses are two or three-dimensional with complicated shapes and the properties of the magnetic materials are often non-linear or non-isotropic. This section reviews the electromagnetic and thermal modelling techniques for magnetic components.

2.5.1 Electromagnetic Modelling

There are two types of eddy current losses in magnetic cores. One is the classical eddy current loss due to the main magnetic flux along the magnetic path. Core materials such as silicon steel, amorphous metal and nanocrystalline are manufactured as finely laminated cores to reduce classical eddy current losses. The other eddy current loss is the gap loss due to the in-plane eddy currents caused by the perpendicular component of the fringing flux around the core air gaps which has been discussed in Section 2.4.3.

Analytical calculations of the classical eddy current loss in laminated cores are available in [142, 143]. They assume uniform flux distributions in the core and the core geometric effects are ignored. However, a finite element (FE) static analysis is performed in [144] showing the flux distribution in laminated cores is not uniform, especially in gapped
cores the flux concentrates at the corner of the cores. To the author’s best knowledge, no analytical calculation of gap eddy current loss in laminated cores has been published so far.

Numerical methods are usually used for eddy current analysis through FE electromagnetic simulations. The FE analysis not only calculates the total losses in the component but also identifies the distribution of losses. Several publications have described FE electromagnetic models for laminated cores to address the eddy currents due to perpendicular leakage flux but the models proposed in the literature are mostly for low frequency applications (several tens of Hz) [144-150]. The laminated core structures can be modelled by two approaches: the direct approach and the homogenisation approach.

In the direct modelling method each lamination sheet is modelled separately. For example, [137] was mentioned in Section 2.4.3 where the gap loss was studied by modelling individual layers of amorphous discs and insulation between discs. The typical lamination thickness for nanocrystalline or amorphous metal cores is around 20 µm and the thin ribbons are wound to form the core with the core leg being more than 10 mm thick. With such geometry, the number of layers to model is very large. Analysing each layer is therefore time consuming and very expensive in computational cost. The number of elements to solve can be reduced by analysing only a small portion of the core, and modelling only a few laminations. However, this requires that the boundary conditions are carefully defined to set up the magnetic field for this particular part of the core.

The homogenisation approach is a space-averaging method that approximates the laminated core by using an equivalent solid bulk of material with anisotropic properties, as shown in Figure 2-13, where the equivalent anisotropic material properties are calculated from the material properties of the lamination materials [149]. Effectively the equivalent permeability $\mu$ and electrical conductivity $\sigma$ are high in the direction along the laminations, denoted by subscript $t$, and low in the direction normal to the lamination planes, denoted by subscript $n$, to account for the insulation between lamination planes.
The computation of the anisotropic properties have been investigated in [148-151] where the core packing factor is used to determine these equivalent properties. The homogenisation method has been widely used in eddy current analysis and proved to be an effective approximation of the laminated core structure in both linear and nonlinear cases, whilst the computational cost is significantly reduced [147]. This method also reduces the complexity of geometric modelling and it is sometimes used in the modelling of multi-strand or multi-layer windings [152].

![Figure 2-13. Schematic diagrams of modelling of laminated core](image)

The eddy currents due to the fringing flux in an air-gapped silicon steel power reactor have been studied in [153]. A small region near the gap is examined and the analysed region is further divided into two parts. The authors combine the direct and homogenisation techniques by modelling the part near the surface of the core, about 1/10 of the analysed region, with thin lamination layers (0.35 mm per layer), whilst the inner part closer to the middle of the core leg is modelled as a solid bulk with anisotropic properties. Simulations have been performed at the fundamental frequency of 60 Hz and the fifth harmonic of 300 Hz. The results show a significant concentration of eddy current loss in the vicinity of the air gap.

### 2.5.2 Thermal Modelling

It is desirable to predict the temperature within a magnetic component during design, especially to identify the hot spot temperature that limits the size and weight reduction of
the component. The thermal prediction relies on accurate estimations of component losses and the distribution of the losses. Therefore, thermal modelling is often coupled with electromagnetic FEA where the loss distribution can be identified.

Reference [154] carries out a survey of thermal modelling for magnetic components. The commonly used thermal modelling techniques include the lumped parameter method which is based on an equivalent thermal-circuit [155-157] and the thermal finite element analysis [76, 158, 159]. The lumped parameter method is easy to apply but the parameters are difficult to determine [159]. Thermal FEA is accurate and allows a true representation of the component’s geometry. What’s more, if the FE simulation software provides multi-physics solver modules, the electromagnetic and thermal FEAs can be directly coupled without additional modelling effort.

A 3D thermal FE model of a nanocrystalline inductor has been developed in [160] to predict the temperature rise in the core due to gap loss. The gap loss was estimated from the published equation (2-16) while the distribution of gap loss was determined from a static magnetic FE analysis of the same core where the distribution of perpendicular fringing flux was investigated. Two Finemet C-cores have been studied, using a single coil winding around one of the core legs. The details of the static magnetic FEA were not provided in the reference. The simulation results show that the perpendicular component of the fringing flux is highest at the core gap edges and reduces through the laminations and when moving away from the gaps along the core surfaces. However, the resolution of the results seems low, possibly due to a large mesh size being used. The author assumed that the gap loss was proportional to the square of the perpendicular flux and that the total gap loss was distributed in the core ends as illustrated in Figure 2-14. Approximately 70 % of the total gap loss was evenly distributed in the red region and the rest 30 % was allocated in the blue regions. Based on the perpendicular flux distribution results from the magnetic FEA, L1 and L3 were chosen to be 5 mm and 15 mm, respectively, and L2 was assumed to be 25 % of the height of the core leg. The thermal conductivities of the core were 10 W/mK in the direction along laminations and 0.5 W/mK through laminations. The thermal FEA predictions were compared with experimental measurements and the
error between the predicted and measured temperature rises relative to the heatsink was about 20%.

Figure 2-14. Gap loss allocation for thermal FEA [160]

### 2.6 Magnetic Component Design and Optimisation

Many inductor and transformer designs have been published in the literature to reduce the component size and weight by using either low loss materials or advanced cooling arrangements.

In [80], a volume reduction of 50% is reported by using a nanocrystalline core instead of a ferrite core for a 2.2 kW, 100 kHz, high current dual active bridge converter. Various magnetic materials are compared as uncoupled or coupled inductors in interleaved boost converters in [161] and the results suggest that the nanocrystalline material Finemet can contribute to the reduction in inductor loss and volume.

A low-profile foil-wound inductor is compared with a conventional litz-wound EFD inductor in [84]. The foil-wound prototype achieves a half reduction in component size comparing with that of the litz-wound component for the same temperature rise when convection cooled by natural air. This is due to the high copper packing factor and superior thermal performance of the foil winding. The prototype is further improved
when encapsulated in an aluminium case for conduction cooling and the volume reduction becomes four times that of the litz-wound structure.

The aluminium casing technique has been employed in [18, 23, 155-157, 159, 162, 163] for compact inductor designs. The inductors are potted into aluminium cases with thermally conductive potting materials to enhance the heat transfer between the component and the case. The encapsulated components are then bolted onto the converter’s water-cooled cold-plate. Reference [163] compares three types of potting material, silicones, epoxies and urethanes, for a potted inductor design, and tests them on a 10 kHz, 40 kW, DC-DC converter, showing that by using high thermo-stable silicone potting material the potted inductor reduces the weight by 75 % compared with conventional design, and the steady state temperature in the component is reduced by 30 °C.

A number of novel designs have also been proposed in the literature by optimising the core or winding structures to reduce losses.

A constant-flux inductor is proposed in [164] for a 500 kHz, 5 kW, boost converter application. The constant-flux inductor is made of several concentric toroidal core cells in order to distribute the magnetic flux and energy uniformly throughout the component. The constant-flux inductor design reduces the volume by 50 % compared with a commercial product with the same inductance.

Several authors suggest using smaller distributed gaps (discrete) instead of a single lumped air gap to reduce the winding loss due to the air gap fringing flux [82, 83, 159]. The effect of the distributed-gap on a foil winding has been investigated experimentally in [57, 61]. The core and winding temperatures were measured in amorphous metal inductors using single-cut and multi-cut cores where the multi-cut core splits the single gap per core leg, 2.625 mm, into three smaller gaps. A temperature drop of 18 °C was observed in the foil winding by using the multi-cut core due to reduced winding loss compared with the single-cut configuration. The core temperature, however, was higher.
in the multi-cut core because the cutting process causes shorts between laminations. The multi-cut structure therefore requires additional effort in both manufacture and assembly procedures.

Another approach reported in the literature to reduce the effect of the air gap on winding loss is to place flux barriers between the core gap and winding, for example an open-circuit copper screen is proposed in [165] and a low permeability magnetic material is used in [166] to create an auxiliary magnetic path to reduce winding loss.

Some researchers have proposed to shape the foil windings to reduce high frequency copper losses. A combination of thick and thin foil windings is described in [167]. The thin foils with lower AC resistance are used as the inner turns closer to the gaps to reduce high frequency losses whilst the thick foils are used away from gaps to reduce DC resistance. In [168], the conductor is cut away from the area that experiences the strongest fringing field.

To reduce the core loss due to air gaps, thin ferrite plates are attached to the cut surfaces of the core, as shown in Figure 2-15 [136, 169]. The principle is that the fringing flux tends to leave the core from the high-resistivity ferrite plates to suppress the eddy currents. However, ferrites have low saturation flux which limits the use of this method under large DC bias conditions.

Figure 2-15. Reducing the in-plane eddy current loss by attachment of thin ferrite plates [136]
In [170], the authors propose to reduce the eddy current loss by breaking the eddy current path with slits in the silicon steel sheets and the simulation results show that the slit reduces the maximum loss density at the core surface by 60%.

As described in Section 2.3.2, planar magnetic technology is one of the promising techniques to reduce the size and cost and improve the thermal performance of magnetic components. A recent publication [87] provides an overview of the state of the art of planar magnetic technologies, and many researchers have proposed planar inductor designs for high frequency power converters [88, 159, 171-173]. Although planar magnetic components can offer several advantages, the established designs tend to be limited to low power applications.

2.7 Summary of Literature Review

Compact DC-DC converters are used in EV power trains to interface the energy sources and storage devices with the traction drives. Many converter topologies have been published in the literature and the dual-interleaved, bi-directional non-isolated DC-DC boost converter with interphase transformer has been identified as a candidate topology for this application. However, for high step-up ratios isolated topologies are used.

Achieving a high power density design of the DC-DC converters requires a reduction in the size and weight of the magnetic components, inductors and transformers, which can account for around 30-50% of the converter weight. One way to achieve high power density magnetic designs is to use low loss core materials and windings. The magnetic materials currently being used in DC-DC converters have been reviewed and the winding techniques have been discussed. Nanocrystalline materials and copper foil windings have several advantages for high-power, high-frequency inductor applications. Nanocrystalline cores offer a high saturation flux density of over 1 T with low loss characteristics as well as a high thermal stability. Foil windings have high filling factor and high current capability. Also the large surface area of copper foils allows the heat to be easily removed.
The loss mechanisms in inductor core and windings have been reviewed. Recent research has focused on the calculation of the traditional core loss and high frequency winding loss whilst the gap loss is not particularly well understood or quantified at the moment, especially at high frequency.

The only empirical equation to estimate the gap loss dates back to the 1940s and was intended for line frequency applications. The accuracy of the established equation at high frequency is questionable. Electromagnetic FE simulations are commonly used for eddy current analysis and can be used to study the gap loss effect. The challenge lies in the modelling of finely laminated cores and this has been overcome in the literature by using the homogenisation technique where the laminations are represented by a solid bulk with anisotropic properties. Once the losses have been estimated and the distribution of the losses has been determined, thermal modelling can be performed to predict the temperature rise in the component and identify the hot spot which allows the design to be optimised and an effective cooling arrangement to be implemented.

Several magnetic designs in the literature have been described. They try to reduce the component volume by using low loss materials, novel core or winding structures or advanced thermal management, for example encapsulating the component has been established as an effective technique to increase the power density of magnetic components.

This thesis first characterises the core losses in nanocrystalline cores to provide a more complete picture of the loss variations with typical operating conditions in high power converter applications. Then, the gap loss, which is less well understood in the literature, is examined by 3D finite element analysis. The FE modelling developed in this thesis is used to calculate the gap loss and its distribution in a high power nanocrystalline inductor under high frequency operation, and therefore the temperature distribution within the component is predicted and the hot spot is identified, which potentially allows a further size and weight reduction to be achieved.
CHAPTER 3

CORE LOSS CHARACTERISATION OF NANOCRYSTALLINE CORES

3.1 Introduction

Whilst a number of empirical models have been proposed for core losses as discussed in Chapter 2, the model parameters are not normally provided by manufacturers. Furthermore, the measured loss data that is available from data sheets tends to be for a limited range of conditions and usually sinusoidal waveforms. Therefore, to understand better the variation of losses under realistic converter operating conditions and to quantify the differences in behaviour between similar materials, a detailed study is undertaken.

In this chapter, the losses of four nanocrystalline cores are experimentally characterised under typical converter operating conditions (squarewave excitation with various duty ratios and DC bias conditions). The measurement results show that the core loss increases with increasing DC bias especially at high DC flux density levels, and the duty ratio also affects the core loss significantly. The measured data can be used as a basis for the design and optimisation of a high power inductor.

3.2 Experimental Set Up

The classic two-winding technique, as discussed in Section 2.4.1.1, has been used to examine the core losses in this chapter due to its fast and simple implementation. Figure 3-1 shows the schematic diagram of the test set up. The current in the primary winding, \( i_p(t) \), and the voltage across the secondary winding, \( v_s(t) \), of the cores under test (CUTs) are measured, and then the core loss is assessed from the B-H loops of the test components. The measurement equipment used is listed in Appendix A.
Four nanocrystalline cores of similar dimensions were tested and characterised. The core parameters are summarised in Table 3-1. These cores are wound with very thin nanocrystalline ribbons, around 18 µm. All the sample cores are cut cores. To minimize the phase error in the measurements and to ensure the gap loss is virtually zero, the cores are wound as gapless inductors. The cut surfaces are carefully aligned and the core halves were fastened with a clamp during the tests.

Table 3-1. Nanocrystalline test cores

<table>
<thead>
<tr>
<th>Material</th>
<th>Finemet FT-3M</th>
<th>Vitroperm 500F</th>
<th>Vitroperm 500F</th>
<th>Vitroperm 500F</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core part number</td>
<td>F3CC0050</td>
<td>W156-01</td>
<td>W156-03</td>
<td>3397</td>
</tr>
<tr>
<td>$A_c$ (mm$^2$)</td>
<td>312</td>
<td>260</td>
<td>260</td>
<td>345</td>
</tr>
<tr>
<td>$l_m$ (mm)</td>
<td>244</td>
<td>190</td>
<td>192</td>
<td>182</td>
</tr>
<tr>
<td>Lamination thickness (µm)</td>
<td>18</td>
<td>18</td>
<td>18</td>
<td>18</td>
</tr>
<tr>
<td>Packing factor</td>
<td>0.78</td>
<td>0.78</td>
<td>0.78</td>
<td>0.83</td>
</tr>
<tr>
<td>Weight (kg)</td>
<td>0.573</td>
<td>0.417</td>
<td>0.407</td>
<td>0.525</td>
</tr>
</tbody>
</table>
Each of the cores under test was wound with a copper foil winding that was split between the two core legs, Figure 3-2. A current probe was used to measure the excitation current in the foil winding. On the secondary side, the voltages were measured from four sense coils made of twisted wires that were placed at different locations around the core magnetic path but still very close to the primary winding so that they were closely coupled. The measurements from the four sense coils were averaged for improved accuracy. The current and voltage probes were calibrated and connected to a digital oscilloscope where waveforms were stored and exported to a PC for further data processing using Matlab.

![Figure 3-2. Test core with primary and secondary windings](image)

In order to measure the core loss under typical operating conditions that arise in DC-DC converters, a 25 kHz, 50 kW, dual-interleaved DC-DC converter with IPT (shown in Figure 2-3 in Chapter 2) was used to excite the CUTs where they were wound and tested individually as the input inductors in the converter. The converter was water-cooled and the measurements were taken rapidly to minimise the influence of temperature changes on the losses. The core temperature was monitored by a thermal camera and kept below 50 ºC for all the measurements.
The converter subjects the inductor to a rectangular voltage waveform with variable duty ratio, combined with a range of DC magnetisations that depend on the converter input current, $I_{DC}$, which can be adjusted by varying the converter’s voltage and load.

The typical inductor voltage and current waveforms are shown in Figure 3-3, where the effective duty ratio, $D$, and period, $T$, of the inductor waveforms can be determined from the converter’s switching duty ratio, $D_{sw}$, and frequency, $f_{sw}$, as in (3-1) and (3-2). The effective operating frequency of the component is 50 kHz, twice the switching frequency due to interleaving.

![Figure 3-3. Inductor voltage and current waveforms](image)

The flux density, $B(t)$, in the core is calculated by integrating the measured secondary voltage $v_s(t)$,

$$B(t) = \frac{1}{N_sA_c} \int_0^t v_s(\tau)d\tau$$  \hspace{1cm} (3-3)

where $N_s$ is the number of turns in the sense winding and $A_c$ is the core cross sectional area.
The current measured from the primary winding, \( i_p(t) \), is used to determine the magnetic field strength, \( H(t) \),

\[
H(t) = \frac{N_p i_p(t)}{l_m}
\]  (3-4)

where \( N_p \) is the number of turns in the primary or excitation winding and \( l_m \) is the mean length of the core magnetic path.

The core loss can be calculated by integrating the product of the measured voltage and current waveforms, and by substituting for \( v_s(t) \) and \( i_p(t) \) using (3-3) and (3-4) the core loss per unit volume is equal to the frequency multiplied by the enclosed area of the B-H loop, as illustrated in (3-5).

\[
P_v = \frac{f \int_0^T N_p v_s(t) i_p(t) dt}{A_c l_m} = \frac{f \int_0^T N_p N_s A_c \frac{dB(t)}{dt} \cdot \frac{H(t) l_m}{N_p} dt}{A_c l_m}
= f \int_{B(0)}^{B(T)} H(B) dB = f \oint H dB
\]  (3-5)

**3.3 Measurement Results**

The core losses in the sample nanocrystalline cores have been measured under varying AC excitations at two effective duty ratios, 50 % and 20 %, with DC bias conditions ranging from 0.23 T to 1 T. Measurement results are provided in this section, showing the variation of core losses with different operating conditions.
3.3.1 Measured Magnetisation Curves

Before determining the core losses, the permeabilities of the cores are first obtained from the measured magnetisation curve. The permeability data is required to calculate the DC flux density, $B_{DC}$, from the knowledge of the magnetic field strength, $H_{DC}$, Figure 3-4, which in turn may be determined from the DC inductor current using (3-6). The purpose was to determine the DC current test conditions to measure the core losses in the sample cores under comparable DC flux density conditions.

$$H_{DC} = \frac{N_p I_{DC}}{I_m}.$$  \hspace{1cm} (3-6)

The magnetisation curves were measured under square-wave voltage excitation with a DC bias, and the operating conditions were chosen so that the cores were saturated and the inductor currents varied from zero to maximum. The magnetisation curves were then obtained from the average of the increasing and decreasing B-H curves from the measurements.

Figure 3-5 shows the measured magnetisation curves of the cores. The four sample cores have similar saturation flux densities, around 1.2 T, which agree with data provided by
the manufacturers. However, the permeabilities of the sample cores are quite different. The Finemet core has the lowest relative permeability in the linear region, about 2500, while the Vitroperm 500F cores saturate at lower $H_{DC}$ levels. At $H = 200$ A/m, the Finemet core is still operating linearly while the Vitroperm cores have already started to saturate. It is noticed that the two sample cores from Vacuumschmelze (VAC) have very different permeabilities even though they are made with the same material by the same manufacturer. Queries have been sent to the manufacturer for further information and the replies explained that the sample core W156-01 was an old generation product.

![Figure 3-5. Measured full magnetisation curves at room temperature](image)

To further verify the measured permeability of the cores under test, the cores were wound as un-gapped inductors with $N$ turns and the inductances, $L$, were measured using a precision impedance analyzer (Appendix A). With the measured inductance values, the relative permeability can be estimated from equation (3-7) using the manufacturers’ $A_c$ and $l_m$ data as listed in Table 3-1. The results are consistent with the measurements, as shown in Table 3-2, with errors less than 10%.

$$L = \frac{N^2 A_c}{l_m \mu_r \mu_0}.$$  \hspace{1cm} (3-7)
### Table 3-2. Core permeabilities and inductances at 50 kHz

<table>
<thead>
<tr>
<th>Sample core</th>
<th>Measured $L$ ($N = 1$)</th>
<th>Estimated $\mu_r$, from measured $L$</th>
<th>Measured $\mu_r$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Finemet (F3CC0050)</td>
<td>3.9 $\mu$H</td>
<td>2402</td>
<td>2483</td>
</tr>
<tr>
<td>Vitroperm 500F (VAC W156-01)</td>
<td>13.2 $\mu$H</td>
<td>7676</td>
<td>7828</td>
</tr>
<tr>
<td>Vitroperm 500F (VAC W156-03)</td>
<td>32 $\mu$H</td>
<td>18804</td>
<td>19310</td>
</tr>
<tr>
<td>Vitroperm 500F (MK 3397)</td>
<td>16.5 $\mu$H</td>
<td>6926</td>
<td>6465</td>
</tr>
</tbody>
</table>

#### 3.3.2 Effect of Duty Cycle on Core Loss

To investigate the effect of duty cycle on core loss, the losses were measured at 50 % and 20 % duty ratio. Figure 3-6 and Figure 3-7 show the example measured waveforms and B-H loops at the two duty ratios under typical DC bias conditions for Finemet cores. The gate voltage $v_{sw}(t)$ in the transistor of one of the converter’s interleaved phases is also presented, showing the switching duty ratios of 75 % and 60 %. The other converter phase is switched with a half cycle delay. The cores under test are therefore operated at twice the switching frequency at 50 % and 20 % duty ratios, respectively, as shown in $v(t)$ and $i_p(t)$.

Figure 3-8 to Figure 3-11 provide the example B-H loops under high DC bias conditions (over 0.8 T) for the four sample cores, respectively. The different permeabilities of the cores can be seen. In addition, the B-H loops are compared at the same DC and AC magnetisations but different duty cycles. The increased area of the B-H loops at $D = 0.2$ is evident, showing that the core loss is higher at 20 % duty cycle as the core loss density per cycle is calculated by integrating the areas of the B-H loops. At $D = 0.2$, the waveform is asymmetric with a steeper rising slope due to the more rapid increase in flux and a slower falling edge for the decrease in flux. According to the modified Steinmetz equation that has been discussed in Section 2.4.1.2, the asymmetric waveforms effectively increase the equivalent frequency, which is calculated from $dB(t)/dt$, and therefore increase the core loss. According to the literature, the core loss variation with
duty ratio is symmetrical about 0.5, implying that the core losses are the same at \( D = 0.2 \) and \( D = 0.8 \) because the calculated equivalent frequencies have the same value.

Figure 3-6. Measured waveforms at \( D_{sw} = 0.75 \), \( D = 0.5 \) (core: Finemet F3CC0050, \( N_p:N_s = 2:1 \), \( B_{DC} = 0.64 \) T)
a. Measured voltage and current waveforms

b. B-H loop

Figure 3-7. Measured waveforms at $D_{sw} = 0.6$, $D = 0.2$ (core: Finemet F3CC0050, $N_p:N_s = 2:1$, $B_{DC} = 0.46$ T)
CHAPTER 3    CORE LOSS CHARACTERISATION OF NANOCRYSTALLINE CORES

Figure 3-8. Measured B-H loops (Finemet F3CC0050, $B_{DC} = 0.95$ T)

Figure 3-9. Measured B-H loops (Vitroperm 500F W156-01, $B_{DC} = 0.81$ T)

Figure 3-10. Measured B-H loops (Vitroperm 500F W156-03, $B_{DC} = 0.95$ T)
3.3.3 Effect of DC Bias on Core Loss

The core loss measurement results under various DC bias conditions are provided in this section. The DC operating current conditions for the tests were determined from the measured magnetisation curves, as shown in Section 3.3.1, to examine the core losses under comparable DC flux density levels. For each DC bias condition, the losses were measured at several AC induction levels to obtain a complete picture of the core loss variations with both DC and AC excitations.

Figure 3-12 to Figure 3-15 compare the measured core losses of the four sample nanocrystalline cores with the manufacturers’ loss data. For the Finemet core, the core loss information was received from the manufacturer via email based on their laboratory tests on F3CC series cut cores. The loss is slightly higher than the datasheet values for the raw material which is likely to be due to the cutting process.
Figure 3-12. Finemet F3CC0050 core losses under different operating conditions.
a. $B_{DC} = 0.23$ T  

b. $B_{DC} = 0.46$ T  

c. $B_{DC} = 0.64$ T  

d. $B_{DC} = 0.81$ T  

e. $B_{DC} = 0.95$ T  

f. $B_{DC} = 1$ T

Figure 3-13. Vitroperm 500F W156-01 core losses under different operating conditions
a. \( B_{DC} = 0.23 \, \text{T} \)

b. \( B_{DC} = 0.46 \, \text{T} \)

c. \( B_{DC} = 0.64 \, \text{T} \)

d. \( B_{DC} = 0.81 \, \text{T} \)

e. \( B_{DC} = 0.95 \, \text{T} \)

f. \( B_{DC} = 1 \, \text{T} \)

Figure 3-14. Vitroperm 500F W156-03 core losses under different operating conditions
Figure 3-15. Vitroperm 500F 3397 core losses under different operating conditions

a. $B_{DC} = 0.23$ T
b. $B_{DC} = 0.46$ T
c. $B_{DC} = 0.64$ T
d. $B_{DC} = 0.81$ T
e. $B_{DC} = 0.95$ T
f. $B_{DC} = 1$ T

CHAPTER 3  CORE LOSS CHARACTERISATION OF NANOCRYSTALLINE CORES
The results show that the core losses increase with increasing AC induction, and at $D = 0.2$ the core losses are higher than those at $D = 0.5$, which have been indicated by the increased area of the B-H loops presented in Section 3.3.2. Furthermore, the increasing DC bias increases the core losses further. For example, at $D = 0.5$ with a low DC bias the losses in the Finemet core are well below the manufacturer’s data which is provided for sinusoidal excitations. This is consistent with the published results in [92] (Figure 2-7), which shows that the core loss under symmetric triangular flux is lower than the loss under sinusoidal flux waveform. However for high DC bias and an asymmetric waveform ($D = 0.2$), the losses exceed the values provided by the manufacturers. Using the manufacturer’s loss data without considering the waveform and DC bias may result in poorly optimised designs which either overheat or are larger in size than necessary.

To further visualise the effect of DC bias on the core losses, the measurement results are curve fitted in Figure 3-16 to show the variations of losses with DC bias at the same peak flux density. The results show that at low DC bias levels, the impact of DC bias on core loss is small. However, a substantial increase in core losses is observed at high DC flux conditions (over 0.8 T).

Among the four cores under test, the Finemet core has the highest core loss density and the loss increases steeply with increasing DC bias, shown in Figure 3-16a. For example, when the peak flux density $B_m$ is 100 mT, the loss density is increased by five times just by increasing the DC flux. Also, the core loss is strongly affected by the duty ratio of the excitation waveform. The losses of Finemet vary by almost an order of magnitude as the DC magnetisation increases and the duty ratio moves away from 0.5 to 0.2.

The Vitroperm 500F cores have lower losses than Finemet. At low DC magnetisations, the three sample Vitroperm cores have comparable losses. For higher DC bias conditions, the W156-01 core loss increases considerably, Figure 3-16b, whilst for the other two Vitroperm cores, W156-03 and 3397, the increase in core losses is relatively small, as shown in Figure 3-16c and d.
a. Finemet, F3CC0050  
b. Vitroperm 500F, W156-01  
c. Vitroperm 500F, W156-03  
d. Vitroperm 500F, 3397

Figure 3-16. Core loss variation with DC bias conditions

Table 3-3 compares the sample cores. The Vitroperm core 3397 from the manufacturer MK Magnetics has a higher continuous operating temperature (180 °C) than the others, which is likely to be an advantage in minimising the size and weight of a wound component. However, Vitroperm components have higher permeability and they tend to saturate at lower DC current conditions.
Table 3-3. Test core comparisons

<table>
<thead>
<tr>
<th>Cores</th>
<th>$\mu_r$</th>
<th>Core loss @ $B_m = 0.1$ T, $B_{DC} = 0.23$ T, 50 kHz (W/kg)</th>
<th>Core loss @ $B_m = 0.1$ T, $B_{DC} = 1$ T, 50 kHz (W/kg)</th>
<th>Continuous operating temperature (ºC)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Finemet F3CC0050</td>
<td>2500</td>
<td>2.8</td>
<td>11.5</td>
<td>155</td>
</tr>
<tr>
<td>Vitroperm 500F W156-01</td>
<td>6800</td>
<td>2.2</td>
<td>6.1</td>
<td>120</td>
</tr>
<tr>
<td>Vitroperm 500F W156-03</td>
<td>20000</td>
<td>2.7</td>
<td>3.4</td>
<td>120</td>
</tr>
<tr>
<td>Vitroperm 500F 3397</td>
<td>8000</td>
<td>2.0</td>
<td>2.9</td>
<td>180</td>
</tr>
</tbody>
</table>

### 3.4 Conclusion

The core losses have been measured in four nanocrystalline sample cores under representative operating conditions that are encountered in typical high power converter applications. The results are presented in this chapter, comparing the core losses under various DC and AC flux levels.

It is shown that the core losses are strongly affected by the duty ratio of the excitation waveforms. Higher losses are observed at asymmetric waveform conditions ($D = 0.2$). Furthermore, the core losses are characterised as a function of DC flux density, showing an increase in loss with increasing DC magnetisation, especially at high DC flux density levels. As the operating condition varies, the core losses can be either well below or higher than the manufacturers’ loss data. The results indicate the weakness of the limited data provided by manufacturers and highlight the importance of detailed characterisation if optimal magnetic component designs are to be produced.

The measurement results presented in this chapter provide a more complete picture of the loss characteristics and the loss data may be used to form a basis for the design and optimisation of an inductor for a high power DC-DC converter.
CHAPTER 4

4 FINITE ELEMENT MODELLING OF LAMINATED CORES FOR GAP LOSS CALCULATION

4.1 Introduction

To address the limited understanding of gap losses in finely laminated inductor cores and the absence of an accurate method for loss calculation, this chapter describes the development of a 3D geometric model and the 3D finite element meshing of the model. The model developed will be used to compute the distributions and the magnitude of the gap losses in the wound component.

4.2 Modelling Method for Laminated Cores

To calculate accurately the additional core losses associated with the in-plane eddy currents around the core air gap, the homogenisation technique is adopted in this work to model the laminated core. This is because detailed modelling of the individual lamination layers, which are approximately 20 μm thick, would present an enormous computational challenge across the core dimensions. The schematic drawing of the homogenisation modelling technique has been presented in the literature review in Section 2.5.1, the diagram is repeated here in Figure 4-1 for convenience.

The laminated core structure is modelled by a solid continuum with equivalent permeability, \( \mu_n \), and electrical conductivity, \( \sigma_n \), in the direction along the laminations, and lower permeability, \( \mu_n \), and electrical conductivity, \( \sigma_n \), in the direction through the laminations.
a. Direct method  b. Homogenisation method

Figure 4-1. Schematic diagrams of modelling of laminated core

The traditional homogenisation technique normally sets $\sigma_n$ to zero to eliminate the classical eddy currents. But this is not sufficient to reflect the magnetic field distribution in the core at high frequency. Therefore, the improved homogenization technique proposed in [150] is used in this work, where the equivalent permeability and conductivity are calculated from the core material properties in the lamination rolling direction, $\mu_m$ and $\sigma_m$, as follows.

$$
\mu_t = F\mu_m + (1 - F)\mu_0 \\
\mu_n = \frac{\mu_m\mu_0}{F\mu_0 + (1 - F)\mu_m} \\
\sigma_t = F\sigma_m \\
\sigma_n = \left(\frac{t_l}{W_{core}}\right)^2 \frac{1}{F}\sigma_m
$$

where $F$ is the core packing factor, typically around 80 % for nanocrystalline wound cores, $\mu_0$ is the vacuum permeability, and $t_l$ and $W_{core}$ are the thickness and width of the lamination strips, respectively.
With $\mu_m = 2500$ and $\sigma_m = 8.33 \times 10^5$ S/m for the Finemet core, the calculated equivalent permeabilities and conductivities are listed in Table 4-1.

Table 4-1. Equivalent material properties in homogenised Finemet Core

<table>
<thead>
<tr>
<th>Equivalent properties</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\mu_s$</td>
<td>2000</td>
</tr>
<tr>
<td>$\mu_n$</td>
<td>5</td>
</tr>
<tr>
<td>$\sigma_t$ (S/m)</td>
<td>$6.67 \times 10^5$</td>
</tr>
<tr>
<td>$\sigma_n$ (S/m)</td>
<td>0.46</td>
</tr>
</tbody>
</table>

The skin depth is an important parameter to consider for the core models because it is often related to the required FE mesh size in eddy current analysis. In the direct laminated model, as shown in Figure 4-1, the skin depth within a lamination is calculated as

$$\delta = \sqrt{\frac{2}{\omega \mu_m \mu_0 \sigma_m}}$$  \hspace{1cm} (4-5)

where $\omega$ is the angular operating frequency. The skin depth can be very small for high permeability cores at high frequency. For the Finemet lamination material, the calculated skin depth is less than 60 µm at 60 kHz. To create a finite-element mesh with element size smaller than this is impractical.

However, considering the homogeneous model an effective skin depth can be determined based on the equivalent permeability and conductivity of the core [150]. The normal flux component, $B_n$, induces in-plane eddy currents that flow along the tangential direction along the laminations. Therefore, the effective skin depth for the homogeneous core model is calculated from (4-6) using $\mu_n$ and $\sigma_t$ listed in Table 4-1 and it is around 1.2 mm at 60 kHz, which allows an achievable mesh size to be used. The implications of skin depth on the element size in the FE model are discussed in Section 4.5.1.2.
4.3 FEA Software

The software used for the modelling of gap loss is Vector Fields ‘Opera’ from Cobham [175]. Opera 3D is a powerful FEA based simulation package for three-dimensional electromagnetic analysis. The embedded module, ELEKTRA/SS, analyses steady-state electromagnetic fields including the effects of eddy currents, and therefore it is used in this work to solve the in-plane eddy currents caused by the fringing field. Opera 3D also has a thermal module, TEMPO, which will be used in the Chapter 6 to predict the temperature rise due to the gap loss.

The flow chart in Figure 4-2 describes the main steps required to build and solve a FE model in Opera-3D. The geometric model is built first. The material characteristics are assigned and volume properties for each part of the model are defined. The model symmetry can be exploited to reduce the size of the model and save computational cost, however the boundary conditions need to be determined correctly. Then the model is ready to mesh and the shape and size of the finite elements must be carefully chosen to achieve the desired accuracy. After meshing the model, a simulation database is generated and then solved at the intended operating frequencies. The results can be displayed and analysed in the post-processor.

In the following sections, the modelling and meshing of the 3D inductor core model will be described. The post-processing of the simulation results is presented in Chapter 5.
4.4 Initial Inductor Modelling

This section describes the initial modelling of a sample inductor, including the geometric modelling and the investigations of model symmetry and boundary conditions.

As mentioned in the literature review in Section 2.5.1, the established electromagnetic models in the literature normally set up only a small section of the laminated core and the flux is defined by either boundary conditions or uniform external magnetic fields.

In this work, the complete core structure is to be modelled and the excitation is provided by current-driven Biot-Savart conductors which represent the inductor windings. The aim
is to represent the true flux density distributions in the component which could be influenced by the core geometry, air gap length as well as the windings. Furthermore, modelling the complete core allows the 3D thermal FEA to be easily correlated with the electromagnetic FEA results.

4.4.1 Core Modelling

Laminated cores may be classified according to the ways they are constructed into stacked cores and wound cores, as shown in Figure 4-3. Both core types could suffer from gap loss due to the air gap fringing flux.

The stacked cores are common in larger power transformer applications. The laminations of stacked cores have a uniform or constant direction throughout the component and therefore they are easier to model, Figure 4-3a. However, in small distribution transformers or inductors, wound cores predominate [176]. Nanocrystalline cores for high frequency converter applications are normally manufactured as tape wound cores. One challenge in the modelling of the wound cores is that the laminations change their directions along the magnetic path, as can be seen in Figure 4-3b.
A Finemet C-core, F3CC0032, from Hitachi Metals [70] has been used in the development of the modelling techniques. The core is wound with 18 µm nanocrystalline amorphous ribbon and a thin layer of epoxy provides electrical insulation between the laminations. The core is cut in the middle of the longer core legs to form the air gaps. Figure 4-4 and Table 4-2 provide the physical dimensions of the core from datasheet [70]. The total air gap length, $l_g$, is set to be 4.4 mm with 2.2 mm in each core leg.

![Figure 4-4. Finemet F3CC0032 core dimensions](image)

<table>
<thead>
<tr>
<th>Core Dimensions</th>
<th>Values (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core strip width, $W_{core}$</td>
<td>30</td>
</tr>
<tr>
<td>Core leg height, $H_{core}$</td>
<td>13</td>
</tr>
<tr>
<td>Core leg length, $L_{core}$</td>
<td>28</td>
</tr>
<tr>
<td>Core window height, $H_{window}$</td>
<td>15</td>
</tr>
<tr>
<td>Total air gap length, $l_g$</td>
<td>4.4</td>
</tr>
<tr>
<td>Mean magnetic path length, $l_m$</td>
<td>194</td>
</tr>
</tbody>
</table>
4.4.1.1 Geometric modelling of the core

For the purpose of FE modelling, the wound core structure is formed by several individual homogeneous blocks with ideal boundaries between each. Figure 4-5 shows the core model where the left core half is expanded to show the individual blocks and the right core half shows the assembled view of the core.

The volume properties are defined for each of the core sections to set the local coordinates in these blocks. The core legs with gaps and the short core legs at the end of the U-shape are modelled with rectangular boxes and Cartesian coordinates, whilst each of the core corners is modelled by a quarter of a cylinder and the cylindrical coordinate system is used to define the volume orientations. As shown in Figure 4-5, the vectors indicate the directions normal to the lamination planes in each core segment, and clearly show how the laminations bend at the corners to form the wound core. The axes in the bottom right corner of Figure 4-5 show the global co-ordinates that do not rotate with the local coordinate systems in the individual core blocks.

![Figure 4-5. Wound C-core model with volume orientation vectors](image)

This chapter presents the modelling of a C-core, however the geometric modelling using the volume orientation technique is broadly applicable to different laminated core shapes, for example a toroidal core or an E core.
4.4.1.2 Material properties

Each of the core blocks is modelled by the equivalent anisotropic characteristics discussed in Section 4.2. To reduce the complexity of the problem, a linear permeability is assumed.

It is assumed that the gap loss is mainly distributed in the regions near the air gaps where the fringing fluxes are large. Therefore, the longer core legs with gaps (light green parts in Figure 4-5) are the major regions of concern and they are assigned with the equivalent permeability and electrical conductivity calculated from typical Finemet core properties, as listed in Table 4-1. As for the dark green regions where the gap loss is assumed to be very small, the same permeability is used, but the electrical conductivity is set to be very low. Using the same permeability along the magnetic path provides a complete magnetic path for the core and allows the flux distribution to be determined, but meanwhile, setting the conductivity to be very low in the dark green regions (Figure 4-5) avoids the calculation of eddy currents, which are due to the crowded flux density near the core corners where the laminations effectively rotate. Moreover, this simplification allows a large mesh size to be used in the dark green regions to reduce the computational effort, because the low conductivity will lead to a large effective skin depth. Ideally in these regions the conductivity could be set to zero but in practice a small non-zero value is used to prevent convergence problems in the FE solution.

4.4.2 Coil Modelling

The modelling of gap loss in this work only investigates the eddy currents induced in the core while the losses in the winding due to the fringe fields are not considered. The coils in the model are therefore merely used as the sources of magnetic fields. These coils are modelled by Biot-Savart conductors in the Opera software.

The Biot-Savart conductors are not part of the finite element mesh and the material properties of the conductors, for example resistivity, do not need to be defined. The
magnetic fields are calculated from the pre-defined current densities in the conductors by the Biot-Savart law [177]. In addition, coils that consist of multiple turns can be modelled with only one turn but with an appropriate current density to represent the amp turns of the actual winding. Using the Biot-Savart conductors can reflect the actual magnetic field generated by the inductor winding, and the conductor dimensions, shapes and positions can be easily altered for different winding designs.

In this model, two racetrack coils, as shown in Figure 4-6, are used to represent the copper foil windings that are wrapped around the gapped core legs. The width of the conductors, $W_{\text{coil}}$, is 51 mm, slightly shorter than core window breath. The total number of turns is split equally between the core legs. The currents in the two coils are oriented such that the fields are additive in the core. The current density in each coil is defined as $NI_m/2A_{\text{coil}}$, where $N$ is the total number of turns in the actual winding, $I_m$ is the amplitude of the sinusoidal excitation current and $A_{\text{coil}}$ is the cross sectional area of the coil conductor in the model.

Figure 4-6. Inductor model with winding

Having a space between the coil and core, especially around the gap, is common practice to minimise the fringe-field-related losses in the winding. Therefore, the coils are placed
3.2 mm away from the core in the model to be consistent with the practical designs and to account for the flux leakage. However, as the Biot-Savart conductors are only used as magnetic sources in the model, they are not flux barriers and will not affect the fringing flux in the simulation. The modelling results presented in Chapter 5 show that at this distance the winding has little influence on the fringe field.

4.4.3 Background Region

The regions surrounding the core and winding are simply modelled as free space. In practice these may be partly occupied by insulation or thermal potting materials, however the electromagnetic properties of these are assumed to be close to those of free space. Furthermore, any electrically conducting structures such as mounting brackets or heat sinks are neglected as it is assumed that they will not affect the fringe field around the air gaps.

Ideally the free space should have an infinite size. In finite element analysis, it is modelled by a background region of air which encloses all the geometries. The background air will also fill any un-occupied gaps between the blocks/cells, such as the core window area, the air gaps between the core halves and the space between the coil and core.

The background volume needs to be large enough to achieve good accuracy but the drawback is the number of elements to be solved will be very big for large models. In this model, the surrounding air is defined as a large box which extends the original model dimensions by three to five times in each coordinate.

4.4.4 Model Symmetry and Boundary Conditions

Whilst the background air region significantly increases the model size, by exploiting model symmetry, only part of the model needs to be solved, allowing the model size as well as the computational requirements to be reduced.
4.4.4.1 Core symmetry and boundary conditions

Based on the geometry of the C-core, the model can be divided into eighths as shown in Figure 4-7, where the field conditions in each section are a mirror image of those in the neighbouring sections. The symmetry planes XY, YZ and XZ are shown in Figure 4-7 where the YZ plane passes vertically through the middle of the air gap.

When solving a one eighth segment of the component which is highlighted in Figure 4-7, appropriate boundary conditions must be applied to the symmetry planes, also called reflection planes, to define the model symmetry.

The directions of the main magnetic flux flow are marked with arrows in Figure 4-7. The magnetic flux lines are perpendicular to the YZ and XZ planes while tangential to the XY plane. Accordingly, the boundary conditions of the symmetry planes are defined, as shown in Figure 4-8, where only tangential or normal flux components pass through the reflection planes.

Figure 4-7. Core symmetry with symmetry planes (background air hidden)
4.4.4.2 Coil symmetry

The two racetrack conductors, one over each core leg, are also symmetrical in geometry. However, the coils in the model are not part of the finite-element mesh. Therefore, all conductor symmetries must be included in the model to form a complete set of coils such that the magnetic field generated can be properly represented [178].

4.4.4.3 Model to solve

Figure 4-9 shows the 1/8th model that is to be analysed including the core and coils after applying model symmetry and boundary conditions. The translucent grey regions represent the background air whilst the large grey box defines the solution space that needs to be meshed with finite elements. The other parts of the model have been effectively trimmed and removed from the simulation. Therefore, the model size as well as the computational cost is significantly reduced.

After the model has been solved, the conditions in the remaining parts of the model can be inferred based on the boundary conditions defined for reflection planes and the complete solution for the core can be created. This procedure only applies to the meshed
parts of the model that exclude the Biot-Savart conductors, which explains why the conductors must be modelled in full.

![Figure 4-9. 1/8th model to solve](image)

### 4.5 Finite Element Mesh

Having created the model structure, it must now be meshed to sub-divide the entire problem into a large number of small sub-domains, referred to as finite elements. 3D finite element meshing takes two steps. Firstly, all the surfaces of the model, both external and internal, are divided into smaller triangular or quadrilateral facets. After that, the volume mesh will be formed to create three-dimensional elements based on the surface mesh in the previous step. The generation of finite elements is important because the quality of the mesh is directly related to the accuracy and the running time of the simulation. A good finite element mesh should be fast to generate and the meshed model should also be fast to solve with high accuracy [177].
4.5.1 Mesh Control

The mesh can be controlled by varying the shape and size of the finite elements. For each block or cell in the geometric model, the mesh can be defined separately. However, the smaller the mesh size is, the better the solution will be, but smaller elements will lead to a larger number of equations to solve and more computer memory will be needed. The mesh therefore needs to be carefully controlled to achieve good accuracy and meanwhile limit the computational cost to a manageable level.

4.5.1.1 Mesh shape

Three-dimensional finite elements mainly have four shapes in the Opera software: hexahedra, prisms, tetrahedral and pyramids, as shown in Figure 4-10 [177]. Tetrahedral elements are most commonly used as they can be generated in cells of any shape, while hexahedra can only be applied to regular hexahedral blocks.

![Figure 4-10. Finite element types](177)
Mosaic meshing was introduced in the version 15 of the Opera software just around three years ago. The mosaic meshing allows different element shapes to be combined in a single mesh. However, it should be noted that for a 3D FE mesh using the mosaic elements, the element shape in one cell will affect the elements in the neighbouring cells. For example, a hexahedral meshed cell has quadrilateral facets on the surfaces while a tetrahedral meshed cell has triangular facets. When these two cells are placed next to each other, a surface mismatch will occur on the shared surface between both. In this case, prisms or pyramids can be used to transit the hexahedral elements to the tetrahedral elements [177].

4.5.1.2 Mesh size

The size of the finite elements determines the accuracy of the field calculations and for eddy current analysis, the mesh size should be no larger than half of the skin depth to capture the field variations due to the skin effect [147, 158, 179].

To avoid a large number of elements, the mesh needs to be fine in the parts of the model where a high level of accuracy is required and where there is a high spatial field variation, while for other parts of the model with smaller field variations, a coarse mesh can be implemented to save computer memory [177].

In this model, the regions around the air gaps where the fringing fields are concentrated must be assigned with the smallest mesh size to capture the details of the eddy current effects. And the element size can be gradually increased away from these problem regions, allowing the mesh to grade into the background air volume. The far end of the background region will have the maximum element size in the model.

Similar to the mesh shape, varying the mesh size in a cell may have significant implications on the element sizes in neighbouring cells of the model [177]. Figure 4-11a shows the mesh size transition in three hexahedral meshed cells. The mesh sizes are 10, 2 and 8 length units in the cells from left to right. The middle cell which has the smallest element size forces the interfaces between cells to be meshed with small quadrilateral
facets. As a consequence, the elements in left and right cells with larger mesh sizes have to be elongated in order to match the surface mesh on the interfaces. Such high aspect ratio elements are undesirable because they normally lead to poor quality finite element meshes with large solution errors [179].

To stop the transition of mesh size, the Opera user manual [177] advises that two tetrahedral meshed cells may be inserted between the hexahedral meshed cells to act as size buffers as illustrated in Figure 4-11b, and the size transition ratio between adjacent cells should be no more than 1:10. This is particularly useful for controlling the mesh in the modelling of the inductor core as different mesh sizes are needed in different regions of the model.
4.5.2 Model Optimisations for Mesh Generation

In the previous sections, the finite element mesh types and sizes have been described and the constraints on the mesh control have been discussed. Generally, the mesh is dependent on the geometry of the model. In addition, the element shape or size in one cell will affect the elements in adjacent cells. Therefore, it is important to specify appropriate meshes in different regions of a model to achieve fast and accurate solutions.

In section 4.4, the initial inductor model has been defined considering the model geometry only and it is not optimised for mesh generation. The initial model failed in the middle of the mesh generation process, mainly because the volume mesh generator failed to connect the common faces between a fine mesh and a coarse mesh. A large number of models have been developed, trying to improve the mesh quality by using various combinations of mesh shape and size.

According to the Opera user manuals [177, 178] and the experience gained during the modelling process, the model can be improved in the following aspects to achieve better mesh quality:

- Complicated cells should be broken down into several blocks of regular shapes, if possible. This will produce more hexahedral or prismatic meshable cells.

- Additional volumes should be used to break up the air gap and core window regions. This will allow additional control over the mesh in these areas and will also simplify the geometry of the background air region.

- Large volumes should be divided into smaller cells as they mesh faster.

- Extra cells should be used to create a gradual transition from a fine mesh to a coarse mesh.
The optimised model can be seen in Figure 4-12, where the additional gap volumes are highlighted. The core legs are cut into smaller blocks and the air volume in the core window is divided into three segments where the two thin layers near the gapped core legs are used to connect the fine mesh in the core legs and the coarse mesh in the middle of the core window. In Figure 4-13, the exterior of the core is coated with a thin layer of air comparable with the air gap length (translucent grey regions) to grade the fine mesh in the core into the coarse mesh in the background air.

Figure 4-12. Core model with additional air volumes filling the gaps and window regions

Figure 4-13. Core model coated with a thin layer of air
With these adjustments, the mesh quality has been significantly improved. The running time of the simulation has reduced from 10 hours for the first successful version of the model to less than 30 minutes for the present version.

### 4.5.3 Error Analysis

The effective skin depth for the homogenous inductor core has been calculated from equation (4-6) in Section 4.2 to be 1.2 mm at 60 kHz. Therefore, the mesh size for the gap regions should be no more than 0.6 mm in order to model accurately the flux distribution.

As mentioned earlier, the mesh quality has a strong implication on the accuracy of the analysis. To investigate the effect of mesh on the field solutions, the model has been meshed and solved with tetrahedral and mosaic elements of various sizes to compare the errors in the results.

#### 4.5.3.1 Effect of mesh shape on field solutions

There is no general rule to determine the element types because the choice of element shapes is strongly limited by the geometry of the model. The generation of mesh is an automatic process in Opera. For the mosaic mesh generator, the hexahedral mesh takes the highest priority in the mesh generation, while for cells that are not hexahedral meshable, for example the core corners, other element shapes will be used.

Figure 4-14a shows the model meshed with purely tetrahedral elements, while Figure 4-14b shows the model with mosaic meshing. The mesh size arrangements for the two models are the same: 0.5 mm around the gap areas and the maximum element size is limited to 5 mm at the outer model boundaries.

During the simulation, the software will report a RMS error in the field solution by calculating the difference between the nodally average field and the value calculated
directly from the element solution potentials for each finite element. The RMS error can then be weighted by the volume of the elements to indicate the effectiveness of the model [178].

Figure 4-14. Meshed models (expanded to show around core regions only)
These two models are solved and some key simulation parameters are compared in Table 4-3 to assess the two meshing methods. It can be seen that the mosaic meshing has several advantages compared to tetrahedral meshing.

Table 4-3. Comparison of tetrahedral and mosaic meshing (frequency: 60 kHz, minimum element size: 0.5 mm, maximum element size: 5 mm)

<table>
<thead>
<tr>
<th>Simulation parameters</th>
<th>Tetrahedral meshing</th>
<th>Mosaic meshing</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of elements</td>
<td>1,659,364</td>
<td>549,137</td>
</tr>
<tr>
<td>Number of equations</td>
<td>2,106,718</td>
<td>835,376</td>
</tr>
<tr>
<td>Database size</td>
<td>442.2 MB</td>
<td>210.8 MB</td>
</tr>
<tr>
<td>Running time</td>
<td>00:25:49</td>
<td>00:08:44</td>
</tr>
<tr>
<td>RMS error in $B$ field</td>
<td>40.177 %</td>
<td>8.983 %</td>
</tr>
<tr>
<td>Weighted RMS error in $B$ field</td>
<td>3.94 %</td>
<td>0.605 %</td>
</tr>
<tr>
<td>RMS error in $J$ field</td>
<td>85.160 %</td>
<td>0.775 %</td>
</tr>
<tr>
<td>Weighted RMS error in $J$ field</td>
<td>60.298 %</td>
<td>0.794 %</td>
</tr>
</tbody>
</table>

Firstly, the scale of the problem is significantly reduced when using mosaic elements. The total numbers of elements and equations are reduced by more than half. The reason is that mosaic meshing involves some hexahedral elements and the model requires fewer hexahedral elements than tetrahedral elements of the same size, which is evident in Figure 4-14. This will also reduce the solution time.

Secondly, for hexahedral elements, the directions of fields are normally aligned with the element edges, making them superior to the tetrahedral elements in terms of solution accuracy [179]. This can be seen in Figure 4-15, where the solution looks blotchy for the model with tetrahedral elements, and on the contrary, the field presents a smooth field variation for the mosaic meshed model.

From Table 4-3, the model with mosaic meshing has more accurate results and is faster to solve, which is in line with the Opera manuals. However, the main limitation of the mosaic mesh is that it is less controllable as the choice of mesh shape is constrained by
the model cell topologies. In this work, the mosaic mesh is used. The areas of interest (regions around the air gaps) are meshed with fine hexahedral elements that have better accuracy and with a smaller number of elements, whilst the core corners and background air regions use tetrahedral elements with larger sizes.

Figure 4-15. Amplitude of flux density in the core at 60 kHz using the two meshing methods (minimum element size: 0.5 mm, maximum element size: 5 mm)
4.5.3.2 Effect of mesh size on field Solutions

To investigate the effect of mesh size on the accuracy of the field calculations, the model has been meshed with mosaic elements of different sizes and the results are compared in Table 4-4. The ‘mesh size’ in the table refers to the smallest element size that is used, which is around the air gaps. The maximum element size is limited to 5 mm in all cases. The results show that the solution errors in the flux density, $B$, and current density, $J$, are lower with a finer mesh. But reducing the mesh size will require more computational effort.

Table 4-4. Comparison of field calculations with different mesh size (mosaic meshing, frequency: 60 kHz)

<table>
<thead>
<tr>
<th>Mesh size (mm)</th>
<th>Weighted RMS error in $B$ field</th>
<th>Weighted RMS error in $J$ field</th>
<th>Solution time</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.3</td>
<td>0.272 %</td>
<td>0.327 %</td>
<td>00:15:12</td>
</tr>
<tr>
<td>0.5</td>
<td>0.605 %</td>
<td>0.794 %</td>
<td>00:08:44</td>
</tr>
<tr>
<td>1</td>
<td>13.117 %</td>
<td>4.315 %</td>
<td>00:03:40</td>
</tr>
<tr>
<td>1.5</td>
<td>21.369 %</td>
<td>6.722 %</td>
<td>00:01:35</td>
</tr>
<tr>
<td>2</td>
<td>28.806 %</td>
<td>9.727 %</td>
<td>00:00:20</td>
</tr>
</tbody>
</table>

To calculate the gap loss, the element size of the concerned areas will be set to 0.3 mm, which is about 1/4 of the equivalent skin depth at the intended operating frequency, ensuring accurate calculation of the eddy currents caused by the fringing flux.

4.6 Conclusion

A FE electromagnetic model has been developed in the Opera software to calculate the loss associated with the air gap fringing flux in a finely laminated tape-wound core. The key objective is to accurately model the eddy currents in the lamination planes. To do this, homogenisation modelling is employed, which uses solid continuums with anisotropic properties to represent the tape wound core. The geometric modelling of the core and winding are described for a foil-wound C-core. The model is based on the dimensions of
a F3CC0032 Finemet core, however the dimensions can be easily changed to examine components of similar structure, but different size. Furthermore, the modelling technique can be easily applied to other inductor designs.

The choices of finite-element shape and size have been discussed and their effects on the simulation errors have been investigated. Compared with the traditional tetrahedral mesh, mosaic meshing that combines different types of elements is preferable as it leads to faster and more accurate solutions. The mesh control has some constrains mainly due to the geometrical topology of the model. Accordingly, the initial inductor model has been improved to allow better control over the mesh quality.

The model developed in this chapter will be used to compute the gap loss and understand its distribution in the component in Chapter 5.
CHAPTER 5

5 FEA GAP LOSS ANALYSIS

5.1 Introduction

In this chapter, the gap loss model described in Chapter 4 is used to analyse the gap losses in a foil wound nanocrystalline inductor. The post-processing of the FE model results will be undertaken to understand the magnetic field and loss distributions in the core. The loss distributions can then be used in a thermal analysis to identify the core’s hot spot and therefore form a basis for thermal management of the inductor.

The simulations show that the gap loss is considerable and highly localised, which may result in significant temperature rises around the air gap regions, therefore, special attention is needed to avoid local overheating.

A sensitivity study has been performed to explore the dependency of the gap loss on the key inductor design parameters, such as operating frequency, gap length, ac flux density and core strip width. Using the results, an empirical equation is derived to provide a convenient design-oriented basis for estimating gap losses.

5.2 Inductor Model Configurations

The results presented in this chapter are based on the modelling of an inductor designed by Dr Gerardo Calderon-lopez for a 25 kW, 30 kHz, dual-interleaved DC-DC converter. The inductor used a Finemet cut core, F3CC0032, wound with 18 µm laminations. The two core halves were placed 2.2 mm apart to form a total gap length of 4.4 mm. The core dimensions and material properties are provided in Chapter 4 (Table 4-1 and Table 4-2).
The winding comprised 6 turns of 0.8 mm thick and 51 mm wide copper foil and was split between the two gapped core limbs with 3 turns around each. The inductance was measured to be 5.4 µH at 60 kHz by a LCR meter. The winding is represented by two single-turn racetrack Biot-Savart conductors of the same width in the model, carrying a sinusoidal current with an equivalent peak current density.

The overall simulation results will be presented in 3D graphs and plots. However, apart from the 3D core view, several 2D patches and lines have been created and shown in Figure 5-1 to help understand the field and loss distributions along the core surfaces and in the middle of the core. Due to the model symmetry, only the upper core leg will be examined. The lower core leg will be a mirror image of the upper leg.

The legends used are explained as follows:

- Cutaway view: a X-Y patch in the centre of the core leg that cuts through all lamination layers, highlighted in Figure 5-1a (core legs are made translucent to show the cutaway patch).
- Outer surface view: the outermost lamination layer, up to 15 mm away from the gap, viewed from +Y, Figure 5-1b.
- Inner surface view: the innermost lamination layer, up to 15 mm away from the gap, viewed from -Y, Figure 5-1b.
- Gap face view: the surface of the core end at the air gap, viewed from +X, Figure 5-1b.
- Centre lines: lines along core surfaces at the middle of the core strip width, \( z = 0 \), Figure 5-1b.
- Parallel lines: lines along the core surfaces, parallel to the core edge at the gap, with a distance, \( d_g \), from the gap, Figure 5-1b.
This section presents the electromagnetic FE solutions of the gap loss calculations. The flux density distribution in the core is studied for the inductor specified in Section 5.2. The model also solves the induced eddy currents in the lamination planes as well as the distributions of the associated losses. The loss distribution profile is then imported to a simple FE thermal analysis to estimate the temperatures in the core in order to identify the hot spot.
5.3.1 Flux Distribution

To analyse the gap loss, the magnetic flux within the core is examined first, especially the flux component that is perpendicular to the lamination planes. Since the overall magnetic field in the core is a combination of the source field created by the winding and that created by the induced eddy currents, this section first studies the source field which generates the eddy currents. In order to isolate this field, the eddy current effects are excluded by setting the electrical conductivity of the core material to zero so that the flux in the core is due only to the source windings.

5.3.1.1 Solid model and homogenous model

First, the flux density distribution in a solid core model is compared with a homogenous model to investigate how laminations will affect the magnetic field in the core. The solid model used isotropic permeability so the laminations are not considered, while the homogenous model employs anisotropic equivalent permeabilities so that the magnetic flux tends to flow along the direction of the laminations, as explained in Chapter 4.

Figure 5-2 shows the flux distributions in the core with no air gap. In the solid core model, the flux densities are almost uniform in the core legs covered with the coils but the fluxes are high around the corners of the core window as this path provides the minimum reluctance/path length, Figure 5-2a. Whilst in the homogeneous core model, Figure 5-2b, the flux densities are high along the inner lamination layers as this is the shorter magnetic path and reduces towards the outer laminations throughout the core, and this is due to the low effective permeability through the lamination planes.

Figure 5-3 shows the flux density distributions in the core with air gaps. Similar to the gapless core, the flux is high around the corners of the core window in the solid core model, Figure 5-3a. However in the homogenous model, the flux densities are higher towards the outermost and innermost lamination layers due to the air gap fringing flux at the surfaces, and because of the low permeability through the laminations the high flux
densities are confined within a few layers close to the core surfaces. The surface flux density is more than twice the magnitude of the flux in the middle of the core, as seen in Figure 5-3d. It is also noticed that the flux density has a larger magnitude at the outer core surface, shown as the top edge in Figure 5-3d. This is because a larger tangential flux component occurs at the top edge, whilst the lower edge which represents the inner core surface experiences an overall larger perpendicular flux, and this will be described in Section 5.3.1.2.

The simulations have shown that the flux density distributes unevenly in the core. Therefore, it is possible that the core experiences some localised saturation while the average flux density is still within the linear range. Furthermore, as the power losses are directly related to the flux density, the losses in the core will also be non-uniform.

![Flux Density](image)

**Figure 5-2. Flux distributions in the core without air gaps**

- a. Solid model
- b. Homogenous model
5.3.1.2 Perpendicular flux component

This section looks at the flux components perpendicular to the lamination planes that causes the gap loss to occur. The distribution of the perpendicular flux will approximately indicate the location of the gap loss.

Figure 5-4 shows the cutaway view of the perpendicular flux distributions through the middle of the upper core leg. The perpendicular component is zero in the middle of the core limb and is largest at the core edges on the surfaces.
Figure 5-4. Cutaway view of perpendicular flux distribution

Figure 5-5 shows the perpendicular flux component reducing as the measurement point moves away from the core edge into the air. The result is consistent with the results in [133], in which the winding loss due to the air gap fringing field is calculated using a 2D finite element analysis. Also in [83], it was concluded that the windings should be spaced at least twice the gap length (per single gap) away from the core to minimise the additional winding loss. From Figure 5-5, the perpendicular flux component reduces to about 10% of its peak value at $l_g/2$ (gap length per core leg) away from the core edge, and to almost zero at a total gap length, $l_g$, away from the core, showing a good agreement with the publications.

Figure 5-5. Perpendicular flux distribution away from the gap edge, $l_g = 4.4$ mm
If the windings are placed closer to the core, the fringing field distribution will be altered due to the effects of induced eddy currents in the windings. As a result, the winding will suffer additional losses. However, this will result in a reduced perpendicular component of flux at the core surface and a reduced gap loss in the core. This project is focused on the analysis of gap loss. The effect of additional winding losses due to the gaps is not considered based on the assumption that a sufficient spacing between coil and core is maintained which is the commonly used method in practice to minimise the high frequency winding losses.

Figure 5-6 shows the surface views of the perpendicular flux distribution, the top plot is looking down on the outer core surface, and the middle plot is looking upwards on the inner surface whilst the bottom plot is looking directly at the gap face.
Figure 5-7 plots the perpendicular flux along the centre lines on the outer and inner core surfaces, which are the top and bottom edges shown in Figure 5-4. It is noticed that there is a slight difference between the fluxes in the inner and outer core leg surfaces, which also can be seen in Figure 5-4.

![Figure 5-7. Perpendicular flux distributions along the central lines on outer and inner core surfaces](image)

The fringing flux has zero perpendicular components in the centre of the air gap and the highest values occur at the edges of the gap. The outermost lamination sheet has a slightly higher peak value of perpendicular flux than the inner layer, while the inside of the core has a broader flux distribution as the perpendicular component is larger in the inner surface away from the gap.

By integrating the flux density over the core surfaces, the inner core surface is subject to a total perpendicular flux of almost 20 % greater than the outer surface. This effect cannot be identified when modelling only a small segment of the core, as seen in the models described in the literature [144, 147, 170] in which the flux is normally defined by either boundary conditions or a uniform external field. However, by constructing the whole core geometry and conductor windings, the field variations in different parts of the core can be identified.
At the outside of the core, the flux fringes into the air only, while in the inside, there will be flux striking into the short side of the core window (bottom of the U-shape) as well as the opposite core leg, creating more perpendicular flux through the inner surface. The perpendicular component has very low magnitude away from the gap, less than 8% of the peak value at 15 mm from the core edge. Therefore the eddy currents induced in the short core legs at the bottom of the U-shape are neglected. The gap loss due to the perpendicular components is assumed to be in the gapped core legs only.

What also can be seen from the surface views shown in Figure 5-6 is that the maximum perpendicular flux experienced by the core is around the corners of the lamination sheet near the gap. In addition, there is a high perpendicular component at the side edges of the core strips. Figure 5-8 plots the perpendicular flux along a line parallel to the gap edge, 5 mm away from the gap, along the Z-axis. The higher perpendicular flux on the inner surface is also evident.

![Perpendicular flux distributions along surface lines parallel to the gap edge, \( dg = 5\text{mm} \)](image-url)
5.3.2 Eddy Current Distribution

To analyse the in-plane eddy currents, the anisotropic conductivities of the core material are restored in this section, which are listed in Table 4-1 in Chapter 4. The conductivity along the lamination planes (X and Z axes) is $6.67 \times 10^5$ S/m and through the lamination planes (Y-axis) it is 0.46 S/m in the gapped core legs.

Figure 5-9 shows the current density through the middle of the core leg. There are no discernable eddy currents in middle of the core where the flux has almost zero perpendicular components. The eddy currents are concentrated at the core surfaces.

![Figure 5-9. Cutaway view of eddy current distributions](image)

The eddy current distributions are consistent with the perpendicular flux distributions seen in Section 5.3.1.2. The eddy current densities are highly localised and occur towards the edges of the cores, particularly along the top and bottom air gap edges. The surface plots in Figure 5-10 show very similar patterns looking down on the outer surface of the core (top plot) and looking up on the inner surface (middle plot), but with slightly higher eddy current levels at the inner surface. The maximum current densities occur towards the centre of the air gap edges and have values greater than 30 A/mm$^2$. The current densities are lower at the core corners, but increase along the side edges with maximum...
values in the region of 15 A/mm$^2$. The lower plot in Figure 5-10 looking at the cut face of the core at the air gap shows that the eddy currents are concentrated within around 1.5 mm of the core surface, which corresponds to the calculated effective skin depth, 1.2 mm, for the homogeneous core model.

These observations are confirmed by the plots in Figure 5-11 showing the surface eddy current distributions along the lines parallel to the gap edge 5 mm from the gap. Figure 5-12 shows the surface eddy current distribution along the centre line of the core moving away from the gap edge and shows that the current density falls to approximately zero at around 7-8 mm from the gap, and then increases due to the eddy current loops.
Figure 5-11. Eddy current distributions along surface lines parallel to the gap edge, $d_g = 5$ mm

Figure 5-12. Eddy current distributions along centre lines on core surfaces
5.3.3 Gap Loss Distribution

The loss density is calculated from time-averaged $J \cdot E$ in the simulations, where $J$ and $E$ are the vectors of current density and electric field strength, respectively [180]. Figure 5-13 shows the gap loss density in the core and the surface plots are shown in Figure 5-14.

The distribution of the gap loss largely corresponds to the eddy current density distributions. The maximum loss densities occur towards the centre of the gap edges with a peak value of around 0.8 W/mm$^3$. Moving away from the centre of the gap edges, the loss density falls to almost zero at around 4-5 mm. The loss densities along the side edges of the core have the maximum value of around 0.2 W/mm$^3$, which occur at 5-7 mm away from the gap edges. The inner core surface (middle plot in Figure 5-14) has a slightly broader loss distribution compared with the outer surface (top plot), and this is explained by the larger perpendicular flux distribution in the inner core surfaces, as shown in Section 5.3.1. The bottom plot in Figure 5-14 shows the loss distribution in the gap face where the loss densities are concentrated within around 1 mm of the top and bottom gap edges.

![Figure 5-13. Gap loss distributions, core view](image-url)
The concentration of the gap loss is remarkable, especially on the surfaces and along the core gap edges. The gap loss is therefore likely to be affected by the core strip width. However, the penetration of loss is insignificant, limited about 1 mm into the core, owing to the low conductivity through the laminations. Also the loss reduces sharply moving away from the gap. It is therefore assumed that the gap loss caused by the fringing flux is relatively independent of the length and height of the core leg. This implies that the gap loss can be reduced by reducing the width of the lamination strips. For example, the design example shown here has a rectangular shape for the core gap face with core strip width (30 mm) larger than the core leg height (13 mm). By reducing the core strip width...
but meanwhile increasing the height in the core leg, the same core cross sectional area can be maintained but the gap loss is likely to be reduced.

The distribution of gap loss density will depend on operating conditions and the air gap lengths. Simulations are undertaken to investigate the effects of frequency, AC induction levels and the air gap length on the gap loss distributions. The different induction levels are provided by altering the AC current excitations in the conductor windings.

Figure 5-15 and Figure 5-16 show the gap loss densities along the centre lines on the outer core surface when moving away from the gap edge at different frequency and excitation conditions. The gap losses are increased with increasing frequency and induction levels. However, varying these conditions does not change the distribution of loss on the core surfaces. When normalised to the peak value that occurs at each condition, the loss distributions appear to be identical.

Figure 5-15. Effect of frequency on gap loss distributions along centre lines on the outer core surface (AC current excitation unchanged, $lg = 4.4$ mm)
Figure 5-16. Effect of excitation level on gap loss distributions along centre lines on the outer core surface (frequency: 60 kHz, $l_g = 4.4$ mm)

Figure 5-17 shows the gap loss densities along the centre line on the outer core surface with varying air gap length. The peak loss density occurring at the gap edges also increases with increasing gap length, but a more concentrated loss distribution is observed for smaller gaps, as shown in Figure 5-17b which plots the loss density normalised to the peak value at the gap edge. The more concentrated loss density towards the gap edge with smaller gaps is directly related to the more concentrated perpendicular flux component around smaller gaps.

The variation of gap loss with the core strip width, frequency and other model parameters will be quantified in later sections.

The total gap loss in the core can be calculated by integrating the loss density over the core volume, $\int (J \cdot E) dV$ [180]. At 60 kHz, the fringing flux causes a power loss of 44 W at a peak flux density of 0.14 T. The hysteresis loss measured at the same condition is 5 W. The gap loss therefore contributes a major part of the inductor core loss for the design example considered here.
Furthermore, the hysteresis loss is distributed throughout the entire volume of the core whilst the gap loss is highly concentrated. By calculating the loss in different core sections, it is found that the majority of the gap loss, nearly 45% of the total loss, is
located around the core edges at the gap. About 15% of the loss is distributed along the core side edges. The rest of the loss is spread around the other parts of the core leg surfaces. Detailed loss distribution data in different core sections are provided in Table 6-5 in Chapter 6 for the gap loss allocation in a thermal analysis. The gap loss is therefore responsible for the largest heat flow that occurs in the core in this design, in particular near the regions around the gaps.

### 5.3.4 Temperature Distribution

The Opera 3D FEA package allows a coupled analysis that is capable of transferring the loss distribution from an electromagnetic model to a thermal one to solve for the associated temperature rises, providing the mesh used for both models is identical.

To predict the core’s hot spot, the loss density distribution shown in Figure 5-13 has been imported to the thermal analysis. The temperature distribution result is presented in Figure 5-18, showing the steady-state core temperatures due only to the gap loss. The thermal conductivities used are 10 W/mK along the laminations and 0.5 W/mK through the laminations [61, 160] to account for the poor heat transfer through the insulation.

![Temperature distribution in the core in free air (Ambient: 30 °C)](image)
Intense heating occurs in the air gap regions caused by the localised gap loss, in particular the inner gap edges. Without thermal management the high temperatures may damage the insulation between core laminations. This temperature distribution is not practical as it only shows the temperature rise caused by the gap loss in free air. However in reality, core hysteresis loss and the winding around the core will increase the heat dissipation and also in the case of the winding affect the cooling paths. This simulation only indicates the possible core hot spot owing to the gap loss, and where additional thermal paths may be required to improve the cooling of the component. A full thermal model including all core and winding losses and thermal management is described in Chapter 6.

5.4 Gap Loss Prediction

In Section 5.3.3, the effects of operating conditions and gap length on the distribution of gap loss on core surfaces have been discussed. This section will investigate the sensitivity of the gap loss magnitude to the inductor design parameters using FEA simulations. The objective is to derive a model-based equation to estimate the total gap loss in inductors using nanocrystalline C-cores.

The preliminary assumption is that the gap loss can be expressed as a power function of the total gap length, \( l_g \), core strip width, \( W_{\text{core}} \), frequency, \( f \), and peak AC induction, \( B_m \), as in equation (5-1), which is in line with the published equation (2-16) [25]. In (5-1), \( k_g \) is a numerical constant, and \( k_{lg} \), \( k_{W_{\text{core}}} \), \( k_f \) and \( k_{B_m} \) are assumed to be constants independent of each other representing the sensitivity of the gap loss to the inductor parameters. The values of these constants are determined through the sensitivity analysis.

\[
P_g = K \left( l_g, W_{\text{core}}, f, B_m \right) \\
= k_g \cdot l_g^{k_{lg}} \cdot W_{\text{core}}^{k_{W_{\text{core}}}} \cdot f^{k_f} \cdot B_m^{k_{B_m}}
\]

The packing factor, \( F \), defines the percentage of core material in the component and accounts for the insulation layers between magnetic ribbons. The homogenisation of the
FE model uses the packing factor to calculate the effective electrical and magnetic properties of the core, and therefore the gap loss is likely to be related to the value of $F$. However, the packing factor of commercial nanocrystalline cores is nearly always around 0.8 [181] so the variation of gap loss with packing factor is not considered here. The packing factor is implicit in the constant $k_g$ in equation (5-1).

5.4.1 Sensitivity Analysis

The sensitivity analysis is performed by varying one inductor design parameter whilst the others are kept constant. The dependency of the loss on each parameter is expressed by curve-fitting the calculated gap losses from FEA with respect to the parameters.

5.4.1.1 Total gap length, $l_g$

To study the gap loss variation with gap length, the total gap length was changed from 2 to 5.2 mm distributed equally between the core legs, while the other parameter remained the same.

The calculated magnitudes of gap loss are plotted against the gap length in Figure 5-19. The loss increases with increasing gap length in a linear manner and the trend lines of the plot extend back to the origin, indicating that the gap loss is zero when there is no air gap, which therefore may be expressed by (5-2), where $K_{lg}$ is the slope of the fitted lines and the exponent $k_{lg}$ is equal to 1. The value of $K_{lg}$ will be dependent on other inductor parameters, $W_{core}$, $f$ and $B_m$. For example, as shown in Figure 5-19, the proportionality between gap loss and $l_g$ is maintained at different frequencies, but $K_{lg}$ will be changed to reflect the different gradient of the lines. A similar pattern is seen with the other parameters.

$$P_g = K_{lg} l_g^{k_{lg}}$$  \hspace{1cm} (5-2)
5.4.1.2 Core strip width, $W_{\text{core}}$

The dependency of gap loss on the core strip width is evident in the loss distributions on the core surfaces where the loss is localised along the width of core laminations around the gap. To examine the effect of different strip widths, four values were simulated: 20, 25, 30 and 35 mm.

Figure 5-20 shows the variation of total gap loss with the core strip width in a log-log plot. The loss, $\log(P_g)$, increases linearly with the core strip width, $\log(W_{\text{core}})$, in the log-log scale, which may be represented by equation (5-3), where $k_{W_{\text{core}}}$ is the slope of the fitted lines and $\log(K_{W_{\text{core}}})$ represents the effects of the other parameters. Figure 5-20 shows that with different frequencies the slopes of the lines are unchanged, but are just moved vertically, represented by a different value of $K_{W_{\text{core}}}$ in (5-3). A similar pattern is seen for different values of the other parameters. Therefore, the gap loss is considered to be proportional to $W_{\text{core}}^{k_{W_{\text{core}}}}$, as shown in (5-4), where the slope, $k_{W_{\text{core}}}$, has the value 1.65.

$$\log(P_g) = k_{W_{\text{core}}} \log(W_{\text{core}}) + \log(K_{W_{\text{core}}})$$ \hspace{1cm} (5-3)

$$P_g = K_{W_{\text{core}}} W_{\text{core}}^{k_{W_{\text{core}}}}$$ \hspace{1cm} (5-4)
5.4.1.3 Frequency, $f$

The simulation frequency was ranged from 40 to 200 kHz, covering the typical operating frequencies of the inductor in the dual-interleaved converter for EV applications.

Figure 5-21 shows a log-log plot of the gap loss against frequency at different induction levels. Similar to the core strip width, the fitted curves are straight lines in the log-log scale, which may be represented by (5-5), and in turn the variation of gap loss with frequency could be expressed by (5-6). Varying the flux density, $B_m$, moves the curves vertically, while the slope of the curves, $k_f$, is constant. A similar pattern is seen for different values of the other parameters. Therefore, the gap loss may be considered to depend on frequency as a power function, (5-6), where $k_f$ is determined from the slope of the fitted lines having the value 1.72.

\[
\log(P_g) = k_f \log(f) + \log(K_f) \tag{5-5}
\]

\[
P_g = K_f f^{k_f} \tag{5-6}
\]
5.4.1.4 Peak flux density, $B_m$

The peak AC flux density is varied from 0.1 to 0.2 T by changing the current density in the Biot-Savart conductors that provide the excitation. As the flux density distributes unevenly in the core, the peak flux density used here is the average flux density across the core cross sectional area.

Figure 5-22 shows a log-log plot of gap loss against the peak flux density. The gap loss increases with increasing flux density, and the linear relation shown in the figure may be expressed by (5-7). Varying the gap length varies the value of $\log(K_{Bm})$ moving the lines vertically, but does not change the slope $k_{Bm}$. A similar pattern is seen for different values of core strip width and frequency. Therefore the gap loss may be considered to be proportional to $B_m^{k_{Bm}}$, (5-8), where $k_{Bm}$ has a constant value of 2, however $K_{Bm}$ is dependent on the values of $l_g$, $W_{core}$, and $f$.

$$\log(P_g) = k_{Bm} \log(B_m) + \log(K_{Bm}) \quad (5-7)$$

$$P_g = K_{Bm} B_m^{k_{Bm}} \quad (5-8)$$
5.4.2 Gap Loss Prediction

Based on the results from the sensitivity analysis, the values of the coefficients $k_{lg}$, $k_{W_{core}}$, $k_f$ and $k_{Bm}$ have been determined, and the empirical equation to estimate the total gap loss in the Finemet core can be written as:

$$ P_g = k_g \cdot l_g \cdot W_{core}^{1.65} \cdot f^{1.72} \cdot B_m^2 $$  \hspace{1cm} (5-9)

where $k_g = 1.68 \times 10^{-3}$, $l_g$ and $W_{core}$ are in mm, $f$ in kHz and $B_m$ in Tesla.

The calculated gap loss using the (5-9) is compared with the published formula (2-16) in Figure 5-23 at two values of $W_{core}$. Equation (5-9) shows a steeper loss variation with frequency and core strip width whilst the published one expresses the gap loss as a linear function of frequency and core width.
In equation (2-16), the numerical constant $G$ in front of the formula has different values for different coil and core arrangements. For example, the gap loss is approximated to be doubled for a one-coil configuration.

The one-coil inductor configuration has been examined using the FE model, where only one Biot-Savart conductor with doubled amp turns is used around the upper core leg. The imbalance of the inductor configuration requires the model symmetry to be doubled, consequently doubling the model size.

Figure 5-24 shows the simulated flux distribution with all the amp turns around the upper core leg. The core leg without a coil is subjected to a lower induction than the core leg covered with a coil due to the leakage flux.

Figure 5-25 shows the gap loss distribution in the core using the one-coil configuration. The total gap loss in the one-coil design is about the same as that in a two-coil configuration, but the loss is distributed unevenly between the two core legs. The core leg with the coil has a larger peak loss density of around $1.2 \text{ W/mm}^3$ at the centre of the gap.
edges, and the uncovered core leg has a lower peak value of around 0.4 W/mm$^3$. About 60% of the gap loss (27 W) occurs in the core leg covered with the coil and 40% (17.5 W) in the uncovered one. Whilst for the two-coil design where the total amp turns are divided equally around the two core legs, the peak loss densities in both core legs are around 0.8 W/mm$^3$, as shown in Figure 5-13 in Section 5.3.3.

![Flux Density (T)](image)

a. Core leg covered with a coil  
b. Core leg without coil

Figure 5-24. Cutaway views of flux distributions with only one coil (average flux density: 0.14 T)

![Gap Loss Density](image)

Figure 5-25. Gap loss distribution with only one coil around the upper core leg
A single-coil winding is cheaper to manufacture, however it is not an optimum design in terms of electromagnetic and thermal performance. The core leg wound with the winding suffers from more gap loss than the other core leg, potentially resulting in a higher hot spot temperature. In addition, the total copper loss is also located around this side of the core, adding to the thermal management challenge in the covered core leg. Furthermore, a single coil containing all the turns is larger in size therefore is likely to increase the overall dimension of the component.

Whilst for a two-coil design, the total number of turns is split equally around both core legs without a particularly large coil diameter, and therefore the packaging requirement is reduced. The construction of the two-coil winding requires extra manufacturing procedures as two coils need to be wound and they must be interlinked. However, the total loss is divided equally between two core legs and therefore the temperature distribution is likely to be symmetrical throughout the core with a lower peak value than the unbalanced design. The overall performance of a two-coil design is superior to the one-coil configuration. Therefore the gap loss in a one-coil inductor is not investigated any further in this research.

The proposed equation (5-9) is derived for a Finemet inductor design where two foil-wound coils are wrapped around each limb of the C-core and the coils are placed a sufficient distance away from the core to minimise the fringing flux striking the foil. The approximation identifies the effects of the inductor design parameters on the total gap loss therefore provides additional loss information for future design optimisations. For other inductor designs or core materials, the gap loss could be approximated following the same approach.

5.5 Conclusion

The magnetic field and eddy current loss distributions in a Finemet C-core have been analysed using the electromagnetic FE model described in Chapter 4. The results have shown a particularly non-uniform flux and loss distribution in the component. The gap
loss is significant in this design example and the loss is highly localised, especially around the core edges at the gaps, which may lead to thermal failure of the component.

The effects of the inductor design parameters on the gap loss have been identified by a sensitivity study via simulations. An empirical expression has been derived from the curve-fitting of the results to estimate the gap loss in the foil-wound Finemet inductor. Compared with the previous equation (2-16) proposed by Lee, the derived loss expression shows the gap loss in high frequency cores depending on the core width and frequency to the power 1.65 and 1.72 respectively as opposed to linearly. The calculated gap loss and its distribution will be used in Chapter 6 to predict the core’s hot spot, allowing the inductor design to be optimised for a size and weight reduction.
CHAPTER 6

EXPERIMENTAL VALIDATION OF FE GAP LOSS MODELLING

6.1 Introduction

This chapter validates the FE gap loss model by using a nanocrystalline inductor. The inductor was designed for a DC-DC converter using Finemet core and foil winding, and was encapsulated in an aluminium case to enhance heat transfer. The experimental tests are used to verify the accuracy of the gap loss calculations and also to examine the thermal performance of the test inductor, therefore identifying possible solutions for future designs.

The FE gap loss model is validated by temperature measurements in two steps. Firstly, the total loss in the inductor is measured and compared with the calculations. Secondly, the steady-state temperatures are measured at different locations on the core surface to verify the loss distributions. The measured temperatures are then compared with a FE thermal analysis which uses the non-uniform gap loss distribution from the FE electromagnetic solutions to estimate the steady-state temperature in the component.

A 70 °C temperature rise relative to the heat sink was found to occur at the core corner near the gap under the rated operating condition. The temperatures are lower away from the gap. The non-uniform temperature distribution in the core corresponds to the predicted localised gap loss distribution. Furthermore it is shown that the hot spot temperature is reduced by 20 °C by using high-thermal-conductivity ceramic heat spreaders in the air gaps, potentially allowing a smaller core to be used.
The measurement results show good agreements with the loss calculations and FE thermal analyses. The likely causes of discrepancies between measurements and FEA results are described.

6.2 Experimental Set Up

6.2.1 Converter

The converter used in the tests is a 25 kW, 300 A, bi-directional dual interleaved DC-DC converter with IPT. The specifications are a minimum input voltage of 75 V, maximum output voltage of 275 V and switching frequency, $f_{sw}$, of 30 kHz. The converter was designed and built by Dr Gerardo Calderon-Lopez as part of a separate research project [51].

6.2.2 Test Inductors

The inductor was designed by Dr Gerardo Calderon-Lopez as the input inductor for the converter specified above. The component was designed to be compact, but this results in a significant gap loss in the core. The core material used and the core dimensions have been described in Section 5.2. Some other features of the component that are related to the thermal validations are provided here.

The air gaps between the core halves, 2.2 mm in each core leg, were set by plastic shims and glass reinforced plastic (GRP) tiles. The plastic spacers were trimmed to fit the dimension of the gap area. Several thermal sensors were fixed to the core surfaces and embedded in the coil to measure the temperature distribution in the inductor. The locations of the thermal sensors are shown in Section 6.4.2.

The winding was wound with copper foil insulated by Kapton HN tape and was equally split between the two core legs. No coil bobbin was used, but small spacers were inserted
between the winding and core to ensure that the winding was at least half the total gap length from the core to minimise the winding losses due to the air gap fringing flux. The placement of two coils was not perfectly balanced during assembly. The innermost foil layer was 2.2 mm from the core inner surface and 3.2 mm from the core outer surface. According to the result presented in Figure 5-5 in Section 5.3.1.2, at these distances the perpendicular component of the fringing flux falls to 10% and 6% of the peak value that occurs at the core gap edges. The implications of the coil-to-core distance on the measurement results will be discussed in later sections.

To improve the thermal performance, the core and windings were encapsulated in a 2 mm thick aluminium case using a thermally conductive epoxy potting compound. The top part of the winding is exposed outside the potting to allow external circuit connections. The aluminium case was milled from a single piece of metal to provide a uniform path for thermal conduction. The space between the component and the aluminium case was filled with thermally conductive epoxy resin potting material to create a thermal path between the inductor and the case. The potting material will also provide electrical isolation and extra mechanical support for the component. A photograph of the potted inductor is provided in Appendix B.

The potted inductor was initially mounted to the converter’s main water-cooled cold plate. In this chapter, the component was separated from the converter’s cooling system and bonded on a small extruded aluminium heat sink [182] cooled by a 12 V fan [183], Figure 6-1. The heat sink and fan assembly was characterised thermally in a separate test with DC power resistors in Section 6.2.4. Superwool thermal insulation [184] was used to seal the entire inductor during tests to ensure that all the heat was dissipated by conduction through the heat sink.

A second inductor, also built by Dr Gerardo Calderon-lopez, was also used in the validation work. The second inductor was identical to the first one except for the air-gap spacers. Instead of plastic spacers with low thermal conductivity (0.35 W/mK), high-thermal-conductivity aluminium nitride (AlN) heat spreaders (90 W/mK) were fitted
between the core halves to maintain the gaps. The ceramic heat spreaders are not electrically conductive but will improve the heat flow around the core gaps. The heat spreaders are slightly larger than the dimension of the gap face to enhance the heat transfer into the potting compound. They extend about 2 mm way from the core edges, Figure 6-2.

Figure 6-1. Potted inductor on the heat sink cooled by a fan

Figure 6-2. Core with AlN heat spreaders

The thermal performance of the inductor with the AlN heat spreaders will be investigated through FE thermal analysis and experimental measurements. The results are then compared with the first inductor to identify the effectiveness of the heat spreaders.
6.2.3 Calibration of Thermal Sensors

The test inductors were fitted with thermal sensors before potting. The sensors used in the components were not of the same type. NTC-10k thermistors manufactured by MURATA [185] were used to measure the core temperatures and type K thermocouples were fitted in the coil to measure the winding temperature.

To calibrate the thermal sensors used in the tests, the sensors were used to measure the temperature of iced water, which ideally would be 0 °C. Several sensors of each type were tested and the average measurement results are shown in Table 6-1.

<table>
<thead>
<tr>
<th>Sensor types</th>
<th>Measured temperatures (°C)</th>
</tr>
</thead>
<tbody>
<tr>
<td>NTC-10k thermistors (MURATA)</td>
<td>2.0</td>
</tr>
<tr>
<td>Thermocouple type K</td>
<td>-0.2</td>
</tr>
</tbody>
</table>

There are discrepancies in the measured temperatures using the two types of thermal sensor. The measured temperatures using thermocouple sensors are closer to 0 °C, whilst the temperatures measured by the thermistors were around 2 °C higher. To further calibrate the sensors, it is important to verify whether they measure the same temperature rises. This was done by connecting the sensors to a data logger to register the changes in ambient temperature for at least 24 hours. The recorded data shows that the temperature rises measured by the two types of sensors are the same over time. The 2 °C discrepancy must therefore be compensated in the temperature measurement results.

6.2.4 Calibration of Heat Sink with Fan

The total power loss in the test inductors, $P_{t_{measured}}$, will be estimated from the measured steady-state temperature rise, $\Delta T_{HS}$, in the heat sink, (6-1). This requires that the thermal resistance of the heat sink and fan assembly, $R_{\theta HS}$, to be carefully calibrated, and this is
done by measuring the temperature rise of the heatsink with known power dissipation from a resistive load.

\[ P_{t,\text{measured}} = \frac{\Delta T_{HS}}{R_{\theta HS}} \]  

(6-1)

The potted inductor mounted on the heat sink and fan assembly has been shown in Figure 6-1. Figure 6-3 shows the top view of the heat sink for calibration where two 100 Ω aluminium-housed resistors are mounted on the heat sink. The two resistors together have approximately the same contact surface area as the potted inductor.

The heat sink has fins and the forced air tends to flow along the direction of the fins, potentially resulting in a non-uniform heat path, and the temperature on the surface of the heat sink is likely to be non-uniform. The thermal resistance of the heat sink must therefore be characterised at the same measurement point so that the total component loss can be accurately estimated. A type K thermocouple was fitted to the heat sink surface, next to the heat source and in the middle along the length of the heat sink to measure the temperature on the heat sink, as shown in Figure 6-3.
The two power resistors were connected in parallel and then connected to a DC power supply to characterise the heat sink and fan assembly at different power dissipation levels. The top of the heat sink and power resistors were covered with Superwool thermal insulation to ensure that all the heat was removed through the heat sink and the 12 V fan underneath the heatsink was operated at full power during the tests.

The thermal resistance was then calculated by dividing the measured steady-state temperature rise by the electrical power dissipation, \((6-2)\). The results are summarised in Table 6-2. At the measurement point, the average thermal resistance under different power dissipation levels is 0.120 °C/W.

\[
R_{\theta HS} = \frac{\Delta T_{HS}}{I^2(50)} \quad (6-2)
\]

Table 6-2. Characterisation of the thermal resistance of the heat sink assembly at the measurement point

<table>
<thead>
<tr>
<th>Power loss (W)</th>
<th>(R_{\theta HS}) (°C/W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>25</td>
<td>0.115</td>
</tr>
<tr>
<td>50</td>
<td>0.123</td>
</tr>
<tr>
<td>75</td>
<td>0.118</td>
</tr>
<tr>
<td>100</td>
<td>0.123</td>
</tr>
<tr>
<td>125</td>
<td>0.121</td>
</tr>
<tr>
<td>150</td>
<td>0.120</td>
</tr>
</tbody>
</table>

For the gap loss validation tests presented in this chapter, the heat sink temperatures were all taken from the same measurement point shown in Figure 6-3.

### 6.2.5 Operating Conditions

Table 6-3 summarises the operating conditions used in the experimental measurements in this chapter. The test inductors are subjected to a triangular current waveform at twice the
converter switching frequency, \( 2f_{sw} \). The DC input current, \( I_{DC} \), was kept the same at 160 A for all measurements. The converter input and output voltages, \( V_{in} \) and \( V_{o} \), were altered to provide five different ripple current levels, \( I_{pk-pk} \), at two converter switching duty cycles, 75 % and 60 %, corresponding to 50 % and 20 % duty ratios in the inductor current waveforms due to interleaving, respectively. The AC flux conditions were not exactly the same for the two duty ratio cases as the achievable peak flux levels were limited by the rated voltage of the converter.

Since the research is particularly concerned with the inductor performance, the term ‘duty ratio’ in later sections refers to the effective duty ratio of the inductor current waveform, as explained in Chapter 3. The two duty ratios represent the scenarios of symmetric and asymmetric waveforms, respectively.

Each test was run for at least 50 minutes to ensure that the component had reached the steady-state temperature. The thermal sensors used in the component were connected to a data logger to record the measured temperatures at a rate of one sample per second during the tests.

Table 6-3. Experimental operating conditions

<table>
<thead>
<tr>
<th>Conditions</th>
<th>( V_{in} ) (V)</th>
<th>( V_{o} ) (V)</th>
<th>( f_{sw} ) (kHz)</th>
<th>Switching duty cycle, ( D_{sw} )</th>
<th>( I_{DC} ) (A)</th>
<th>( I_{pk-pk} ) (A)</th>
<th>( B_{m} ) (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>22</td>
<td>88</td>
<td>30</td>
<td>0.75</td>
<td>160</td>
<td>41.5</td>
<td>0.05</td>
</tr>
<tr>
<td>2</td>
<td>31</td>
<td>124</td>
<td>30</td>
<td>0.75</td>
<td>160</td>
<td>57</td>
<td>0.075</td>
</tr>
<tr>
<td>3</td>
<td>44</td>
<td>176</td>
<td>30</td>
<td>0.75</td>
<td>160</td>
<td>84</td>
<td>0.11</td>
</tr>
<tr>
<td>4</td>
<td>50</td>
<td>200</td>
<td>30</td>
<td>0.75</td>
<td>160</td>
<td>102</td>
<td>0.13</td>
</tr>
<tr>
<td>5</td>
<td>70</td>
<td>280</td>
<td>30</td>
<td>0.75</td>
<td>160</td>
<td>129</td>
<td>0.16</td>
</tr>
<tr>
<td>6</td>
<td>56</td>
<td>140</td>
<td>30</td>
<td>0.60</td>
<td>160</td>
<td>46.5</td>
<td>0.05</td>
</tr>
<tr>
<td>7</td>
<td>94</td>
<td>228</td>
<td>30</td>
<td>0.60</td>
<td>160</td>
<td>69</td>
<td>0.09</td>
</tr>
<tr>
<td>8</td>
<td>114</td>
<td>280</td>
<td>30</td>
<td>0.60</td>
<td>160</td>
<td>85</td>
<td>0.11</td>
</tr>
<tr>
<td>9</td>
<td>117</td>
<td>291</td>
<td>30</td>
<td>0.60</td>
<td>160</td>
<td>90</td>
<td>0.12</td>
</tr>
<tr>
<td>10</td>
<td>120</td>
<td>300</td>
<td>30</td>
<td>0.60</td>
<td>160</td>
<td>94</td>
<td>0.125</td>
</tr>
</tbody>
</table>
6.3 Total Loss Measurements

This section describes the first stage to validate the gap loss model. The total inductor loss is measured and compared with the calculated losses. The loss mechanisms in the test inductor are considered to comprise three parts: copper losses, core loss and gap loss. The copper losses have DC and AC loss components. Two loss mechanisms occur in the core, the core loss and gap loss, where the core loss is mainly the hysteresis loss. The classical eddy current loss is assumed to be negligible due to the fine laminations.

In this section, the inductor was separated from the converter’s cooling system, allowing the total losses dissipated to be estimated by the steady-state temperature rise in the individual heat sink. The measured total losses in the component show good correlations with the calculations.

6.3.1 Loss Calculations

The operating conditions used in the experimental tests involve a DC bias plus triangular AC excitation at different duty cycles. The non-sinusoidal current waveform is taken into account in the loss calculations. The inductor loss components are calculated individually and then summed up to predict the total loss in the inductor, $P_{t,\text{calculated}}$, (6-3), where $P_{cu}$ is the copper loss, $P_{hy}$ is the hysteresis core loss and $P_{g}$ is the gap loss.

$$P_{t,\text{calculated}} = P_{cu} + P_{hy} + P_{g}$$  (6-3)

6.3.1.1 Copper losses, $P_{cu}$

The copper losses are calculated from published standard loss equations for foil wound windings [131]. The skin and proximity effects are taken into account. The effect of air gap fringing flux on the winding loss is not included. The calculations are summarised as follows.
The copper loss consists of DC and AC losses and the AC losses are calculated from all harmonics of the inductor current, as in (6-4).

\[
P_{\text{cu}} = P_{\text{cu,DC}} + P_{\text{cu,AC}} = R_{\text{DC}} I_{\text{DC}}^2 + \frac{1}{2} \sum_{n=1}^{\infty} R_{\text{AC,n}} I_n^2
\]  

(6-4)

where \( I_{\text{DC}} \) is the inductor average current, \( I_n \) is the amplitude of the \( n \)-th harmonic component of the inductor current. \( R_{\text{AC,n}} \) is the AC resistance at the \( n \)-th harmonic and \( R_{\text{DC}} \) is the DC coil resistance which is calculated from the copper resistivity, \( \rho_{\text{cu}} \), number of turns, \( N \), the mean length per turn, \( \text{MLT} \), the coil width, \( W_{\text{coil}} \), and the thickness, \( t_{\text{cu}} \), of the copper foil, as in (6-5).

\[
R_{\text{DC}} = \frac{\rho_{\text{cu}} N \text{MLT}}{W_{\text{coil}} t_{\text{cu}}}
\]  

(6-5)

The skin depth of copper at the \( n \)-th harmonic considering the frequency doubling effect in the inductor current ripple is

\[
\delta_{\text{cu}}(n) = \sqrt{\frac{\rho_{\text{cu}}}{n \pi \mu_0 (2f_{\text{sw}})}}
\]  

(6-6)

The stretch factor \( \eta \) is determined from the foil width and the total core window breath, \( \eta = W_{\text{coil}}/(2L_{\text{core}} + l_g/2) \). And the porosity factor at the \( n \)-th harmonic is calculated as

\[
\xi(n) = \frac{t_{\text{cu}} \sqrt{\eta}}{\delta_{\text{cu}}(n)}
\]  

(6-7)

Dowell functions [127] are used to calculate the AC resistance, \( R_{\text{AC,n}} \), where the first term in (6-8) accounts for the skin effect and the second term represents the proximity effect.

\[
\frac{R_{\text{AC,n}}}{R_{\text{DC}}} = \xi(n) \left[ \frac{\sinh 2\xi(n) + \sin 2\xi(n)}{\cosh 2\xi(n) - \cos 2\xi(n)} + \frac{2(N - 1)^2}{3} \frac{\sinh \xi(n) - \sin \xi(n)}{\cosh \xi(n) + \cos \xi(n)} \right]
\]  

(6-8)
6.3.1.2 Hysteresis loss, $P_{hy}$

The hysteresis loss was determined based on the measurements on an un-gapped inductor wound with the same core using the B-H loop method, as described in Chapter 3. The effects of DC bias and varying duty cycle are taken into account.

6.3.1.3 Gap loss, $P_g$

The calculation of gap loss has been described in Chapter 5. The empirical equation (5-9) derived from the simulations is used to predict the gap loss. The frequency used in the equation is twice the converter switching frequency due to interleaving. However, (5-9) is derived based on sinusoidal inductor current which is the default driving function in the steady-state electromagnetic solver in the Opera software.

Since the gap loss is an AC loss associated with eddy currents, it is therefore assumed that it is unaffected by the DC current. Three approaches have been used to approximate the effect of the triangular AC excitation. The total inductor losses calculated using these gap loss approximations are compared with the measurements in Section 6.3.2.

Approximation 1 considers the same flux swing in the core, that is a sinusoidal waveform with the same peak-to-peak value as the triangular one is used to estimate the gap loss. This approach assumes the gap loss is related to the peak flux therefore the effect of duty cycle is not included.

Approximation 2 uses a Fourier analysis. The estimated gap loss is the superposition of the loss calculated from the first three harmonic components of the original current waveform. The calculations for the two duty ratio conditions are slightly different. The 50 % duty ratio waveform only has odd harmonic components therefore the fundamental and third harmonics are considered, whilst the gap loss calculated at 20 % duty ratio includes the fundamental, second and third harmonic components.
Approximation 3 only uses the fundamental component of the triangular waveform to calculate the gap loss.

### 6.3.2 Experimental Results

The total inductor loss was estimated from the steady-state temperature rise measured at the measurement point shown in Figure 6-3 on the heatsink, using (6-1) in Section 6.2.4.

Figure 6-4, 6-5 and 6-6 compare the calculated and measured total inductor losses where the gap losses are calculated using the three approximations described in Section 6.3.1.3, respectively. The stacked columns on the left are the calculated loss components and the purple columns on the right are the total inductor losses estimated from the temperature rise in the heat sink. The percentage errors, $\varepsilon$, are marked on the plots for each test condition, where

$$\varepsilon = 100\% \cdot \frac{|P_{t,\text{calculated}} - P_{t,\text{measured}}|}{P_{t,\text{measured}}} \quad (6-9)$$

All three loss components increase with increasing AC excitation, resulting in a higher total loss at larger flux levels. For 50 % duty ratio conditions, the gap losses calculated from Approximation 2 (Figure 6-5a) and Approximation 3 (Figure 6-6a) are lower than those calculated from the equivalent sinusoidal waveform with the same flux change that is Approximation 1 shown in Figure 6-4a. Whilst for 20 % duty ratio conditions, the gap losses calculated from Approximation 1 (Figure 6-4b) and those calculated by using the first three harmonics, Approximation 2 (Figure 6-5b) are about the same, however, when only the fundamental component is used in the gap loss calculation (Figure 6-6b), the total losses calculated are lower.
Figure 6-4. Calculated and measured total inductor losses (gap loss calculated using Approximation 1)
CHAPTER 6  EXPERIMENTAL VALIDATION OF FE GAP LOSS MODELLING

Figure 6-5. Calculated and measured total inductor losses (gap loss calculated using Approximation 2)
Figure 6-6. Calculated and measured total inductor losses (gap loss calculated using Approximation 3)
The average percentage errors in the total loss estimations are compared in Table 6-4. The first two calculation approaches to approximate the gap loss with triangular waveforms result in smaller errors in the total inductor loss estimations, with the average errors less than 10% for Approximation 1 and around 5% for Approximation 2, while the errors are larger when using only the fundamental component to calculate the gap loss (Approximation 3), that is 11% at D = 0.5 and almost 20% at D = 0.2.

Whilst Approximation 1 and Approximation 2 lead to similar error levels in the total loss estimations, Approximation 2 offers better accuracy at high loss conditions and will be used to account for the non-sinusoidal waveform in the gap loss calculations in subsequent results.

<table>
<thead>
<tr>
<th>Gap loss calculation approach</th>
<th>Average ε at D = 0.5</th>
<th>Average ε at D = 0.2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Approximation 1</td>
<td>8.9 %</td>
<td>8.6 %</td>
</tr>
<tr>
<td>Approximation 2</td>
<td>4.7 %</td>
<td>5.0 %</td>
</tr>
<tr>
<td>Approximation 3</td>
<td>11.0 %</td>
<td>18.1 %</td>
</tr>
</tbody>
</table>

When comparing Condition 3 and Condition 8 with the same AC flux, the measured total loss is higher at D = 0.2 than the loss at D = 0.5. Both core loss and AC copper loss are increased with asymmetric waveforms. The effect of waveform shape on the gap loss is not yet well understood. However, the gap loss calculated from the harmonic components, Approximation 2, shows an increase in the loss at D = 0.2. The modified Steinmetz equation described in the Chapter 2, which is used to predict the traditional core loss under non-sinusoidal waveforms, relates the loss to the rate of change of the flux density, $\frac{dB}{dt}$. This could be another approach to estimate the gap loss under asymmetric waveforms.

The proximity of the coils to the core was not considered in the gap loss and copper loss calculations. The test inductor was designed to minimise the fringe-field-related loss in the winding, however according to the simulation, the winding was placed at a position at
which the perpendicular component of the flux is not entirely zero, but falls to less than 10\% of its peak value. There would be an additional copper loss component caused by fringing flux hitting the foil turns. Neglecting this effect may result in a slight overestimation in the gap loss and an underestimation in the copper loss. Effectively, part of the losses due to fringing flux could be transferring from the core to the coil. Whether the total amount of loss is affected remains unknown.

Overall, the calculated total losses agree with the experimental measurements at all test conditions with errors less than 5 W when using Approximation 2, where the gap loss contributes a significant part, almost 50\%, of the total loss in this design example. Furthermore, possible areas to improve the gap loss calculations have been identified. The effect of non-sinusoidal waveform and the spacing between coils and core on the gap loss calculation will form part of the future work described in Chapter 7.

6.4 Component Temperature Measurements

Whilst the measurements in Section 6.3 have confirmed the accuracy of the overall inductor loss predictions and by implication the total gap loss magnitude, this section attempts to verify the distribution of gap loss by measuring the inductor core and coil temperatures at different positions and comparing them with FEA predictions.

The calculated loss components using the gap loss Approximation 2 (Figure 6-5) are used in a 3D thermal FEA to estimate the steady-state temperature rise in the potted inductor. The electromagnetic FE model used to calculate the gap loss involves the mesh of the core and the background air region as described in Chapter 4, however, the thermal model must also include the air-gap spacers, coils, potting compound and the aluminium heat sink. The mesh in the thermal model is therefore no longer identical with the electromagnetic one, and the loss density distribution cannot be used directly in the thermal analysis. To overcome this problem several small volumes are determined from the electromagnetic loss density results over which the loss density may be assumed to be uniform. The details of the FE thermal model will be provided in Section 6.4.1.
The FE thermal analysis is validated by laboratory test results in Section 6.4.2. The results show that the cores have non-uniform temperature distributions. Significant temperature rises occur around the air gaps, indicating a localised loss distribution. Furthermore, the core hot spot temperature can be significantly reduced by using ceramic heat spreaders which may allow a smaller core to be used.

6.4.1 FE Thermal Modelling of the Potted Inductor

The model is built based on the geometry of the potted inductors designed by Dr Gerardo Calderon-lopez and it is solved in the Opera 3D software using the steady-state thermal solver, TEMPO. The model can be used to predict the steady-state temperature rise within the inductor and identify the hot spot. Furthermore, it can be used to evaluate and compare different thermal designs, and it therefore allows a more effective thermal management to be selected.

6.4.1.1 Potted structure

Figure 6-7 shows the FE thermal model of the potted component. The coils are wrapped around the core legs. One coil is made translucent to show the heat spreaders between the core halves. The top parts of the coils are outside the potting material. The light grey part is the aluminium case which is hollowed to a shell with two shoulders at the bottom of the case to support the core ends outside the winding. The electrical isolation is provided by a thin layer of Kapton HN tape between the core ends and the shoulders. There is approximately a 2 mm gap between the inductor and the aluminium can. The blue volumes represent the potting material that fills all the gaps between the inductor and the case, assuming the filling is even and there are no air pockets in the potting compound.

The bottom face of the aluminium case is assigned with a ‘fixed temperature’ boundary condition which is equal to the steady-state temperature of the heat sink. All other outer surfaces of the potted component are defined as perfect insulators. Therefore, all the heat will flow through the bottom of the aluminium case to the heatsink.
6.4.1.2 Core modelling

The thermal characteristics of the core are modelled in a similar homogenisation process as was used for the electrical and magnetic properties in Chapter 4. The volume orientations and the anisotropic thermal material properties are used to represent the laminated core structure. The thermal conductivities used are 10 W/mK in the direction along the laminations and 0.5 W/mK through laminations to take into consideration the poor heat transfer caused by the insulations between laminations [61, 160].

The hysteresis loss is assumed to be distributed uniformly in the entire core volume, whilst the gap loss is concentrated in the regions around the gaps. The 3D gap loss distribution result from the FE electromagnetic simulations cannot be imported directly to the thermal one due to the different meshes. Therefore, the core legs are divided into separate small blocks to account for the non-uniform gap loss distribution, the loss distribution being uniform in each block, as shown in Figure 6-8.

The core leg surface regions, where the gap loss is mainly distributed, are divided into four parts, as sketched in Figure 6-8a. They are: the core edges at the gap, Gout and Gin; the core side edges, SideOut and SideIn; then the rest of the core surface regions are divided into two parts: Out1 and Out2, In1 and In2. The depth of the surface areas, \( d \), is 1
mm which is slightly less than the effective skin depth of the homogenous core (1.2 mm). The middle of the core leg, *MidCore*, is treated as a single block where the loss density is minimal.

The dimensions of the surface regions as labelled in Figure 6-8b are determined from the gap loss density distributions on the core surfaces shown in Figure 5-14 in Chapter 5. The maximum loss density occurs at the edges of the core surface laminations. $L_{Gout}$, $L_{Gin}$ and $L_{Side}$ are chosen to be 1 mm, 1.2 mm and 1 mm, respectively. At these distances, the loss
densities drop to 30% of their peak values at the core gap and side edges. $L_{\text{SideOut}}$ and $L_{\text{SideIn}}$ are selected as 10 mm and 15 mm, respectively. About 70% of the losses along the core side edges are contained within these lengths. As illustrated in Chapter 5, varying the AC flux level or frequency does not appear to change the loss distribution on the core surfaces at fixed gap length.

The losses occurring in these small blocks are calculated individually from the electromagnetic analysis through integration. The allocations of the gap loss in these small blocks are listed in Table 6-5. The outer core surface regions contribute around 37% of the total gap loss, whilst around 58% of the gap loss occurs in the inner core surfaces. About 44% of the gap loss is concentrated in the gap edges, $G_{\text{in}}$ and $G_{\text{out}}$, and 15% of the loss in distributed along the side edges, $\text{Side}_{\text{Out}}$ and $\text{Side}_{\text{In}}$.

<table>
<thead>
<tr>
<th>Block name</th>
<th>Percentage of total gap loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>$G_{\text{out}}$</td>
<td>17.7%</td>
</tr>
<tr>
<td>$\text{Side}_{\text{Out}}$</td>
<td>6%</td>
</tr>
<tr>
<td>Out1</td>
<td>8%</td>
</tr>
<tr>
<td>Out2</td>
<td>4.8%</td>
</tr>
<tr>
<td>$G_{\text{in}}$</td>
<td>26.5%</td>
</tr>
<tr>
<td>$\text{Side}_{\text{In}}$</td>
<td>8.6%</td>
</tr>
<tr>
<td>In1</td>
<td>11.7%</td>
</tr>
<tr>
<td>In2</td>
<td>10.7%</td>
</tr>
<tr>
<td>MidCore</td>
<td>6%</td>
</tr>
</tbody>
</table>

6.4.1.3 Coil modelling

The two coils have equal numbers of turns so that the temperatures are assumed to be the same in both coils. Therefore, the electrical connection interlinking the coils is not included in the model, and the winding is modelled by two separate coils.
The foil wound coils with insulating tape between the turns have a similar layered structure to the laminated core. Therefore, they are also modelled by rotated orientations together with equivalent anisotropic thermal conductivities along the turns and through the turns. Figure 6-9 shows the coil model with eight individual blocks for each coil. The orientations of the blocks are set so that the foil layers are effectively bent at the corners of the coils. The dimensions are determined from the actual winding used in the test inductors. The calculated copper losses are assumed to be evenly distributed in the coils.

![Figure 6-9. Coil model for thermal FEA](image)

Each coil in the test inductor has three turns of 0.8 mm copper foil insulated with 0.13 mm Kapton tape, and the total thickness of the coils is around 4.5 mm. Even though the coils were wound as tightly as possible, there were air gaps between layers. Apart from the insulating tape, the air gaps between the turns are also responsible for the poor heat flow through the winding. Furthermore, when potted into the aluminium case, the air gaps between turns could be filled with epoxy resin. The potted part of the coil may therefore have a slightly better thermal transfer than the top part that is exposed to the air.

The effective thermal conductivity of the coil in the direction along the turns, $k_t$, and through the turns, $k_n$, are approximated from (6-10) and (6-11) [54]. The original equations from the reference [54] consider only the copper and insulation, whilst in this work additional terms representing the air gaps between copper turns are included.
\[ k_t = \frac{k_{cu}t_{cu\_coil} + k_{tape}t_{tape} + k_{gap}t_{gap}}{t_{coil}} \]  \hspace{1cm} (6-10)

\[ k_n = \frac{t_{coil}}{t_{cu\_coil}/k_{cu} + t_{tape}/k_{tape} + t_{gap}/k_{gap}} \]  \hspace{1cm} (6-11)

where \( t_{cu\_coil} \) is the total thickness of the copper foils used in a coil, \( t_{coil} \) is the thickness of the entire coil, \( t_{tape} \) and \( t_{gap} \) are the total thickness of the insulation tape and the gaps in the coil, respectively, \( k_{cu} \) and \( k_{tape} \) are the thermal conductivities of copper and Kapton tape, whilst \( k_{gap} \) is the thermal conductivity of air for the top parts outside the potting compound or the thermal conductivity of the potting compound for the potted regions of the coils.

### 6.4.1.4 Summary of material thermal properties

The thermal conductivities of the model components used in the FE thermal analysis are listed in Table 6-6.

<table>
<thead>
<tr>
<th>Components</th>
<th>Thermal conductivity (W/mK)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminium case</td>
<td>236</td>
</tr>
<tr>
<td>Epoxy resin potting compound</td>
<td>1.8</td>
</tr>
<tr>
<td>Coils</td>
<td></td>
</tr>
<tr>
<td>along the turns</td>
<td>218.5</td>
</tr>
<tr>
<td>through the turns (potted parts)</td>
<td>1.05</td>
</tr>
<tr>
<td>through the turns (exposed parts)</td>
<td>0.08</td>
</tr>
<tr>
<td>Kapton HN tape</td>
<td>0.12</td>
</tr>
<tr>
<td>Finemet core</td>
<td></td>
</tr>
<tr>
<td>along the laminations</td>
<td>10</td>
</tr>
<tr>
<td>through the laminations</td>
<td>0.5</td>
</tr>
<tr>
<td>Aluminium nitride (AlN) heat spreaders</td>
<td>90</td>
</tr>
<tr>
<td>Plastic air gap spacers</td>
<td>0.35</td>
</tr>
</tbody>
</table>
6.4.2 Experimental Results

The calculated loss data provided in Figure 6-5 were used in the FE thermal simulations to estimate the steady-state temperature rise within the inductor structure. The simulation results are compared with the experimental measurements taken at different sensor points to verify the temperature distributions in the component.

Figure 6-10 shows the locations of the thermal sensors. CoG1 and CoG2 were used to measure the core temperatures at the corners around the gap. CG1 and CG2 were placed in the middle of the core leg along the laminations to measure the temperatures at the gap and underneath the end of the coil, respectively. TG1 and TG2 were fitted along the centre line of the outermost lamination at the core edge and half way down the length of the core leg, respectively. T_{coil} was placed in the mid-turn of the coil to monitor the temperature in the centre of the winding.

Figure 6-10. Locations of thermal sensors (The other coil not shown)

The FEA solutions at different sensor locations use the average values in a square patch of 2*2 mm\(^2\) on the core surfaces at the corresponding positions, which is about the size of the contact surfaces of the thermal sensors.

The FEA and measurement results are provided in the following sections under several typical test conditions and more results are provided in Appendix C.
6.4.2.1 Maximum AC flux condition

The measurement and FEA simulation results under Condition 5, as in Table 6-3 in Section 6.2.5, are compared in this section, which represents the worst case scenario with maximum current ripple in the inductor at the converter’s rated operating condition.

Figure 6-11 shows the transient temperature measurement results in the inductors using plastic air gap spacers (without AlN heat spreaders). The temperatures reached their steady-state after about 30 minutes of operation.

![Temperature Measurement Results](image)

**Figure 6-11. Measured temperatures in the inductor using plastic spacers, Condition 5, $T_{HS} = 38 \, ^\circ C$**

The measured hot spot temperature is 108 °C which is a 70 °C temperature rise above the heat sink. The regions near the gaps, CoG1, CoG2, CG1 and TG1, experience the highest temperatures. By comparing CG1 and CG2, the core temperature reduces by 17 °C moving away from the gap along the middle of the core leg to the end of the core leg. On the outer core surface, the temperature drops by 14 °C from TG1 to TG2 moving halfway down the core leg from the gap. The high temperatures around the gap regions confirm the non-uniform gap loss distribution as predicted.
Figure 6-12 is the test results using the inductor with AlN heat spreaders. The hot spot temperature is reduced by 20 °C by inserting the heat spreaders between the core halves. A 10 - 15 °C reduction is achieved at the other sensor points around the gaps.

![Temperature vs. Time](image)

Figure 6-12. Measured temperatures in the inductor with AlN heat spreaders, Condition 5, $T_{HS} = 38$ °C

Figure 6-13 shows the predicted steady state temperature distributions from the FE thermal analysis and Figure 6-14 shows the temperature profiles in the cores only. The effectiveness of the AlN heat spreaders is clearly visible. A more uniform core temperature is observed in the inductor with AlN heat spreaders, Figure 6-14b. The hot spots shown in the simulations are slightly inside the core rather than on the core surfaces. This is because the potting compound has effectively improved the heat transfer from the core surfaces.

The top core surface shown as the front surface in the Figure 6-14 is hotter than the surface at the back which is at the bottom of the potted structure closer to the aluminium case. The direction of the main heat flow in the inductor without heat spreaders is from the core edges at the gap towards the far end of the core legs along the laminations and
then down through the bottom core surface to the shoulders of the aluminium case. For the inductor with AlN heat spreaders, additional heat paths are created to spread the heat around the gaps to the potting compound and across the cross section of the core.

![Thermal FEA results showing the potted inductors, Condition 5, T_{HS} = 38 °C](image)

Figure 6-13. Thermal FEA results showing the potted inductors, Condition 5, T_{HS} = 38 °C

![Thermal FEA results showing the cores only, Condition 5, T_{HS} = 38 °C](image)

Figure 6-14. Thermal FEA results showing the cores only, Condition 5, T_{HS} = 38 °C

The top edges of the aluminium case extend by 1 mm above the core. Assuming the epoxy resin potting compound is uniform, the thickness of potting material on top of the core is also 1 mm. Simulations have been performed to slightly increase the height of the
aluminium case, effectively increasing the potting thickness over the core top area. A 2 ºC reduction is achieved in the top core surface by increasing the amount of potting material above the core by 1mm. The reduction is not significant. The amount of potting material is not a key factor in the optimisation of thermal performance. Using more potting material will not significantly reduce the core temperatures but will increase the overall component size and weight. Therefore, the clearance between the inductor and the case (currently 2 mm) is not worth increasing. A better solution to reduce the temperatures on the core top surface would be to implement a lid to the aluminium case, which can be considered in future designs.

Table 6-7 compares the FEA predicted and measured steady-state temperature rises above the heat sink at different sensor positions (shown in Figure 6-10) under Condition 5, and the results are plotted in Figure 6-15 to show the temperature rise distributions.

Table 6-7. Summary of FEA and measured steady-state temperature rises above the heat sink, Condition 5, $T_{HS} = 38$ ºC

<table>
<thead>
<tr>
<th>Sensor points</th>
<th>Inductor without heat spreaders</th>
<th>Inductor with AlN heat spreaders</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>FEA (ºC)</td>
<td>Measurement (ºC)</td>
</tr>
<tr>
<td>CoG1</td>
<td>68</td>
<td>57</td>
</tr>
<tr>
<td>CoG2</td>
<td>58</td>
<td>70</td>
</tr>
<tr>
<td>CG1</td>
<td>54</td>
<td>59</td>
</tr>
<tr>
<td>CG2</td>
<td>41</td>
<td>42</td>
</tr>
<tr>
<td>TG1</td>
<td>56</td>
<td>52</td>
</tr>
<tr>
<td>TG2</td>
<td>41</td>
<td>38</td>
</tr>
<tr>
<td>$T_{coil}$</td>
<td>34</td>
<td>42</td>
</tr>
</tbody>
</table>
The simulated core hot spot is CoG1 which is the internal core corner at the gap, whilst the measured hot spot is CoG2, the outer core corner. One possible explanation is that the sensors at CoG1 and CoG2 were mistakenly swapped during assembly. Because intuitively, the heat would flow from inside of the component towards the heat sink at the exterior. A higher temperature is expected around the middle of the potted component. Another possible reason is that the higher temperature in the exterior could be related to the proximity between winding and core. As explained earlier, the effects of coil-to-core distance on the gap and copper loss calculations were not considered in this research. The coils of the test inductors were placed closer to the inner core surface and slightly away from the outer core surface which could possibly result in more fringing flux around the outer core regions and therefore a larger gap loss could possibly occur at the core exterior. However in the inductor with AlN heat spreaders, the measured temperature at CoG1 is higher than CoG2, and the same patterns are seen at all other test conditions. The heat spreaders will only affect the heat flow but cannot alter the loss distributions. Therefore, it is believed that the loss occurring in the inner core regions were higher than those in the outer regions. The ‘swapped’ sensor position seems to be a more reasonable explanation.
for the discrepancy between the simulation and measurement results. As the component has already been potted, the positions of CoG1 and CoG2 cannot be verified.

If the measured temperatures at CoG1 and CoG2 are swapped over, the FEA and measurement results will closely match with each other. There is a 2 ºC difference between the measured and predicted core hot spot temperatures, which is less than 3% error in the temperature rise with respect to the heat sink. A 20 ºC error in the temperature rise prediction around the gap area was seen when using the gap loss allocation described in the previous work [160]. This research has significantly improved the accuracy of temperature predictions by using the proposed gap loss calculations and a thermal FEA using the predicted gap loss distribution.

Another relatively large error seen in the inductor without heat spreaders is between the simulated and measured coil temperatures at the middle of the winding, $T_{coil}$. The predicted temperature rise is 8 ºC lower than the measurement. One possible reason to explain the error is that the sensor was placed on the mid-layer of the coil which is copper, whilst in the model the coil is represented by an equivalent bulk of conductor and the temperature is taken from the middle of an averaged continuum. Furthermore, the tape used to provide insulation in the coils has a low thermal conductivity of 0.12 W/mK. The tape at the coil surfaces is responsible for the poor heat exchange between coils and the potting material. However in the model, the averaged thermal conductivity only accounts for the low heat transfer between turns whilst the poor thermal contact at the boundary of the coils is not included, which may result in an underestimation in the coil temperature. In addition, the coil is modelled with a constant cross sectional area throughout the entire volume over which the loss density is assumed to be uniform. In practice, the actual coil has a thicker dimension at the shorter ends where the sensor is placed (top coil regions outside the potting compound). A photograph of a foil winding is provided in Appendix B. This is due to extra copper pieces soldered to the turns to interlink two coils and connect the coils to the converter circuit. Therefore, there may be more losses in the top parts of the coils. Also the solder, screw caps, washers etc. may affect the heat dissipation in the coils.
During the test, a thermal camera was used to monitor the temperatures in the entire converter. A temperature of 85 °C was observed at the winding connection point. As copper is also a heat conductor, the winding is not entirely thermal-isolated. The heating in the connections will affect the coil temperature measurements, but the temperature at the connection points is not easy to predict. This measurement identifies another possible aspect to optimise the thermal performance of the converter and the components involved by improving the thermal contacts at the connection points. For example, additional small heat sinks can be used to create a heat path for the winding connections to the converter’s cold plate.

6.4.2.2 Varied frequency conditions

The inductor without heat spreader was connected to a 25 kHz DC-DC converter to validate the gap loss model at a different frequency. The converter was stepping from 57 V to 225 V with an input current of 160 A. The inductor was subjected to a symmetric triangular waveform with a peak flux of 0.16 T, the same as Condition 5 in Table 6-3, but at a lower frequency, 50 kHz.

The total loss estimated from the measured temperature rise in the heat sink is 80 W under the 50 kHz condition, which is about 33 W lower than the total loss measured at 60 kHz. The calculated total loss is 85 W, distributed as: $P_{cu} = 41.5$ W, $P_{hy} = 4.5$ W and $P_g = 40$ W, which has an error of about a 6.3 % comparing with the measurement.

The measured and predicted steady-state temperature rises at different sensor points are presented in Figure 6-16, comparing the temperature rise at the two frequencies. At 50 kHz, the measured temperature rise at the hot spot is about 13 °C lower than that under the 60 kHz condition due to the lower losses. A good match is achieved between the FEA and measurement results except for the swapping of CoG1 and CoG2. The errors between the predicted and measured hot spot temperature rises are within 2 °C or 4 % for both conditions.
6.4.2.3 50 % and 20 % duty ratio conditions

Figure 6-17 shows the temperature rise prediction and measurement results under Condition 3 and Condition 8 (as in Table 6-3), which have the same peak flux density, but at 50 % and 20 % duty ratio, respectively.

When changing the duty ratio of the current waveform from 50 % to 20 %, the measured temperature rises around the air gap, such as CoG1, CoG2, CG1 and TG1, were increased by around 6 - 8 °C, while at CG2 and TG2 which are away from the gap the measured temperatures rises were increased by around 4 °C.

The FEA predicted temperature rise for the hot spot is 2 °C below the measurement at 50 % duty ratio. An error of 7 °C (around 15 %) is seen at 20 % duty ratio and the error is lower away from the gap. This could be due to an underestimation of gap loss under asymmetric waveforms. At 20 % duty ratio, the gap loss is possibly larger than expected or more localised, which will cause a higher temperature rise around the gap regions.
Figure 6-17. FEA and measured steady-state temperature rises at 50 % and 20 % duty ratio, Condition 3 and Condition 8, $T_{HS} = 36$ °C

### 6.5 Conclusion

The objective of this chapter has been to validate the FE gap loss calculations. The total inductor loss measurements have verified the predicted magnitude of the gap loss in Section 6.3. The measurements of the component’s internal temperatures have confirmed the non-uniform loss distribution in Section 6.4.

By using the calculated losses, the FE thermal model is able to estimate the steady state core temperatures in the potted component with acceptable accuracy at various operating conditions. A closely matching temperature distribution in the core has been achieved between the FE simulations and experimental measurements, which corresponds to the predicted gap loss distribution. An improved coil model is needed to accurately predict the temperatures in the winding.

The effectiveness of high-thermal-conductivity heat spreaders in the gaps to significantly reduce the core hot spot temperature has been demonstrated, which may allow a smaller
core to be used and therefore the size and weight of the inductor could be reduced. Other possible solutions to further improve the thermal performance of the component have also been identified, such as a lid could be added to the aluminium case to reduce the temperatures on the top core surfaces, also the winding temperature may be reduced by improving the electrical connections and creating additional heat paths for the connections to the heat sink.
CHAPTER 7

CONCLUSIONS AND FUTURE WORK

7.1 Introduction

This chapter summarise the research work reported in this thesis. The key contributions are highlighted and suggestions for future work are described.

7.2 Summary of Research

The research has focused on the losses in high-power, high-frequency nanocrystalline inductor cores, including the traditional core loss and the gap loss associated with air gap fringing flux.

7.2.1 Nanocrystalline Core Loss Characterisation

The core losses in four commercial nanocrystalline sample C-cores using Finemet and Vitroperm 500F materials have been experimentally characterised in Chapter 3 under typical operating conditions that are encountered in high power converter applications. The classical two-winding measurement method is used and the core losses are assessed by measuring the B-H loops of the cores. A dual interleaved boost converter with interphase transformer (IPT) is used to subject the test components to triangular flux waveforms with variable duty ratio together with a range of DC bias conditions. The measured full magnetisation curves of the cores allow the core losses to be characterised as a function of DC flux density. Higher losses are observed under asymmetric waveform conditions and the increasing DC bias increases the core loss further, especially at high DC flux density levels. The Vitroperm cores were seen to be superior to the Finemet core in terms of losses, especially with high levels of DC flux.
7.2.2 Finite Element Gap Loss Modelling in Nanocrystalline Cores

A 3D FE electromagnetic model has been developed in Chapter 4 to examine the effect of the gap loss in a Finemet C-core under high frequency operation. The laminated core with gaps is modelled by several individual homogeneous blocks with equivalent anisotropic electrical and magnetic properties.

The model is developed in the Opera software. The volume orientation technique along with the use of Biot-Savart conductors allows the inductor to be represented in full, and therefore the actual flux distribution in the core with a winding can be examined. The model size is then reduced by defining the model symmetry with appropriate boundary conditions. The modelling method can be applied to other core shapes and winding designs.

The shapes and sizes of the 3D finite elements and their impact on the mesh generation process are considered along with the implications for the solution of the model. The model is then modified to improve the mesh quality. A faster and more accurate solution has been achieved by using a mosaic mesh where the mesh size is determined from the equivalent skin depth of the homogeneous core model. A mesh size of 0.3 mm was used and the effective skin depth in the homogeneous core model was 1.2 mm at 60 kHz.

7.2.3 Gap Loss Analysis and Prediction

The FE model developed in Chapter 4 has been used with sinusoidal winding current to examine the gap loss and its distribution in Chapter 5.

The flux distribution in the core is studied first, especially the distribution of the perpendicular flux component that generates the eddy currents in the lamination planes. The results show a non-uniform flux distribution in the laminated core. The flux density is higher towards the core surfaces and lower in the middle of the core leg throughout the
magnetic path, whilst the perpendicular flux is concentrated at the core surfaces around the edges at the gap.

Then the distribution of the eddy currents induced by the perpendicular flux and the associated losses are examined. The results show that the maximum loss density occurs towards the centre of the core gap edges, and the side edges of the surface core laminations strips also experience a high loss density. Altering the amplitude and frequency of the excitation did not appear to change the loss distribution on the core surfaces, whilst the gap loss is more concentrated towards the gap edges for smaller gap lengths. The loss distribution results from the simulations show that the gap loss is likely to be increased with increasing core strip width, but is relatively independent of the length and height of the core legs.

The total gap loss is very significant in the design example and it is highly concentrated, especially around the gap edges, leading to a large heat flow. Without thermal management, the localised loss may lead to an excessive temperature rise and thermal failure of the core.

A sensitivity study has been undertaken using the simulation model to investigate the effects of the inductor design parameters, such as air gap length, frequency, flux density and core strip width, on the total gap loss. Assuming the gap loss is a power function of these parameters, a model-based expression has been derived from the sensitivity study to predict the magnitude of the gap loss in a Finemet inductor at high frequency.

### 7.2.4 Experimental Validation

The FE gap loss model has been validated by experimental measurements in Chapter 6 using two Finemet inductors designed by Dr Gerardo Calderon-lopez for the input inductor for a dual interleaved DC-DC converter with IPT. The inductors are encapsulated in aluminium cans using a thermally conducting potting compound.
The predicted magnitude of the gap loss has been validated by comparing the measured and calculated total losses in the test inductor, where the total calculated losses are determined from the measured core loss in Chapter 3, the gap loss predicted from the equation proposed in Chapter 5 and the winding loss calculated from the published equations from the literature. Three approaches have been used to approximate the gap loss under non-uniform excitation waveforms, and by using the first three harmonic components of the triangular waveform, the calculated total losses show good agreement with the measurements with errors less than 5 W. The gap loss contributes a significant part of the total inductor loss, almost 50%. Accurate prediction of the total loss therefore provides high confidence in the gap loss calculation.

The predicted distribution of the gap loss has been confirmed by measuring the core temperatures within the test inductor at different positions and comparing them with the predictions using a 3D thermal FEA. The FE thermal model has been developed using a non-uniform gap loss distribution approximated from the electromagnetic modelling results reported in Chapter 5. The measurements show a non-uniform temperature distribution in the core that largely corresponds to the predicted gap loss distribution. The inductor core’s hot spot is located at the core corner near the gap and has a 70 °C temperature rise relative to the heat sink. The temperatures are lower away from the gap. The effectiveness of high-thermal-conductivity heat spreaders between core halves to reduce the hot spot temperature has also been demonstrated. The simulations have an excellent match with the measured steady-state temperatures under various test conditions, with errors of less than 5 % in the hot spot temperature predictions. The possible reasons for the errors are discussed.

### 7.3 Contributions

Nanocrystalline tape wound cores offer low core losses and high saturation flux density, making them attractive for compact inductor designs. However, to optimise the component design accurate loss data is required.
Several empirical models have been published in the literature to predict the core loss for non-sinusoidal waveforms however they are not always easy and practical to use. Furthermore, none of the existing core loss equations combine the effects of non-sinusoidal waveforms with DC bias conditions. The core loss measurement results presented in this thesis provide a more complete picture of the core loss variations with both DC and AC excitation. In addition, the effect of DC bias on the core loss is examined as a function of DC flux density which is more practical to use than DC magnetic field strength. The measurement results show that the core losses are higher under asymmetric waveform conditions and will increase with increasing DC bias, especially at high DC flux density levels. The core losses can be either well below or higher than the datasheet values as the duty ratio and the DC bias are varied. The measured loss data can be used as a basis for the design of high power inductors.

The gap loss associated with air gap fringing flux is not particularly well understood in the literature and it is often neglected in inductor designs. The established equation to estimate gap losses can be dated back to the 1940s and was derived for power line frequency applications. In this research, a 3D electromagnetic finite element (FE) model has been developed to examine the distribution of flux as well as the induced eddy currents and associated gap losses in the core at high frequency. The results show that the gap loss can contribute a major part of the total loss, almost 50% in the design example. Furthermore, the gap loss distribution is not uniform in the core. The loss is highly localised around the core edges at the gaps, which may result in local over heating in these regions. The loss distribution results provide a sound basis for inductor designs and thermal management, either to reduce gap loss or to extract heat from the hot spot regions caused by the gap loss. An empirical equation is proposed based on the modelling results to provide a design oriented estimation of the gap loss in nanocrystalline inductor cores at high frequencies. Whilst being similar to the original gap loss equation proposed by Lee for power line frequencies, this proposed equation for finely laminated high-frequency cores includes non-unity exponent terms for gap width and frequency.
A 3D thermal FE model has also been developed based on the loss distribution results to predict the steady state temperature rises in foil-wound Finemet inductors encapsulated in aluminium cases. The thermal FE model can also be used to examine different thermal management methods to optimise the thermal performance of the component, which may potentially allow a further size and weight reduction.

The calculations of the gap loss magnitude and its distribution, together with the thermal predictions, have been validated experimentally, showing good correlations between predictions and experimental measurements.

7.4 Future Work

7.4.1 Core Loss Characterisation

Multi-cut cores that produce smaller distributed gaps have been proposed in the literature to reduce the high frequency losses in the windings. However, the cutting process increases the core losses. This thesis has measured the core losses in four single-cut sample cores. Characterising the core losses in multi-cut cores will further complete the loss information required for a high power inductor design.

7.4.2 Electromagnetic Gap Loss Modelling

As mentioned above, using multiple gaps can reduce the fringe-field-related winding losses, and therefore potentially reduce the gap loss in the core and distribute the loss more evenly. The FE models presented in this thesis can be extended to assess both the electromagnetic and thermal performance of the multi-gap inductor designs.

The inductor hot spot temperature is largely dependent on the gap loss and its distribution. The experimental measurements closely match the temperature predictions with 50 % duty ratio conditions. However, the predictions under 20 % duty ratio conditions are less
accurate. Further studies on the effect of non-sinusoidal waveform on the gap loss are required.

The FE gap loss model in this thesis examines the gap loss in the inductor and the winding losses caused by the fringing flux are not included, whilst in the literature the winding losses are studied without considering the gap loss effect. Replacing the ideal Biot-Savart conductors in the FE model described in Chapter 4 with representative copper volumes will potentially allow the fringe-field-related losses in both the core and winding to be analysed in a single model, also allowing the impact of the coil-to-core spacing on the gap loss and winding loss to be examined. However, copper has a very small skin depth at high frequency that will create more challenges in the finite element mesh control.

7.4.3 Thermal Finite Element Modelling

The thermal FEA has been shown to be accurate for steady state temperature predictions in the core, but relatively large errors are seen in the winding temperature predictions. Further refinements in the coil modelling are required, including a more accurate approximation of the loss distribution in the coils and a better characterisation of the thermal interface between the coil and potting material.

In addition, extending the thermal analysis to transient conditions would allow the component’s performance limits to be assessed during overload conditions.

7.4.4 Inductor Design Optimisation

The highly localised gap loss distribution reported in this thesis has identified several areas of future research for optimised inductor design either to reduce the gap loss or to develop new thermal management methods to reduce the hot spot temperature.
Alternative aspect ratio core designs could be considered with reduced core strip width to reduce the eddy current paths at the gap. The increased core volume and core loss would need to be traded against the reduced gap loss.

Additional heat spreading structures could be considered around the core in the region of the gap, for example high-thermal-conductivity strips could be placed along the core lamination surfaces to provide additional thermal paths from the gap edges and core sides to the potting material.

Finally, the high temperature observed at the winding connection points during the laboratory tests identifies the need to improve the winding connections in future designs.
REFERENCES


REFERENCES


[73] R. Garcia, M. Escobar-Mejia, K. George, and J. C. Balda, "Loss comparison of selected core magnetic materials operating at medium and high frequencies and


REFERENCES


REFERENCES


REFERENCES


REFERENCES


APPENDIX A. CORE LOSS MEASUREMENT EQUIPMENT

Table A-1: Measurement equipment

<table>
<thead>
<tr>
<th>Digital oscilloscope</th>
<th>LeCroy WaveRunner 6 Zi</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage probe</td>
<td>LeCroy PP008-1</td>
</tr>
<tr>
<td>Current Probe</td>
<td>LeCroy CP500</td>
</tr>
<tr>
<td>Impedance Analyzer</td>
<td>Agilent 4294A</td>
</tr>
</tbody>
</table>
APPENDIX B. PHOTOGRAPHS OF THE TEST INDUCTOR

Figure B-1. Picture of Potted inductor (winding not interlinked)

Figure B-2. Picture of foil winding with similar dimensions to the one used in the potted structure
APPENDIX C. THERMAL FEA AND TEMPERATURE MEASUREMENT RESULTS

C.1 60 kHz Conditions

This section provides additional FEA temperature predictions and experimental measurement results that are not included in Section 6.4.2. The test conditions are listed in Table 6-3 in Section 6.2.5 and the loss calculations are provided in Figure 6-5 in Section 6.3.2.

![Temperature Rise Chart]

Figure C-1. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 1, D = 0.5, \( T_{HS} = 29.5 \, ^\circ C \)
Figure C-2. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 2, D = 0.5, $T_{HS} = 31.5$ °C

Figure C-3. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 4, D = 0.5, $T_{HS} = 37$ °C
Figure C-4. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 6, D = 0.2, $T_{HS} = 33 \, ^\circ C$

Figure C-5. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 7, D = 0.2, $T_{HS} = 34 \, ^\circ C$
Figure C-6. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 9, D = 0.2, $T_{HS} = 37$ °C

Figure C-7. FEA and measured steady-state temperature rises in the inductor without heat spreaders, Condition 10, D = 0.2, $T_{HS} = 37$ °C
C.2 50 kHz Conditions

This section provides additional FEA temperature predictions and experimental measurement results under 50 kHz conditions that are not included in Section 6.4.2.

Figure C-8. FEA and measured steady-state temperature rises in the inductor without heat spreaders, $I_{DC} = 160$ A, $B_m = 0.05$ T, $f = 50$ kHz, $D = 0.5$, $P_{hy} = 1$ W, $P_{cu} = 16$ W, $P_g = 5.5$ W, $T_{HS} = 28$ °C
Figure C-9. FEA and measured steady-state temperature rises in the inductor without heat spreaders, $I_{DC} = 160$ A, $B_m = 0.9$ T, $f = 50$ kHz, $D = 0.5$, $P_{hy} = 1.4$ W, $P_{cu} = 21$ W, $P_g = 10.3$ W, $T_{HS} = 30.5$ ºC

Figure C-10. FEA and measured steady-state temperature rises in the inductor without heat spreaders, $I_{DC} = 160$ A, $B_m = 0.12$ T, $f = 50$ kHz, $D = 0.5$, $P_{hy} = 2.5$ W, $P_{cu} = 16$ W, $P_g = 23$ W, $T_{HS} = 33$ ºC
Figure C-11. FEA and measured steady-state temperature rises in the inductor without heat spreaders, $I_{DC} = 160 \, \text{A}$, $B_m = 0.21 \, \text{T}$, $f = 50 \, \text{kHz}$, $D = 0.5$, $P_{hy} = 7.5 \, \text{W}$, $P_{cu} = 64 \, \text{W}$, $P_g = 72 \, \text{W}$, $T_{HS} = 43 \, ^\circ\text{C}$