Methods to Quantify and Reduce Rotor Losses in a Solid Rotor Yoke Permanent Magnet Machine

by
Dominic Wills

Dissertation presented in fulfilment of the requirements for the degree of Doctor of Philosophy in Engineering at Stellenbosch University

Promotor: Professor Maarten Kamper
Department of Electrical and Electronic Engineering

March 2010
DECLARATION

By submitting this dissertation electronically, I declare that the entirety of the work contained therein is my own, original work, that I am the owner of the copyright thereof (unless to the extent explicitly otherwise stated) and that I have not previously in its entirety or in part submitted it for obtaining any qualification.

March 2010
Abstract

Certain types of electric machines are particularly susceptible to the proliferation of eddy currents flowing within the solid conducting regions in the rotor. Single-layer, non-overlapping windings within uneven open slots are some stator properties that can produce damaging, asynchronous magnetic field harmonics which manifest in the rotor as eddy currents. The ohmic losses caused by these eddy currents are a source of inefficiency and can cause a marked increase in the temperature of the rotor. This temperature rise can be dangerous for the magnets, which have to be kept within temperature limits to avoid partial or full demagnetization.

The research work presented here is concerned with reducing the effect of eddy currents in the rotor magnets and solid rotor yoke of an electric machine. The work presents analytical methods to calculate the magnetic fields, eddy currents and solid loss in an electric machine due to current in the winding and due to the interaction of the permeance variation in the stator with the magnets in the rotor. A method is also suggested where the analytical theory can be used with a magnetostatic finite element solution to produce a transient solid loss result. The research work also investigates a method for optimal segmentation in both level and penetration, and provides some design suggestions.

The work presents the method of partial magnet segmentation, which is a technique whereby thin incisions are made into the magnet material from one or both sides. Another method of partial rotor segmentation is also presented where the incisions are made into a portion of the magnet-facing solid yoke. These methods attempt to interrupt the flow of eddy currents and increase the resistance 'seen' by the eddy currents, while also keeping construction difficulty and cost to a minimum. The methods are verified using finite element calculations which are compared to measured results.

The result is that partial magnet segmentation is a very useful, effective and practical method of segmenting magnets. The loss reduction profile can be similar to that of traditional full segmentation. The method of partial rotor segmentation also shows a large reduction in rotor power loss. With implementation of these methods on a test machine, one can expect an efficiency increase of more than 4%.
Opsomming

Sekere tipes van elektriese masjiene is veral sensitief vir die vloei van werwelstrome in solied geleidende gebiede in die rotor. Enkellaag, nie-oorvleuelende wikkelings in oneweredige oop gleewe is enkele stator eienskappe wat skadelike, asinchrone magneetveld harmonieke tot gevolg kan hê, wat as werwelstrome in die rotor manifesteer. Die ohmise verliese wat deur hierdie werwelstrome teweeg gebring word is 'n bron van ondoeltreffendheid en kan lei tot 'n merkbare toename in die temperatuur van die rotor. Hierdie temperatuur styging hou gevaar in vir die magnetes en moet binne temperatuur limiete gehou word om gedeeltlike of self volle demagnetisering te vermy.

Die navorsing vervat in hierdie document is gemoeid met die vermindering van die effek van werwelstrome in die rotor magnete en in die soliede rotor juk van 'n elektriese masjien. Die werk bied analitiese metodes aan vir die berekening van die magneetvelde, werwelstrome en soliede verliese in 'n elektriese masjien as gevolg van strome in die wikkelings en die interaksie van die permeansie variasie van die stator met die magnetes in die rotor. 'n Metode word ook voorgestel waar die analitiese teorie saam met 'n magnetostatiese eindige element oplossing gebruik word om 'n resultaat vir die oorgang soliede verliese te verkry. Die navorsingswerk onderzoek ook 'n metode vir die optimale segmentering in beide vlak sowel as penetrasie, en verskaf sekere ontwerp voorstelle.

Die werk bied die metode aan van gedeeltelijke magneet segmentering, wat 'n tegniek is waarvolgens dun insnydings gemaak word aan een of beide kante van die magneet materiaal. Nog 'n metode van gedeeltelike rotor segmentering word beskou waar die insnydings in in 'n gedeelde aan die magneetkant van die soliede rotor juk gemaak word. Hierdie metodes poog om die vloei van werwelstrome te onderbreek en die weerstand soos "gesien" deur die werwelstrome te verhoog, terwyl konstruksie kompleksiteit en koste tot 'n minimum beperk word. Die metodes word bevestig deur eindige element berekeninge wat met gemete resultate vergelyk word.

Die gevolg is dat gedeeltelijke magneet segmentering 'n baie nuttige, doeltreffende en praktiese metode van die segmentering van magnetes is. Die verliesverminderingsprofiel van gedeeltelike segmentering kan soortgelyk wees aan dit van tradisionele volle segmentering. Die metode van gedeeltelike rotor segmentering toon ook 'n groot afname in rotor drywingsverlies. Met die implementering van hierdie metodes op 'n toetsmasjien, kan 'n mens 'n verhoging in benuttingsgraad verwag van meer as 4 %.
Acknowledgements

The following people/organisations have been instrumental in allowing me to complete this work by contributing guidance, advice and emotional support.

Professor Kamper, my supervisor for his guidance and support, open mindedness, flexibility and extremely keen interest.

Johannes Potgieter, who worked on the design, construction and testing of the test machine.

Adriaan Lombaard, who completed the mechanical design of the machine.

Pietro Petzer for supervising the technical work with mechanical and electrical construction.

Andre, for his work with machine construction especially with the rotor magnets.

To the SA National Antarctic Programme, for the funding and necessary infrastructure to make the travel opportunities and this research possible.

To Stellenbosch University and the postgraduate office for all the bursaries that I have received.

From Loher, GmBH, Ruhstorf, Germany

Jacques Germishuizen and Andreas Joeckel for affording me the opportunity to work at Loher, GmBH, where I gained invaluable experience in using FE software.

To my friends, family for all their support, especially

Peta, for her love and unconditional support and encouragement despite some long travel commitments,

My Mother, for always being there for me as a pillar of strength and comfort.
Contents

1 Introduction .............................................................................................................................................................. 1
   1.1 Background to Study........................................................................................................................................... 1
       1.1.1 Challenges facing Machine designers for WG Applications ....................................................... 2
       1.1.2 Permanent Magnet in Wind Generator Design................................................................................ 3
   1.2 Problem Statement........................................................................................................................................... 3
   1.3 Scientific Approach to the Research.............................................................................................................. 4
       1.3.1 Analytical Calculations ...................................................................................................................... 5
       1.3.2 Finite Element Analysis..................................................................................................................... 5
       1.3.3 Measured Results ................................................................................................................................ 6
   1.4 Scope and Limitations........................................................................................................................................ 6
   1.5 Research contribution........................................................................................................................................ 6

2 Literature Review ................................................................................................................................................... 8
   2.1 Introduction ........................................................................................................................................................... 8
   2.2 Magnet and Rotor Loss publications............................................................................................................ 8

3 Rotor Loss Calculation Methods .................................................................................................................... 13
   3.1 Analytical Model Assumptions ..................................................................................................................... 13
       3.1.1 End Effects are Ignored.................................................................................................................. 13
       3.1.2 Hysteresis Losses Ignored ............................................................................................................ 14
   3.2 Analytical Prediction of Magnetic Fields .................................................................................................. 14
       3.2.1 Magnetic Field due to Stator Winding...................................................................................... 14
       3.2.2 Rotor Magnetic Field Calculation ............................................................................................... 16
       3.2.3 Prediction of Magnetic Vector Potential in Airgap, Magnets, Yoke.............................. 18
   3.3 Eddy Current Calculation ................................................................................................................................ 21
       3.3.1 Calculating Eddy Currents from Magnetic Vector Potential..................................................... 21
       3.3.2 Eddy Currents in Solid Conductors ........................................................................................... 21
       3.3.3 Eddy Currents in Segmented Magnet Conductors ........................................................................ 21
       3.3.4 Eddy Currents in Segmented Rotor Yoke Conductors .............................................................. 22
   3.4 Loss Calculation .................................................................................................................................................. 23
       3.4.1 Loss Calculation from Current Density........................................................................................ 23
       3.4.2 Loss Calculation using the Skin Depth........................................................................................ 23
   3.5 Finite Element Modelling ................................................................................................................................ 24
       3.5.1 Model ..................................................................................................................................................... 24
       3.5.2 Boundary Conditions ....................................................................................................................... 24
       3.5.3 Excitations ........................................................................................................................................... 24
4.2.1 Resistivity Model ................................................................. 27
4.2.2 Current Density Subtraction Model ........................................ 29
4.2.3 Segmentation Model Comparison ........................................... 30
4.3 Full Magnet Segmentation .......................................................... 30
4.3.1 Description .............................................................................. 30
4.3.2 Types of Segmentation ............................................................. 31
4.3.3 Manufacturing .......................................................................... 31
4.3.4 Disadvantages of Full Magnet Segmentation ......................... 32
4.4 Partial Magnet Segmentation ....................................................... 32
4.4.1 Description .............................................................................. 32
4.4.2 Model ................................................................................... 34
4.4.3 Manufacturing .......................................................................... 34
4.4.4 Limitations ............................................................................. 35
4.5 Partial Rotor Yoke Segmentation .................................................. 35
4.5.1 Description and Purpose .......................................................... 35
4.5.2 Model ................................................................................... 36
4.5.3 Manufacturing .......................................................................... 36
4.5.4 Limitations ............................................................................. 37
4.6 Optimal Magnet Segmentation ..................................................... 37
4.6.1 Static Segmentation Model ....................................................... 38
4.6.2 Segmentation Model Including Time ......................................... 41
4.6.3 Optimal Segmentation in Design .............................................. 42
5 Implementation of Analytical Rotor Loss Calculation in FEA ............ 44
5.1 Introduction ............................................................................ 44
5.2 Model .................................................................................... 44
6 Results and Comparison ............................................................... 48
6.1 Test Machine ........................................................................... 48
6.2 Analytical and FEM Calculation Comparison ............................. 50
6.2.1 Harmonic Speed Calculation .................................................. 51
# List of Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>magnetic vector potential</td>
<td>V.s.m(^{-1})</td>
</tr>
<tr>
<td>B</td>
<td>flux density</td>
<td>T</td>
</tr>
<tr>
<td>H</td>
<td>magnetic field strength</td>
<td>A.m(^{-1})</td>
</tr>
<tr>
<td>(\omega_s)</td>
<td>Stator Synchronous Frequency</td>
<td>radians</td>
</tr>
<tr>
<td>t</td>
<td>Time</td>
<td>s</td>
</tr>
<tr>
<td>(\mu)</td>
<td>Harmonic number</td>
<td></td>
</tr>
<tr>
<td>(N_s)</td>
<td>Number of Slots</td>
<td></td>
</tr>
<tr>
<td>(C_{sl})</td>
<td>Conductors per slot</td>
<td></td>
</tr>
<tr>
<td>(l_m)</td>
<td>Machine length</td>
<td>metres</td>
</tr>
<tr>
<td>(d_{si})</td>
<td>Stator inner diameter</td>
<td>metres</td>
</tr>
<tr>
<td>(n_s)</td>
<td>Rotor speed</td>
<td>rad.s(^{-1})</td>
</tr>
<tr>
<td>(N_p)</td>
<td>Number of poles</td>
<td></td>
</tr>
<tr>
<td>(\kappa)</td>
<td>Conductivity</td>
<td>S.m(^{-1})</td>
</tr>
<tr>
<td>(R_s)</td>
<td>Stator radius adjacent to airgap</td>
<td>metres</td>
</tr>
<tr>
<td>(g')</td>
<td>Effective airgap</td>
<td>metres</td>
</tr>
<tr>
<td>(h_y)</td>
<td>Yoke height</td>
<td>metres</td>
</tr>
<tr>
<td>(\mu_r)</td>
<td>Relative permeability</td>
<td></td>
</tr>
<tr>
<td>(J)</td>
<td>Current density</td>
<td>A.m(^{-2})</td>
</tr>
<tr>
<td>(T_s)</td>
<td>Segment width</td>
<td>radians</td>
</tr>
<tr>
<td>(P_s)</td>
<td>Magnet segmentation penetration</td>
<td>%</td>
</tr>
<tr>
<td>(P_y)</td>
<td>Yoke Segmentation penetration</td>
<td>%</td>
</tr>
<tr>
<td>(b_{sl})</td>
<td>Slot width</td>
<td>radians</td>
</tr>
<tr>
<td>(r_p)</td>
<td>Pole Pitch</td>
<td>radians</td>
</tr>
<tr>
<td>(r_{mc})</td>
<td>Magnet centre radius</td>
<td>metres</td>
</tr>
<tr>
<td>(h_{m})</td>
<td>Magnet height</td>
<td>metres</td>
</tr>
<tr>
<td>(l_s)</td>
<td>Segment length</td>
<td>metres</td>
</tr>
<tr>
<td>(N_{ss})</td>
<td>Number of Segments</td>
<td></td>
</tr>
<tr>
<td>(N_{rs})</td>
<td>Relative Degree of Segmentation</td>
<td></td>
</tr>
<tr>
<td>(\mu_0)</td>
<td>Permeability of free space</td>
<td>H.m(^{-1})</td>
</tr>
<tr>
<td>(b_0)</td>
<td>Stator slot opening width</td>
<td>radians</td>
</tr>
<tr>
<td>(K_w)</td>
<td>Winding factor</td>
<td></td>
</tr>
<tr>
<td>(K_c)</td>
<td>Carter Factor</td>
<td></td>
</tr>
</tbody>
</table>
1 Introduction

The focus of this research is on reducing the eddy current loss within solid conducting regions in the rotor of an electric machine. The work in this study has a range of applications in different machine topologies. However in this particular work, the focus and background of the study was within the field of PM machine design for wind generator applications. WG design will remain the focus; however, the reader should note that the research can be applied to many other applications too.

1.1 Background to Study

Wind Generation is the fastest growing form of renewable energy. Since 2003, global capacity has risen from 40,000 MW to 94,000 MW at the end of 2007, growing at an average annual rate of approximately 25% [1]. Ambitious renewable energy targets aim to maintain this trend, especially in Europe where leaders have penned a resolution aspiring to have renewable energy comprising 20% of Europe’s power generation by 2020.

Wind technology has reached a level of maturity in some areas such as blade and tower design where further efficiency improvements are being made through a process of evolution rather than revolution. However, in the area of drivetrain and power electronic design, the field is divided between

- direct drive, low speed PM generator with a fully rated converter,
- gearbox coupled, high speed PM generator with fully rated converter,
- high speed doubly fed induction generator coupled with a multi-stage gearbox and partially rated converter.

Each of these designs have their advantages and disadvantages, however we are likely to see a dominant technology emerge in the future.

One factor that benefits the low speed direct drive technology is the production of permanent magnets, in particular Neodymium Iron Boron (Nd-Fe-B) rare earth magnets, which in the sintered form, are the strongest type of permanent magnet available. Due to the high saturation magnetization of around 1.6 Tesla, Nd-Fe-B magnets have a largest energy yield which can be 11-20 times higher than ordinary ferrite magnets and twice that of Samarium Cobalt rare earth magnets [2].

Permanent Magnets are used in machines to create a static field in the rotor, which interacts with a rotating stator winding field to produce torque. PM machines have an advantage over other types of synchronous machines in that they replace the rotor windings, thus eliminating the need for slip rings and reducing associated copper losses. In direct drive application for
wind generators, which require slow rotational speeds, high pole numbers can also be used in machines with increased diameter, rendering the gearbox redundant.

It is important to briefly underline the challenges facing wind generator machine designers before being able to highlight the advantages of using permanent magnet direct drive machines.

1.1.1 Challenges facing Machine designers for WG Applications

The wind generator manufacturing industry is very competitive and coupled with slim power generation margins, generator efficiency is a make-or-break design consideration. Higher machine efficiency can be used to mitigate any additional cost as more power is produced for a given wind speed. Some challenges include:

- **Low blade speeds dictate low angular velocities** of generator shaft. This forces machine designers to either use a low-speed generator, a gearbox-coupled high-speed generator or an intermediate arrangement with a medium-speed generator used in conjunction with a reduced stage gearbox.

- Maximum instantaneous **cogging torque** values have to be lower than the minimum stationery torque provided by the blades at the cut in speed so as not to interfere with the start up performance.

- As wind generators provide an increasing proportion of national grid capacity, the grid codes that the wind turbines need to adhere to are becoming stricter. These codes dictate that wind generators behave like power stations, implementing stern power quality and fault ride through requirements.

- Wind generator technology has to **compete with other forms of renewable energy** and conventional power generation, which places cost as an important consideration when selecting machine materials and manufacturing techniques.

- **Reliability** is a consideration which affects overall cost and efficiency as downtime for maintenance causes loss in generator revenue in addition to the cost of repairs.

- **Acoustic emissions** play an important role in the approval of environmental impact assessments, so drivetrain designers need to keep noise within limits.

- **Machine size and weight** are important considerations for the overall turbine design as they affect the dynamic and static performance of the wind turbine system. They also affect overall cost as heavier machines and drive trains require more sturdy foundations, towers and nacelle.

The current industry standard with a few exceptions is to use a doubly fed induction machines mated with a multi stage gearbox in the drivetrain.
1.1.2 Permanent Magnet in Wind Generator Design

Considering the main design criteria listed above, one can now look at how a permanent magnet machine could contribute and improve on the current technology.

- A low speed, high pole, direct drive machine can eradicate the gearbox completely. This saves on some of the initial cost and also the additional maintenance cost of replacing the gearbox which can occur 1-2 times during its lifetime. Efficiency can also potentially be improved as the mechanical losses associated with the gearbox are eliminated.
- A permanent magnet machine requires the use of a fully rated converter in order to regulate the power produced by the machine for grid connection. This adapts very easily to the various grid codes as there is a ‘soft’ connection between the turbine and the grid. This also applied to gearbox-coupled, high speed PM machines.

These benefits of opting for permanent magnet direct drive machines are causing an increasing number of manufacturers to investigate the technology. Direct drive technology would be a step in a new direction as it requires:

- Redesign of the nacelle to incorporate the large diameter, high pole number generator.
- Changing to a fully rated converter.
- New stator and rotor design in a research field that is relatively new.

The permanent magnet rotor is one under developed area, as it is dissimilar to other machine topology rotor types. It contains an iron yoke, into which the permanent magnets are either embedded, or they are surface mounted. The performance of the rotor is crucial as not only does it play a strong role in overall efficiency, but also in magnet temperature regulation.

1.2 Problem Statement

The role of the rotor yoke is to provide a return flux path for the magnetic field caused by the magnets. As the magnets are constant in strength and are in a fixed position, the rotor yoke field is essentially static. Therefore points in the yoke operate in one quadrant of the BH curve only and have to tolerate less flux variations than in the stator whose points operate throughout the 4 quadrants in the BH curve. This lower flux variation in the yoke raises the question of whether it is necessary to use laminated steel in the rotor yoke. An alternative option which simplifies manufacture and reduces build cost is to roll the rotor yoke from solid steel. But even with the reduced flux pulsations in the rotor yoke, there still exists enough flux variation to induce significant eddy currents due to the high conductivity of the steel. These currents reduce efficiency and the ohmic power loss produces unwanted heat. To summarise:

“A solid steel rotor yoke exhibits great susceptibility to eddy current induction and power loss due to its proximity to harmful asynchronous field harmonics and high conductivity.

This loss constitutes a source of heat generation and inefficiency. Design needs to be
performed in order to minimize this loss while keeping cost and manufacturability a priority.”

The rare earth magnet Nd-Fe-B has a conductivity approximately 1/10\textsuperscript{th} that of steel. Despite this considerable difference in conductivity, the magnets are still highly susceptible to conducting eddy currents due to their proximity to the stator field winding harmonics and their low permeability also increases eddy current skin depth. With enough heat produced by these eddy currents, magnets can exceed their operating temperature range causing irreversible demagnetization. The power loss due to heat production also decreases overall machine efficiency. Design and construction methods to combat this effect need to be developed, which leads to the next problem statement:

“The magnets exist very close to the source of asynchronous field harmonics and combined with their moderate conductivity, they provide a medium in which eddy currents can flow. Design needs to be performed to neutralize the influence of these harmful effects in order to increase efficiency and keep magnet temperatures optimal.

This design must also keep manufacturability a priority.”

These questions can only be answered if one has an accurate and reliable method of calculating the eddy current loss in the magnets and solid rotor yoke. In some finite element packages, there exists a function than can calculate eddy current losses and therefore solid loss. However, this requires time-consuming, simulations over many fine time instants in order to gain accurate results. This eddy calculation process coupled with an iterative design optimization algorithm would too lengthy for practical purposes. A faster technique that uses a single time instant FEM calculation predicting rotor loss is required.

“Design and especially optimization in FEA demands an incorporated, single time step, rotor loss calculation for no load and full load conditions that is fast, accurate and reliable”

1.3 Scientific Approach to the Research

The approach of this study is to present alternative methods to improve machine efficiency by minimizing eddy current losses in the rotor magnets and yoke. The specific design used in this study is a concentrated coil, single layer, open slot, surface mounted magnet, 15 kW machine for wind generator application. This 40-pole, 48-slot machine constitutes a design that is regarded as being conducive to inducing large eddy currents in the rotor for the following reasons:

- The single-layer, non-overlapping, winding layout contains a magnetic field with a large, deep-penetrating, asynchronous MMF sub-harmonic.
• The stator permeance variation due to open slots cause larger flux pulsations in the rotor and magnets than if the slots were closed or semi-closed.
• The wound stator teeth are different in size and shape to the non-wound stator teeth.
• The surface mounted rotor magnets receive no protection from any magnetic field transients emitted from the stator.

These factors make this machine design one that is vulnerable to many contributors to rotor and magnet losses, and therefore a good test subject in this research. This work does not attempt to redesign the stator or any other part of the machine in order to minimize rotor losses. Instead, the work focuses on a machine design that is simple, inexpensive and manufactureable. Given this constraint, the research seeks methods to mitigate magnet loss by redesigning the rotor with manufactureability a priority.

There are three main methods that are used to complete calculations for this research and each method is briefly outlined below.

### 1.3.1 Analytical Calculations

In order to explore the origins of the eddy currents in the rotor yoke and magnets, it is necessary to investigate the PM machine field theory. The theory provides a mathematical model explaining observed behaviour, providing understanding and knowledge on which specific design ideas can be based. The model uses the Laplacian Differential Equation, which is used to calculate the 2-D magnetic vector potential fields due to the magnets and stator winding. Implementation of the machine model on a software platform creates a fast, useful calculation tool for design and comparison. The model uses machine information such as rotor speed, current, machine dimensions and number of slots and poles to produce a very fast solution. Another advantage is that design changes can be quickly implemented and evaluated enabling use in design optimization algorithms.

### 1.3.2 Finite Element Analysis

The most commonly used design tool for electromagnetic machine design is Finite Element Analysis (or FEA). This software is the most versatile, robust and most accurate method of predicting magnetic fields within a machine. The software works by breaking down a machine structure into programmable ‘elements’. For each time step a magnetic field value is calculated at the ‘nodes’ which are located at the corners of each element. The software then collates all the data from the individual nodes and returns values such as flux linkage, torque and voltage. The software can also be used to calculate other values that require auxiliary calculations such as flux density, current density, core loss, and ohmic loss.
In this study, a commercial package called Ansoft Maxwell 2D® is used. One benefit of this software is the capability to calculate eddy current loss in the rotor, which is not available in some other types of FE software. Due to the machine being broken up into elements, it is possible to extract information from specific points and monitor them as a function of time, or extract magnetic flux or current density values from complete lines from a radius in a machine. This ability helps especially when comparing the software to results obtained analytically.

1.3.3 Measured Results
The final method used to compute results is from measurement of a test machine. The test bench has an induction drive coupled to the machine via a gearbox and a torque and speed sensor. The goal to determine the benefits of various rotor configurations was achieved by building three different rotors, each differing by one design modification. These rotors are all tested with the same stator, shaft and bearing for consistency. The rotors are all tested at no-load and full load.

1.4 Scope and Limitations
This study was undertaken to investigate whether for a given PM machine, one could improve the way the solid yoke and magnets are designed and constructed in order to keep eddy current loss minimal. This study does not investigate stator design to reduce the presence of harmonics, as this has already been done in [4]. The research specifically targets novel ways in which to construct the rotor yoke and magnets so that for any given machine, one can implement these methods to reduce rotor losses.

Limitations on the research were mainly that when dealing with the built machines, there was a performance difference between segmented and non-segmented magnets, due to the magnet material lost in the segmentation process. Also, the construction of a partially segmented yoke was not possible in time for full testing and measurement.

1.5 Research contribution
When attempting to calculate the losses in the magnets and rotor yoke of a machine, research in [3,4] use a model of the stator winding that defines a linear current density sheet on the surface of the stator to model the stator current induced MMF harmonics. The Laplacian equation governing the behaviour of the ensuing magnetic fields is then used to extrapolate these fields to form a 2-D magnetic field solution in the rotor. This method is completely correct, however it fails to incorporate the rotor losses due to the permeance variation caused by stator slotting which can be significant especially for machines with open slots. This leads to the first contribution:
• Analytical calculation of no-load magnet and rotor yoke loss due to stator slotting only and implementation of this method into a FE program using data from one time step. This method splits the magnetic field calculated in the airgap due to the magnets into its Fourier components. Using the slot and pole information, one can ignore the stationary field harmonics due to the permanent magnets. The magnitude and relative speeds of the asynchronous harmonics are then calculated and losses are computed.

• The methods of single-sided and double-sided partial magnet segmentation are presented. These methods are aimed at improving the ease of magnet manufacture while still deriving the performance benefits of normal segmentation. Full analytical, FEA and measured results are presented.

• The method of partial rotor yoke segmentation is presented. This method aims to interrupt the path of eddy current flow in the rotor by incorporating partial thin radial cuts into the solid rotor yoke steel, however the solid steel disk remains is one piece to keep construction difficulties to a minimum.

• Investigation into optimal magnet segmentation is also considered, where a property of relative magnet segmentation pitch is introduced. This concept can be used in design where prior knowledge of destructive eddy current harmonic orders can be used to aid decisions on the degree of segmentation.
2 Literature Review

2.1 Introduction

Neodymium Iron Boron Magnets were invented by a consortium of companies and organisations in 1982 in response to the rising costs of Samarium Cobalt rare earth magnets. However, it was only in the 1990’s, when falling material costs made the incorporation of Nd-Fe-B magnets feasible in machine design. The major difference between Nd-Fe-B magnets and its predecessor rare earth types lies in their strength and conductivity. These changes have introduced new options and design challenges in machine design. This review takes a look at some of the work on magnet and rotor losses that has been published in the field of PM machines over the past 12 years.

2.2 Magnet and Rotor Loss publications

The work of Polinder et. al on magnet loss points out that prior to this research, designers often ignored the solid loss in magnets as ferrite magnets were used predominantly which have relatively high resistivity values. In addition to this, magnets were modelled as a cylinder, which fails to consider that isolated magnets cannot conduct between themselves. This prompted the study in [3] which generated a model of magnet loss due to the time harmonics in the stator current waveform. This work assumed that the magnetic field in the airgap was one-dimensional, it ignored reaction fields in the magnets, it assumed constant flux density over the magnet breadth, ignored end effects, and also ignored the losses due to the space harmonics caused by the stator winding current, it also ignored the effects of stator slotting. With all these assumptions the author admits that ‘the loss is not calculated very accurately, but a reliable approximation is obtained’.

Polinder continued the work in [4] on magnet loss with the same mathematical model; however this work was modified to include the effect of segmentation. The results show a reduction in magnet loss proportional to the square of the level of segmentation, while the author also notes that magnet loss increases as a square of speed.

Kawase et. al work on a Finite Element Model in [5] to quantify the effect of segmentation in permanent magnet machines calculating the magnetic vector potential with a more efficient finite element technique called the ‘double node technique’. They conclude that the eddy current density is larger at the surface of the magnet than at the back. They also conclude that increasing magnet segmentation strongly diminishes the magnet loss.

In the earliest of their works on magnet losses, Atallah et. al in [6] show how magnet losses can be significant in non-overlapping machines as the torque is produced by a higher order
harmonic. In this topology, lower order harmonics rotating asynchronously to the rotor can give rise to significant eddy currents. The work goes on to quantify the effectiveness of magnet segmentation.

Toda et. al continue the work from [6], and performed a comparison between a modular 24 slot/22 pole machine with a conventional 36 slot/24 pole machine using previously published analytical techniques [8]. This work concludes that:

- Significant eddy current loss exists in both machines
- Segmentation is very effective in reducing magnet losses
- The analytical model used was less accurate in the slotted machine due to 'the fact that eddy-current loss due to the stator slot openings is more significant (than expected), and is not considered in the analytical model'.

Zhu et. al furthered their research in [17] to expand their analytical model of calculating magnet loss. This polar coordinate-based model is based on two dimensional field calculations in the airgap and magnet region and takes into account the eddy current reaction field. The model also allows for calculation of eddy currents in a retaining sleeve as well as a variety of winding configurations, but neglects the effect of slotting. The authors apply the model to a brushless DC traction machine. Conclusions include:

- Better agreement is achieved with the improved model between measured and calculated results at high speed due to inclusion of the eddy current reaction field.
- The machine considered in the experiment had partially closed slots, and losses due to slotting were found to be negligible.
- The losses due to time harmonics in the armature reaction field were found to be more significant than the effect of the space harmonics due to the winding configuration.

The work in [6, 8] is continued in [9] where the authors extend their magnet loss model to include single layer fractional slot machines where wound teeth differ in width to unwound teeth. Chief assumptions include ignoring slotting and the effect of reaction fields in the magnets. The conclusions of this study include that single layer windings produce double the magnet losses than double layer windings and that unequal tooth widths result in the highest eddy current loss.

Yasuaki Aoyama et. al perform one of the first published physical experiments on Nd-Fe-B magnets [10]. A magnet was thermally insulated and placed in the resultant field of a solenoid coil driven by a function generator. The magnet is then subjected to flux pulsations of varying magnitude and frequency. The temperature is recorded over time and a thermal model is used to verify the results of the same test performed in a finite element simulation. They had good agreement between both models and found that segmentation effectively suppressed the power losses in the magnet.
Polinder et al. begins to look at losses in the solid back iron yoke of the rotor [11]. Analytical methods are used to calculate losses in machines with single and double layer windings, while investigating the losses generated from various slot/pole combinations. This work also includes an interesting experiment with a machine that has a rotor and no magnets. The stator is pulsed with frequencies of varying amplitudes resulting in a pulsating field in the yoke. The resulting current and power is measured which includes the eddy current losses in the rotor yoke. The experimental results agree strongly with the measured results with the major conclusion being that **single layer windings produce excessive losses in the yoke** and should be avoided.

On the subject of single layer windings, in the next publication [12], the authors focus on magnet losses in modular (single layer wound) machines. The advantages being to do with easier construction techniques using preformed coils and electrical isolation which improves fault tolerance. As in [9], the authors highlight that torque production occurs through interaction of high order winding harmonics, with the low order, asynchronous harmonics causing large magnet losses. The work goes onto **describe a new efficient finite element method of calculating the effect of axial segmentation** without resorting to standard 3-D computational methods. Comparison is also made with circumferential segmentation and favourable results are obtained with both methods.

Markovic et al. [13] work on a solution for a slotless PM machine with similar analytical methods as used in [6,9,12]. Due to the ironless nature of the stator, the machine experiences no losses due to permeance variation between the tooth and slots. They use a double fourier series that includes space and time harmonics especially those time harmonics anticipated from commutation events in the drive circuitry. As this machine has very little inductance due to its air core, these time harmonics have the potential to be especially destructive. **The thrust of the paper is in developing the model and good agreement is found with the results of FEA.**

Nuscheler [16] proposes a paper that details most of the steps in developing the analytical model to calculate eddy current losses in the magnets and the rotor yoke. **The method takes the reaction field into account so therefore ‘not restricted to cases where the skin depth is large compared to the geometrical dimensions of the magnets and the rotor yoke’**. This work covers all of the calculation detail step-by-step, which makes it a very useful educational tool. The work analytically computes the effect of segmentation in the magnets not excluding the reaction field. The research considers two different machines with identical rotors, but with stators of 9 and 12 slots each. It compares the two machines and looks at the effect of enlarging the airgap and the magnet heights. Conclusions include:

- That the solid rotor yoke reduces the losses in the magnets compared to a laminated yoke due to the reaction field in the rotor yoke causing a damping effect on the flux pulsations.
• Increases in magnet height and airgap were effective in decreasing the magnet and yoke losses.

• Operation of concentrated coil machines at high speed causes excessive rotor losses which can only be kept within tolerable limits with segmented magnets and a laminated rotor yoke.

• Careful attention should be paid to the choice of stator for a given rotor configuration.

Sergeant et al. look at different methods of magnet segmentation in their work in [15]. They look at the magnet losses caused by the time harmonics in square voltage PWM waveforms using a 3-D finite element method. Their conclusions include that axial and circumferential segmentation is effective in reducing magnet loss and that segmenting magnets is only useful when using a laminated yoke due to the shielding effect of the eddy currents when the yoke is solid.

Bianchi et al. focus on investigating rotor losses in a particular group of PM machines [18]. These machines have a fractional slot winding and are known to be particularly susceptible to high magnet losses due to the large number of eddy-current inducing space harmonics in the airgap. The analytical model used includes a current sheet to describe the winding configuration by defining the individual harmonic magnitudes, wavelengths and speeds. The model also defines an airgap and uniform conducting rotor region, without any magnets or yoke specified. Although the model is very accurate, it is not meant to be used to calculate actual rotor losses, but rather to be able to rapidly compute the effect of various pole-slot combinations on rotor losses. A relationship is developed between power loss and variables such as stator current, wavelength, specific wavelength, rotor speed, conductivity, permeability and machine dimensions. This work concludes that:

• Rotor Losses increase with specific wavelength and harmonic order. However these two variables are inversely proportional to each other.

• Losses increase as a square of the winding factor, so sub-harmonics which are more destructive, should have as low winding factors as possible.

• The airgap acts as a low pass filter.

That completes the list of research publications that was investigated before and during this study. To summarise the work completed on the topic of eddy current rotor loss, the research base covers:

• Mathematical problem formulation and identification of the primary eddy current sources, and the major factors that affect their magnitudes,

• The impact of machine dimensions on eddy current loss,

• Investigations into the effect of segmentation on reducing eddy currents,
• Investigation into the effect of eddy currents and their reaction fields on design,
• Methods to redesign the stator and winding in such a way as to minimize the eddy currents induced in the rotor.

The work in this research begins with the premise that there is a set stator and winding design and given this constraint, attempts to determine:

• what can be done to reduce the magnitude of the eddy currents induced in the magnets and solid rotor yoke of a machine,
• how can one ensure loss reduction while maintaining cost effectiveness and ease of manufacture.

In the ensuing chapters of this research, the answers to these questions will be attempted.
Chapter 3  
Rotor Loss Calculation Methods

In order to describe the eddy currents induced in the rotor and calculate magnet losses, one must isolate the origins of the magnetic field in a machine and define a model that accurately describes it. This chapter steps through the calculations that are used to compute the analytical results.

3.1 Analytical Model Assumptions

Before continuing with the description of the analytical model, it should be noted that a few assumptions are made for simplification. These assumptions are thought to have minimal impact on the accuracy of the results, however if the methods are used in other applications, the assumptions made might not still hold true. For this reason, the reader should be prudent in understanding these assumptions and their limitations.

3.1.1 End Effects are Ignored

Real life eddy currents flowing within solid conducting regions have no spatial restrictions and can potentially flow in any direction within the (x,y,z) plane. Three dimensional modelling is computation intensive compared with 2-D or 1-D modelling and for this reason, various assumptions are made in order to bring down computation time and complexity. The analytical model seeks simplification by defining areas where insignificant activity is observed and then discounting these areas to simplify calculations.

In the case of eddy currents flowing within the magnets and solid rotor yoke, the currents are assumed to only flow in the positive and negative z-direction. The reason for this is due to the geometry of the magnet in question which has a much larger length in the z direction than the width, which is in the x direction. Conducting regions have especially narrow x-dimensions after implementation of radial segmentation which further amplifies the relative contribution of eddy current in the z-direction over the x-direction. For this reason, magnet loss due only to the effects of eddy currents flowing in the z-direction is considered to comprise a vast majority of the total magnet loss, hence the assumption is to omit the effect of eddy currents flowing in the x-direction. In the finite element analysis, the same assumption is made, and a 3-D package is needed in order to quantify the effects of the eddy currents flowing in the x-direction. It should be noted that for complete accuracy, the 3-D model should be used and this limitation is noted.
3.1.2 Hysteresis Losses Ignored

The rotor yoke in a permanent magnet machine is designed to provide a return path for flux on the non-airgap side of the magnets. The magnet strength doesn't vary significantly, so one can expect the flux density at each point in the rotor yoke to remain within one quadrant.

![Hysteresis Curve](image)

**Fig. 3-1:** A typical hysteresis curve showing the flux density locus as it travels through the four quadrants of the B-H plane.

Hysteresis losses are typically generated in regions where the material experiences flux density changes that cause the material to operate in more than one quadrant in the B-H curve. In steel, the locus of the flux density plot travels a different path depending whether the magnetic field strength is increasing or decreasing, and this is especially significant if the magnetic field strength changes sign. In situations where the magnetic field strength does not change sign, i.e., it stays within one quadrant, one expects the hysteresis losses to be negligible as the forward and return flux density paths are approximately equal. It should also be noted that the machine referred to in this work was designed so that the flux density in the rotor yoke exceeded the saturation flux density ($B_s$) which pushes the material into extreme saturation. From Fig. 3-1, one can see that at this operating point, even with significant magnetic field strength pulsations, there is no hysteresis loop and therefore hysteresis losses are negligible.

3.2 Analytical Prediction of Magnetic Fields

3.2.1 Magnetic Field due to Stator Winding

For simplicity, it is convenient to unroll a conventional PMSM machine to create a 2-D, linear machine model in Cartesian coordinates as shown in Fig 3-2. As this work is focused on the rotor, the co-ordinate axes are fixed to the moving rotor reference frame. The airgap, magnets and rotor yoke each have differing conductivities and permeabilities, so for this reason they are treated as separate regions each with their own boundaries. The y dimension is defined as
being zero at the surface of the stator and runs perpendicular to the boundary interfaces while the x dimension runs in the direction the boundary interfaces.

![Fig. 3-2: PMSM model showing the coordinate system, the different regions of interest and their dimensions.](image)

The stator winding consists of phase-shifted, AC, current-carrying coils, each phase spatially distributed along the stator surface. When summed, these phase harmonics produce a ‘working’ magnetic field harmonic which interacts with the fundamental rotor harmonic to produce torque. Therefore, it is convenient to model the stator field in one expression as a thin current sheet which represents the sum of the harmonics produced by the balanced three phase winding.

\[ H_{(\mu)}(x_s, t) = \sum_{\mu=1}^{\infty} h_{(\mu)} \cos(\omega_s t - \pi \mu x_s), \]  

(3.1)

where

\[ h_{(\mu)} = \frac{\sqrt{2} N_s C_{sl} I_K(\mu)}{\pi d_{sl}} \]  

(3.2)

where \( \mu \) is the spatial harmonic, \( h_{(\mu)} \) represents the current loading magnitude of the \( \mu \)th spatial harmonic, \( \omega_s \) represents the stator electrical angular frequency, \( x_s \) represents the stator x variable, and \( t \) is time. The higher order time harmonics are ignored as pure sinusoidal stator current waveforms are assumed.

The winding current can be represented in a method suggested in [20]:

\[ i_{x,ABC}(k) = \sum_{k=1}^{N_s} i_{x,ABC}(k) \]  

(3.3)

where

\[ H(x) = 0 \text{ for } x \neq \text{ slot opening} \]

In equation (3.3), \( i_{x,ABC}(k) \) is the expression containing the current for each slot for any phase along the stator surface, \( x_k \) is the geometric centre of the slot \( k \). This expression in (3.3) is aimed at simulating the exact pattern of magnetic vector potential along the stator surface, which depends on manually simulating the winding in the function \( i_{x,ABC}(k) \). This method equips one with the flexibility to be able to simulate any possible stator winding. The fourier transform follows:

\[ h_{(\mu)} = \frac{1}{\pi} \int_{0}^{2\pi} H(x)[\cos(\mu x) + \sin(\mu x)]dx \]  

(3.4)
The harmonic number \( \mu \) takes any positive integer. For harmonics that do not exist, equation (3.5) gives a value of zero. The modulus can be written as:

\[
h_{(\mu)} = \frac{\sqrt{2}N_s C_{sl}}{\pi d_{sl}} \sin \left( \frac{\mu (b_{sl})}{d_{sl}} \right) \sum_{k=1}^{N_s} i_{x,ABC}(k) \left[ \cos(\mu x_k) + i \sin(\mu x_k) \right]
\]

(3.5)

The method of calculating the winding factor and hence, the magnitude of each harmonic in (3.6) is based on the premise that the winding will be ‘created’ in software. Given that the analytical calculations are computed in software, this method of analytical harmonic calculation was found to be appropriate.

The next step is to transform this stator current loading onto the moving rotor reference frame. The rotor speed has no effect on the magnitude of the stator current loading; however the relative speed of each harmonic must be adjusted accordingly. The relative speed can be defined as follows:

\[
\omega = \frac{2\pi n_s}{60} \left( \frac{N_p}{2} - \mu \right)
\]

(3.7)

One can see from equation (3.7) the inclusion of the harmonic order, \( \mu \). This indicates that each stator winding harmonic moves at its own speed within the rotor. It can be noted at this point that when \( \mu \) equals the working harmonic \( N_p/2 \), the relative speed \( \omega \), equals zero. One expects this result, as the working harmonic rotates synchronously with the rotor, therefore the relative speed equals zero.

### 3.2.2 Rotor Magnetic Field Calculation

The previous section focused on calculating the current harmonics in a stator winding by defining a current sheet on the surface of the stator. This section will focus on calculating the magnetic fields induced by the rotor magnets. Given that the goal of this section is to set up the necessary functions needed for rotor eddy loss calculation due to slotting, this section will represent the rotor field in a manner appropriate for the calculations. The method is similar to the previous section in that it seeks to represent the rotor magnetic field harmonics with a current sheet, and aims to determine the harmonic magnitudes and their relative speeds.
The no-load magnetic fields consist of two major groups of harmonics. There are those harmonics caused by the rotor's static magnetic field, which rotate synchronously with the rotor. The second group of harmonics arises due to the interaction of the static rotor fields and the permeance variation of the slots and teeth in the stator. The rotor magnetic flux function is:

\[ B_r(x, y) = \sum_{k=1,3,5,...}^{\infty} B_{rem}(y) \cos \left( k \frac{N_p}{2} x \right) \]  

(3.8)

where

\[ B_{rem}(y) = \frac{4 r_{mc} B_{rem}}{\pi N_p k} \sin \left( kp \tau_p \right) \]  

(3.9)

where \( r_{mc} \) represents the radius at the magnet centre. In order for eddy currents to be induced in a conductor there must be a relative speed between the conduction medium and the field harmonic. This group of harmonics in (3.8) rotates at a frequency of \( 2\pi n_s \), which is synchronous with the rotor. However, when one combines this static magnetic field harmonic with the permeance variation of the stator slots, a new asynchronous set of harmonics is produced.

The permeance variation function is described in [19] in a 2-D model which uses a conformal transformation assuming a unit magnetic potential applied between the rotor and stator surfaces and assumes infinitely deep rectilinear slots. The permeance function is defined as:

\[ \lambda(x, y) = \sum_{\mu=0}^{\infty} \lambda_\mu(y) \cos \left( \mu N_s x \right) \]  

(3.10)

where

\[ \lambda_0(y) = \frac{1}{K_c} \left( 1 - 1.6 \beta \frac{b_0}{\tau} \right) \]  

(3.11)

and

\[ \lambda_\mu(y) = -\beta(y) \frac{4 \pi \mu}{\pi \mu} \left[ 0.5 + \frac{\left( \mu \frac{b_0}{\tau} \right)^2}{0.78125 - 2 \left( \mu \frac{b_0}{\tau} \right)^2} \right] \sin \left( 1.6 \pi \mu \frac{b_0}{\tau} \right) \]  

(3.12)

and

\[ \beta(y)|_{y=R_c} = \frac{1}{2} \left[ 1 - \frac{1}{\sqrt{1 + \left( \frac{b_0}{2g} \right)^2}} \right] \]  

(3.13)

As this calculation is only interested in the permeance variation as seen by the rotor surface, the simplified version of the \( \beta(y) \) function is used, where \( v=0 \). It is also worth noting that in (3.11), the Carter Factor has been used in the definition of the DC permeance variation Fourier coefficient to account for the overall reduction in flux due to slotting. The definition of the Carter factor is:

\[ K_c = \frac{\tau}{\tau - \gamma g} \]  

(3.14)
where

\[ \gamma = \frac{4}{\pi} \left[ \frac{b_0}{2g'} \tan^{-1}\left( \frac{b_0}{2g'} \right) - \ln \left( 1 + \left( \frac{b_0}{2g'} \right)^2 \right) \right] \tag{3.15} \]

and

\[ \tau = \frac{2\pi R_s}{N_s} \tag{3.16} \]

One quantifies the effect of slotting given various slotting dimensions by multiplying the permeance variation function by the static field created by the rotor.

\[ B_{rotorsurface}(x, y) = \sum_{\mu=0}^{\infty} A_\mu(y) \cos (\mu N_s x_s) \times \sum_{k=1,3,5,...}^{\infty} B_{rem}(y) \cos \left( k \frac{N_p}{2} x \right) \tag{3.17} \]

In order to simulate the machine’s movement, the rotor is defined as the stationary reference frame and the stator is moved:

\[ x_s = x + \omega t \tag{3.18} \]

This gives:

\[ B_{rotorsurface}(x, y) = \sum_{\mu=0}^{\infty} \sum_{k=1,3,5,...}^{\infty} A_\mu(y) \cos (\mu N_s (x + \omega t)) \cdot B_{rem}(y) \cos \left( k \frac{N_p}{2} x \right) \]

\[ = \sum_{\mu=0}^{\infty} \sum_{k=1,3,5,...}^{\infty} \frac{A_\mu(y) B_{rem}(y)}{2} \cos \left[ (\mu N_s \pm k \frac{N_p}{2}) x + \mu N_s \omega t \right] \tag{3.19} \]

The important thing from a rotor loss perspective is to see that each space harmonic is a function of the rotor field and the stator permeance harmonics. The space harmonics also operate at frequencies that all are asynchronous to the rotor frequency which makes them able to induce eddy currents. In the chapters where we compute the magnetic fields in the rotor, the function in (3.19) is required to be expressed as magnetic vector potential:

\[ A(x, y) = -\int B_y dx = -\sum_{\mu=0}^{\infty} \sum_{k=1,3,5,...}^{\infty} \frac{A_\mu(y) B_{rem}(y)}{2} \sin \left[ (\mu N_s \pm k \frac{N_p}{2}) x + \mu N_s \omega t \right] \tag{3.20} \]

### 3.2.3 Prediction of Magnetic Vector Potential in Airgap, Magnets, Yoke

In the previous two sections, a thin current sheet on the stator surface was defined to represent fields due to the stator winding and the rotor fields. In this section, the magnetic field produced from this loading definition is calculated throughout the airgap, magnets and rotor regions. In this work, a similar process is followed as in [16]. For this calculation, the magnetic vector potential is used as an intermediate variable, from which other field calculations can be made.

\[ \nabla \times A = B \tag{3.21} \]

\[ \nabla \cdot A = 0 \tag{3.22} \]

The governing differential equation of the magnetic vector potential is Poissonian:

\[ \nabla^2 A - j\omega \mu A = 0 \tag{3.23} \]
The second term in equation (3.23) is dependent on the conductivity of the material, so in the airgap where the conductivity is zero, this term vanishes. Solving the differential equation in (3.23) relies on an assumption that the function is separable. That is to say:

\[ A(x, y) = A(x)A(y) \]  

(3.24)

The separation of variables method then returns a new form of equation (3.23):

\[ \nabla^2 \mathbf{A} - j\omega \kappa \mu \mathbf{A} = \frac{d^2 A(x)}{dx^2} \frac{1}{A(x)} + \frac{d^2 A(y)}{dy^2} \frac{1}{A(y)} - j\omega \kappa \mu = 0 \]  

(3.25)

This produces the general solution for \( A \):

\[ A(x, y) = (C_y e^{ry} + D_y e^{-ry})e^{-j\alpha x} \]  

(3.26)

where \( \gamma^2 = a^2 + j\omega \kappa \mu \), and \( \alpha = \mu \pi \)  

(3.27)

Equation (3.26) is the general solution for conductive regions; however in the airgap the general solution can be simplified to:

\[ A_{airgap}(x, y) = (C_y e^{ay} + D_y e^{-ay})e^{-j\alpha x} \]  

(3.28)

The magnetic vector potential in each region is governed by either equations (3.23) or (3.28) in 2-D space depending whether it is a conductive region or not. In order for these equations to be used, one needs to define the constants \( C_y \) and \( D_y \). Each region has a different value of permeability and conductivity, giving rise to different constants for the airgap, magnets and yoke. In order to solve for these constants, one needs to setup a series of boundary conditions at the stator surface, the outer yoke surface and the interfaces of each of the region. The conditions defining each boundary include:

\[ B_{n,k} = B_{n,k+1} \]  

(3.29)

\[ H_{t,k+1} - H_{t,k} = A_k \]  

(3.30)

\[ B_x(x, y) = \frac{dA}{dy} = \gamma(C_y e^{ry} - D_y e^{-ry})e^{-j\alpha x} \]  

(3.31)

\[ B_y(x, y) = -\frac{dA}{dx} = j\alpha(C_y e^{ry} + D_y e^{-ry})e^{-j\alpha x} \]  

(3.32)

\[ H_x = \frac{B_x(x, y)}{\mu_0 \mu_r} \]  

(3.33)

\[ H_y = \frac{B_y(x, y)}{\mu_0 \mu_r} \]  

(3.34)

Equation (3.29) is simply an implementation of the condition in Gauss’ Law stating that the magnetic flux entering a surface must equal the flux leaving. In this case, the condition applies to the normal flux component in the y-direction. Equation (3.30) states that any discontinuity in the tangential magnetic field strength from one region to the next is due to the presence of a current sheet. Equations (3.31-3.34) comprise the expressions from which (3.29) and (3.30) are built.
In the airgap region (region 1), the stator surface contains a current loading defined in (3.1), and the airgap/magnet interface obeys conditions (3.29) and (3.30).

Region 1:

\[ H_{y_1}(x, y) = \frac{1}{\mu_0} \frac{dA}{dy} = C_1 + D_1 e^{-ax} = h_{(\mu)} e^{-j\alpha x} \]  

(3.35)

\[ B_{y_1}(x, y) = -\frac{dA}{dx} = j\alpha(C_1 e^{ay} + D_1 e^{-ay})e^{-j\alpha x} \]  

(3.36)

\[ B_{x_1}(x, y) = \frac{dA}{dy} = \alpha(C_1 e^{ay} - D_1 e^{-ay})e^{-j\alpha x} \]  

(3.37)

In the magnet region, the conductivity is non zero and permeability is similar to air:

Region 2:

\[ B_{y_2}(x, y) = -\frac{dA}{dx} = j\alpha(C_2 e^{y_2 y} + D_2 e^{-y_2 y})e^{-j\alpha x} \]  

(3.38)

\[ B_{x_2}(x, y) = \frac{dA}{dy} = \gamma_2(C_2 e^{y_2 y} - D_2 e^{-y_2 y})e^{-j\alpha x} \]  

(3.39)

The rotor yoke region is made from solid steel and a boundary condition of no leakage flux in the back yoke is enforced. This condition is implemented in equation (3.42).

Region 3:

\[ B_{y_3}(x, y) = -\frac{dA}{dx} = j\alpha(C_3 e^{y_3 y} + D_3 e^{-y_3 y})e^{-j\alpha x} \]  

(3.40)

\[ B_{x_3}(x, y) = \frac{dA}{dy} = \gamma_3(C_3 e^{y_3 y} - D_3 e^{-y_3 y})e^{-j\alpha x} \]  

(3.41)

\[ H_{y_3}(x, y) = \frac{1}{\mu_0} \frac{dA}{dy} = \gamma_3(C_3 e^{y_3 y} + D_3 e^{-y_3 y})e^{-j\alpha x} = 0 \]  

(3.42)

Using equations (3.35-3.42) a list of interface conditions can be compiled in linear equation form:

Interface 1:

\[ B_{y_1}(x, y) = B_{y_2}(x, y) = j\alpha(C_1 e^{ay} + D_1 e^{-ay}) = j\alpha(C_2 e^{y_2 y} + D_2 e^{-y_2 y})e^{-j\alpha x} \]  

(3.43)

\[ B_{x_1}(x, y) = B_{x_2}(x, y) = \alpha(C_1 e^{ay} - D_1 e^{-ay})e^{-j\alpha x} = \gamma_2(C_2 e^{y_2 y} - D_2 e^{-y_2 y})e^{-j\alpha x} \]  

(3.44)

Interface 2:

\[ B_{y_2}(x, y) = B_{y_3}(x, y) = j\alpha(C_2 e^{y_2 y} + D_2 e^{-y_2 y}) = j\alpha(C_3 e^{y_3 y} + D_3 e^{-y_3 y})e^{ay y} \]  

(3.45)

\[ B_{x_2}(x, y) = B_{x_3}(x, y) = \gamma_2(C_2 e^{y_2 y} + D_2 e^{-y_2 y})e^{-j\alpha x} = \gamma_3(C_3 e^{y_3 y} + D_3 e^{-y_3 y})e^{-j\alpha x} \]  

(3.46)

Combining (3.43-3.46) with (3.35) and (3.42), a system of linear equations is produced:

<table>
<thead>
<tr>
<th>( C_1 )</th>
<th>( D_1 )</th>
<th>( C_2 )</th>
<th>( D_2 )</th>
<th>( C_3 )</th>
<th>( D_3 )</th>
<th>( \mu_0 h_{(\mu)} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>( e^{ay y} )</td>
<td>( e^{-ay y} )</td>
<td>( -e^{y_2 y} )</td>
<td>( -e^{-y_2 y} )</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>( a e^{ay y} )</td>
<td>(-ae^{-ay y} )</td>
<td>(-e^{y_2 y} )</td>
<td>( ye^{-y_2 y} )</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>( e^{y_2 y} )</td>
<td>( e^{-y_2 y} )</td>
<td>(-e^{y_3 y} )</td>
<td>(-e^{-y_3 y} )</td>
<td>0</td>
</tr>
</tbody>
</table>

(3.47)

(3.48)

(3.49)

(3.50)
This set of linear equations (3.47-3.52) can be solved to produce the constants needed to define the magnetic vector potential throughout the airgap, magnet and rotor yoke regions in the general solution shown in (3.26) and (3.28). All subsequent calculations pertaining to eddy currents and solid loss are also reliant on these constants.

3.3 Eddy Current Calculation

Thus far in the chapter, the following quantities have been defined:

- the current sheet defining field harmonics due to the stator winding and stator slotting,
- the speed at which each of those harmonics are moving relative to the rotor,
- the magnitude of each harmonic in space throughout the airgap, magnets and rotor.

This section will focus on using all of this information to compute the eddy current and losses in each region.

3.3.1 Calculating Eddy Currents from Magnetic Vector Potential

The expression for the electric field induced in a conductor is defined as the time rate of change of the magnetic vector potential, summed with a \( \text{grad} \) term which is constant in the \((x,y)\) plane,

\[
E(x, y) = -j\omega A(x, y) + \text{grad}\phi \tag{3.53}
\]

In this work, currents will be assumed to flow only in the positive and negative \(z\) direction. The current density is computed as:

\[
J(x, y) = J_z(x, y) = -j\omega \kappa A(x, y) + \mathcal{C}
= -j\omega \kappa (C_y e^{j\gamma y} + D_y e^{-j\gamma y}) e^{-j\alpha x} + \mathcal{C} \tag{3.54}
\]

3.3.2 Eddy Currents in Solid Conductors

In a solid conductor, eddy current flow is limited only by the material conductivity. Using the magnetic vector potential, harmonic frequency and conductivity, the ohmic loss due to heat produced by eddy currents can be calculated by:

\[
P = \frac{l}{2k} \cdot \overline{J(x, y) \cdot J(x, y)^*} \tag{3.55}
\]

3.3.3 Eddy Currents in Segmented Magnet Conductors

Kirchhoff’s Law states that electrical currents at a point must sum to zero. Application of this law to a solid conductor containing infinite points, dictates that circulating eddy currents in a conductor must sum to zero. The constant term in the current density function is used to offset any residual DC current so that the forward and return eddy currents balance. This is especially relevant in segmented conductors as this condition is enforced on each segment.
To enforce this condition, the eddy currents in each isolated conductor (segment) are averaged. This average current density value is then subtracted from the eddy current density function to leave the region with a zero average value of current density.

\[ J_{\text{segment average}}(x, y) = \int_0^{h_m} \int_0^{l_s} J(x, y) \, dx \, dy \]  

(3.56)

Where, \( h_m \) and \( l_s \) are magnet height and segment length respectively. The expression for the current density in a segmented conductor ensuring the zero total conductor current condition follows:

\[ J_{\text{post segmentation}}(x, y) = \sum_{i=0}^{N_p-1} \sum_{k=0}^{N_{SS}-1} J(x, y) - \int_0^{h_m} \int_0^{l_s} \left( \frac{2\pi}{N_p} \frac{2\pi(k+1)}{N_{SS}} \right) R_s \, J(x, y) \, dx \, dy \]  

(3.57)

This expression redefines the eddy currents calculated in 3.2.1 by subtracting the average value of current in each isolated, separate segment. This phenomenon is simply a mathematical deployment of the logic that an induced eddy current of wavelength larger than a segment pitch cannot flow between insulated conductor segments.

It should be noted at this junction, that the subtraction of the average value of the eddy current as indicated in equation (3.57) does have an impact on the field solution. The added imposed current loading acts to retrospectively influence the field solution which is not taken into account in the solution as this occurs in post processing. The influence of this is to possibly reduce the eddy current reaction field in the magnets which can shield the rotor yoke from excessive changes in flux density. To mitigate the effect of this inaccuracy, another calculation method is needed. One method would be to institute an iterative method that repeats the field solution calculation making a change each time to ensure that the zero current boundary condition for each isolated conductor.

3.3.4 Eddy Currents in Segmented Rotor Yoke Conductors

In a conductor which has no poles, the segmentation calculation is simpler as it disregards the condition of zero total current in each pole. In this case, such as with rotor segmentation, the following condition applies:

\[ J_{\text{post segmentation}}(x, y) = \sum_{n=0}^{N_{SS}-1} J(x, y) - \int_0^{h_y} \int_0^{l_y} \left( \frac{2\pi}{N_{SS}} \right) R_s \, J(x, y) \, dx \, dy \]  

(3.58)
3.4 Loss Calculation

The final calculation in this chapter determines the solid conductor losses in the rotor of a machine based on the previously computed eddy current calculations.

3.4.1 Loss Calculation from Current Density

The eddy current loss follows the current density calculation and requires very little more processing to compute the result. This solution requires the current density to be computed at points throughout the surface.

\[
P = \frac{2\pi l_r l_m}{2\kappa} \left| J(x, y) \cdot J(x, y) \right|^*\tag{3.59}
\]

The average power loss due to ohmic losses in a solid conductor is proportional to the conductor volume and conductivity, and the square of the current density and harmonic angular velocity.

3.4.2 Loss Calculation using the Skin Depth

The skin effect is a phenomenon that describes the tendency of alternating current to distribute itself non-linearly in a conductor. The current is largest at the surface or 'skin' and decays exponentially with distance. The rate of decay can be described with a quantity called the 'skin depth', which is the distance at which the current density has decayed to \(1/e\) of its surface value.

The current density in a conductor with an AC voltage across its ends can be described as:

\[
J = J_s e^{-\frac{y}{\delta}}\tag{3.60}
\]

where the skin depth is given by:

\[
\delta = \frac{2}{\sqrt{\omega \sigma \mu}}\tag{3.61}
\]

The skin depth can be used in loss calculations to provide a relatively easy method of calculating the losses in a solid conductor while also taking the reaction field into account. Mathematically, the surface current value multiplied by the skin depth equals the current density function integrated over infinity. This is to say:

\[
\int_0^\infty J_s e^{-\frac{y}{\delta}} dy = J_s \delta
\]

In a conductor where the thickness is far larger than the skin depth, the thickness can be approximated as infinity. In this case, the skin depth can be used to calculate current density and average power:

\[
P_{\text{ave}} = \frac{2\pi l_r l_m \delta}{2\kappa} \frac{J_s}{\delta} \tag{3.62}
\]
3.5 Finite Element Modelling

The advantage of analytical calculations is that the formulas do not only produce results, but also explain them, which is helpful in design as it promotes understanding. However, often during the analytical process, assumptions are made in order to simplify models, which can reduce accuracy. Finite element computation methods aim for complete accuracy and versatility.

A machine can either be described by a small number of complicated equations, or it can be described by many simple equations. This is the essential difference between an analytical model and a finite element model. Finite element analysis breaks up a machine into a mesh of individual nodes and performs simple field calculations for each point based on material properties and proximity to excitation. These results are collated and summed to produce an overall result.

In this study, the finite element analysis is used to verify calculations performed analytically and also provide insight into the machine’s behaviour by providing additional visual graphics.

3.5.1 Model

The FE software used in this study was Ansoft Maxwell 2D. The machine model is input with a simple CAD drawing of the machine containing all dimensions. Each material is defined according to its physical properties. The rotor is comprised of the magnets and solid yoke which is defined as a moving band.

3.5.2 Boundary Conditions

Every finite element package has to have boundary conditions stipulated. The software uses these for calibration, continuity and also to save computation by including periodicity. The outer boundaries are designed to enclose the machine and the FE calculation region by defining lines of zero magnetic vector potential. The master and slave boundaries allow one to model just a fraction of the machine and duplicate the results with positive or negative periodicity.

3.5.3 Excitations

The FE software defines windings by grouping current or voltage carrying coils into phase belts. Aligning the current axes with the magnetic axes can be done by injecting current into the d-axis and adjusting the machine angle until a torque value of zero is measured. The excitations are not limited to the armature as one also needs to define regions of zero current for insulated segmented conductors. If this is not selected, segmentation will have no effect on solid loss, rendering the results inaccurate.
3.5.4 Mesh Operations

The mesh definition guides the software in determining the position and quantity of the finite elements (nodes) required to construct the mesh. Each part of the machine can be selected and the minimum distance between nodes within the material and its edges can be defined. The general guideline is that the FE model becomes more accurate as the number of nodes increases. In this study, a very fine mesh was selected in the rotor yoke and magnets as these were areas of particular interest.

3.5.5 Solve Setup

Another important factor in the FEM model’s accuracy is the length of the solution time step. This quantity defines how much time separates each solution. Maxwell 2D computes some quantities directly from the magnetic vector potential such as flux linkage and magnetic field strength. These results are usually very accurate as they depend only on excitation and mesh geometry. However, some quantities require the use of a time derivative for their calculation, e.g. current density or voltage. The time derivative is a continuous function modelled as a discrete function, which can create inaccuracies if the chosen time step is not small enough.

3.5.6 Specific Eddy Current Calculation

The method for specific eddy current calculation in Maxwell 2-D follows in the flow diagram shown in Fig. 3-4. These four checks are essential for accurate solid loss results. The first step is to define the conductivity of the materials which must be consistent with the materials used in construction. The next step is to allow eddy currents in the regions to be able to flow. This automatically also sets the calculator to include the reaction fields due to eddy currents in the field solution. The boundary conditions must be set to define solid regions, and in order to achieve this, one defines the total current in each isolated conductive region as zero. The final
step is to use the fields calculator which can compute solid loss, which calculates eddy current loss in all the conducting regions, or ohmic loss which allows one to calculate region specific solid loss.

**Fig. 3-4:** Flow chart representing the steps that need to be taken in Maxwell 2D in order to calculate solid loss.

The calculation of solid loss is done internally within the software and it automatically looks at the multiplication factor of the machine and the stack length and performs the necessary multiplications. For example, if you are modelling 1/8 of a machine with a stack length of 100mm (1/10m), then the model will automatically multiply the results by 8 and divide by 10, however if one uses the ohmic loss calculator, this gives more control to the user, but the user needs to perform this scaling multiplication manually.
4 Partial Segmentation of Magnets and Rotor Yoke

4.1 Introduction to Magnetization

The presence of a strong magnetic field in a material acts to align the particle poles. In permanent magnet material, the particle poles remain aligned after the field is removed producing a static magnetic field. This process is called magnetization. Demagnetization is caused by interference in the molecular pole alignment, caused by:

- a strong external magnetic field, or
- a strong thermal disturbance.

In an electric machine, the performance is dependent on the ability of the magnet to perform at full strength. Under normal operating conditions, demagnetization due to a strong magnetic field from the stator is unlikely; however overheating of the magnets poses a threat to the magnetization. Heat generation in a machine can be attributed to many sources, but the magnets are most vulnerable to heat that is produced from within, by eddy currents. For this reason, eddy currents are highlighted not only as a potential source of inefficiency, but also as a threat to the magnet’s condition.

In machine design, one can take a myriad of design decisions aimed to reduce the effect of harmful harmonics causing the flow of eddy currents in the rotor. Decisions on winding layout [17], airgap thickness [16,18], buried or surface mounted magnets, magnet thickness, closed or open stator slots are some factors that can affect the rotor loss in a machine. These designs often include a tradeoff penalty such as loss of torque or voltage due to leakage flux, or more expensive and time-consuming construction methods. One needs to optimize the machine so that the design benefits and trade-offs are balances by cost, performance and efficiency.

If one aims to reduce magnet losses without making geometric or design alterations, improvements must be made to the material itself to increase resistivity. The properties of the material are approximately constant, but using insulated fragments or segments can increase the resistivity ‘seen’ by eddy currents. This brings about the focus of this chapter which describes segmentation.

4.2 Magnet Segmentation Models

For analytical calculations, various segmentation models are presented.

4.2.1 Resistivity Model

This method of segmentation acts to interrupt the flow of current within a magnet and forces the currents to flow through a narrower area and over longer distances.
Chapter 4  Partial Segmentation of Magnets and Rotor Yoke

Fig. 4-1: The diagram shows an unsegmented magnet, its dimensions and the flow of eddy currents within its boundaries. The labelling of its axes are consistent with the model in chapter 3.

One way of describing the effectiveness of segmentation is to look at an unsegmented magnet as in Fig. 4-1 and describe the resistance of the path 'seen' by the induced eddy current.

\[ R_z = \frac{\rho \text{ length}}{\text{area}} = \frac{2\rho l}{wh} = \frac{4\rho l}{2wh} \]

(4.1)

and similarly

\[ R_x = \frac{4\rho w}{lh} \]

(4.2)

\[ R_{\text{total}} = \frac{4\rho l}{wh} + \frac{4\rho w}{wl} = \frac{4\rho}{w/lh} [l^2 + w^2] \]

(4.3)

It becomes clear from this formula that the eddy current resistance path is proportional to the width and length of the magnets, and also to the square of the level of segmentation. In segmentation in the z-direction, the \( N^2 \) term is the co-efficient of the \( l \) term which is usually

Fig. 4-2: This is the same magnet as in Fig.4-1, however this magnet is segmented into \( N \) insulated pieces each with a width of \( w/N \).

If the magnet is segmented down the length of the machine, i.e.: radial segmentation, the resistance 'seen' by the eddy currents changes as the area and length of its path of travel are different. So the new resistance is:

\[ R_z = \frac{2N \cdot \rho l}{wh} = \frac{4N^2 \rho l}{wh} \]

(4.4)

and including \( x \) resistance,

\[ R_{\text{total}} = 4 \left[ \frac{N^2 \rho l}{wh} + \frac{\rho w}{lh} \right] = \frac{4\rho}{w/lh} [N^2 l^2 + w^2] \]

(4.5)
much larger than the $w$ term. If one chooses segmentation in the x-direction, the effect of the segmentation is not as great as the $N^2$ term becomes the co-efficient of the much smaller $w$ term. The relationship in equation (4.5) can be used as a guide, but more importantly as a way of understanding why segmentation can be effective.

![Fig. 4-3](image1)

**Fig. 4-3:** An unsegmented magnet with a sinusoidal current density waveform.

The resistivity model only provides a guide to segmentation and should not be relied upon for accurate results.

### 4.2.2 Current Density Subtraction Model

![Fig. 4-4](image2)

**Fig. 4-4:** A segmented magnet with a sinusoidal current density waveform.

The second method of describing how magnet segmentation is effective involves a visual interpretation of equation (3.58), which is also suggested in [20]. This method is used by the analytical and finite element analysis calculations to implement segmentation. It works on the
principle that forward and return currents in an isolated conductor must balance each other. Therefore, the average current value in a conductor must be subtracted from each value of current in that same conductor which can give an overall reduction in current. If one segments the magnet into two pieces as in Fig 4-4, the current can no longer flow across the two halves of the magnet. This causes it to flow between magnetic field gradients within its own bounds which are often substantially lower causing smaller eddy current magnitudes and reducing the overall solid loss.

### 4.2.3 Segmentation Model Comparison

If one compares the two models with single cut magnet segmentation, the results show a disparity. Assuming current in the z-direction only, the resistivity model shows a loss reduction of 75%, whereas the current density model shows a loss of 55%. This discrepancy is further explored in section 4.5.

The resistivity model makes the incorrect assumption that there are even currents flowing throughout the magnet at all times. Rotating harmonics of differing wavelengths penetrate different parts of the magnet periodically, giving rise to a complex time varying current density topography in the magnet. For this reason, the resistivity model is not the preferred calculation method as resistivity is not always the only factor limiting currents.

The average current density model takes the current density waveform which can be calculated by FEM or analytically and computes the average value and subtracts. This approach renders the model particularly attractive due to its versatility. The average current density model is therefore the preferred method to calculate the effect of segmentation and will be used throughout the rest of this study.

### 4.3 Full Magnet Segmentation

In this study the normal magnet segmentation will be described as full magnet segmentation (FMS) in order to differentiate it from other segmentation types.

#### 4.3.1 Description

Full magnet segmentation describes a construction process where solid magnets are split into a series of electrically insulated, geometric fragments and rebonded together to reconstruct the original magnet dimensions.
4.3.1 Partial Segmentation of Magnets and Rotor Yoke

One assumes that the segmented magnet has the same total strength as the original magnet:

\[ M_{total} = \sum_{i=1}^{Np} M_i = \sum_{i=1}^{Np} \sum_{k=1}^{Nss} M_{i,k} \]  \( (4.6) \)

In practical applications, this can be a false assumption as there exists a non-conductive and non-magnetic epoxy layer designed to insulate each segment. This will influence the strength of the magnet, but for purposes of demonstration, this study will ignore these effects.

4.3.2 Types of Segmentation

The two main types of segmentation differ in the direction in which the segmentation cuts are aligned.

The implementation of cuts aligned in the z-direction or in the direction of the machine's axial length is called radial segmentation, and those cuts that are aligned concentrically about the machine's shaft (x-direction) implement axial segmentation. In this work, the main focus is on radial segmentation; however the research is equally applicable to axial segmentation.

4.3.3 Manufacturing

The manufacture of FMS requires knowledge of the dominant physical properties of the NdFeB magnet:

- **Britleness** - the magnets originate from a powder and are heated up to sub-melting point temperatures in a bonding process called sintering. This renders the magnets...
quite brittle which makes cutting a specialized process to avoid fracture and fragmentation.

- **Strength** – the NdFeB magnet is magnetically one of the strongest of all magnets, and processes such as cutting and bonding become especially more difficult post magnetization due to the huge attractive forces between magnet pieces.

To avoid the almost intolerable manufacturing complications that arise with the handling of magnetized segment pieces, it is easier for magnets to be segmented before magnetization. The fragments must also be bonded with a strong epoxy to overcome the internal forces that are present within the magnet after magnetization. Some machines with internal rotor magnets can accept axially segmented magnet pieces as these can be slid into the rotor by means of a machine.

### 4.3.4 Disadvantages of Full Magnet Segmentation

For the reasons outlined in chapter 4.3.3, the disadvantages of FMS are mostly concerned with the manufacturing process:

- **Cost** – Due to the intensive cutting and bonding process, the production of a segmented magnet is time consuming, labour intensive and therefore costly.
- **Lack of versatility** – The segmentation process must take place pre-magnetization and therefore by the manufacturer of the magnets. This does not give the machine manufacturer any flexibility about degrees of segmentation after
- **Structural weakness** – A solid magnet is more structurally stable and strong than a segmented magnet that has bonded surfaces.

### 4.4 Partial Magnet Segmentation

Full magnet segmentation has been shown to have a good effect on machine performance by reducing losses, however these benefits can be outweighed by manufacturing complications and cost. Partial magnet segmentation (PMS) aims to address these drawbacks while still achieving to some degree the gains and benefits of FMS.

#### 4.4.1 Description

Partial magnet segmentation as shown in Fig. 4-7 describes a manufacturing technique where the incisions only penetrate a portion of the solid magnet. The purpose of this method is to keep the magnet in one piece during construction so that various manufacturing benefits can be realised. These advantages include:

- Due to the magnet not being cut into separate pieces, the bonding process is eliminated which saves time, cost and complexity.
• The method can be implemented pre or post magnetization, giving more flexibility to the machine manufacturer.

• Customization is improved as this technique can be implemented at the machine factory level instead of only at the magnet manufacturer.

• Speed is improved in mass production as dedicated cutters can be used in the machine factory assembly line.

**Fig. 4-7:** A partially segmented magnet, showing cuts in the y-direction and the cut penetration.

### 4.4.1.1 Segmentation Penetration

The degree of segmentation describes the depth of the incision in the y-direction, if the degree of penetration is 100%, then one effectively has a fully segmented magnet. For performance reasons, one would like as high a penetration as possible. However the internal forces that exist within a magnetized magnet are the chief limiting factor to penetration depth as they can cause weakening of the magnet leading to possible fracture. An epoxy coating and fill between the segments is helpful to prevent structural failure by reducing magnet’s susceptibility to damage during construction. It also protects the magnet against the non-normal attractive forces from the stator teeth.

### 4.4.1.2 Single-Sided and Double-Sided Partial Magnet Segmentation

There are two major types of partial magnet segmentation suggested in this work. Single-sided PMS employs incisions into the magnet from one side only, whereas double-sided PMS allows additional incisions to be made from the other side of the magnet in the opposing direction. DS-PMS aims to increase the effectiveness of partial segmentation which still retaining the major benefit of keeping the magnet in one piece. One expects the implementation of DS-PMS to achieve a similar loss reduction profile achieved by FMS given that the combination of incisions from both directions adds up to full segmentation. Both of these techniques are looked at in detail in this study.
4.4.2 Model

The analytical work performed in chapter 3 forms a good base of understanding for how PMS can be effective in reducing losses in a machine. The theory is able to describe the current density functions in the $x$ and $y$ dimensions, which is used to guide decisions regarding the degree of segmentation, penetration percentage and optimal placement of segments. The current density functions can also be used to model the effect of segmentation for accurate quantification of results, where a model, derived from the combination of two models described in chapter three is used to build the PMS model. A partially segmented magnet is essentially a combination of a segmented magnet portion and a non-segmented magnet portion:

\[
J_{PMS}(x, y) = \sum_{n=0}^{N_p-1} j_m R_s \left( \frac{2\pi n N_p+1}{N_p} \right) R_s \int_0^{N_p N_{ss}} j(x, y) \, dx \, dy
\]

\[
+ \sum_{n=0}^{N_p-1} \sum_{k=0}^{N_{ss}-1} j(x, y) \int_{h_m R_s}^{h_m R_s+1} \left( \frac{2\pi n+k}{N_p N_{ss}} \right) R_s \, dx \, dy
\]

(4.7)

4.4.3 Manufacturing

One of the main advantages of partial magnet segmentation is that it allows the possibility of segmentation post magnetization. This possibility presents its own challenge in manufacture for the following reasons:

- Ferrous cutting materials such as steel blades are easily deflected by the magnet's field.
- Brittle nature of the sintered magnetic material makes it susceptible to small shards chipping off from the magnet itself.

It is clear that the main manufacturing challenge lies in the cutting process, which has to be done as thinly and neatly as possible.
4.4.4 Limitations
When considering the possibility of implementing partial magnet segmentation, one needs to weigh up a few disadvantages of the technology. These disadvantages include:

- **Material Loss:** During the segmentation process, magnet material is cut away leaving a small gap that is filled with epoxy. This reduces the overall amount of magnet material and, hence the strength.
- **Cost:** Magnet segmentation incurs extra cutting costs in tooling and blades. The extra time that it takes to cut the magnets will also increase manufacturing time and the overall cost of the machine.
- **Reduced structural strength during cutting:** Partially segmented magnets are weakened by the incisions made into them, so care must be taken during manufacture to avoid fracture before the epoxy is applied.

4.5 Partial Rotor Yoke Segmentation
Permanent magnet machines have large equivalent airgaps as the permeability of NdFeB magnet material is almost identical to that of air. Consequently, the rotor yoke is more isolated and protected from harmful harmonics than if the equivalent airgap was smaller. The rotor yoke acts as a return path for flux in the machine, and even though the flux is essentially static, there still can exist enough pulsating harmonics to induce eddy currents. The yoke of the rotor with non-overlap stator windings is traditionally made from laminated steel to prevent the flow of these eddy currents, however the question arises as to whether one can use a solid yoke rotor instead.

The laminations in a rotor yoke need to be encased in a hub to keep the laminations together and give them the structural rigidity that they need. A solid yoke rotor becomes an option as it:

- Eliminates the need for punching or laser cutting laminations.
- Contains enough mechanical strength to be used without support.

The solid yoke rotor is unfortunately also an electrically conductive region, aiding the flow of eddy currents. The method of partial rotor yoke segmentation aims to reduce this eddy current flow and the ensuing solid loss.

4.5.1 Description and Purpose
When a given set of asynchronous field harmonics are present in a solid conductor, the induced eddy current magnitudes are determined by the resistivity and the geometric dimensions. In order to reduce losses, one must find a way to increase the resistivity without changing the material. This section will describe a method of increasing the resistivity ‘seen’ by the eddy currents called *partial rotor yoke segmentation (PRYS)*.
Chapter 4  Partial Segmentation of Magnets and Rotor Yoke

Fig. 4-9: The diagram shows the solid yoke segments positioned on the magnet side of the solid rotor yoke.

PRYS is a technique where the solid yoke is finely cut along the surface adjacent to the magnets. The aim of the cut is to create isolated conducting regions and interrupting the eddy current path. The depth of the cut is called the segmentation penetration depth and an optimum value depends on the conductivity, permeability, and the frequency of the field harmonic in question. In most cases, a value much less than the thickness of the yoke can be selected.

4.5.2  Model

The current density model of PRYS is based on the analytical calculations in chapter three which describes the current density in the rotor yoke for each harmonic. In similar fashion to the partial magnet segmentation, PRYS is modelled by ensuring that forward and return currents in each segment are balanced by subtracting average values, shown in (3.58).

The main difference between modelling PRYS and PMS lies in the segmentation penetration depth. In the rotor yoke, due to the small skin depth caused by the steel’s high permeability and conductivity, the penetrating depth can reach an optimum level at a fraction of the yoke thickness. Therefore only single-sided partial rotor yoke segmentation is considered. For this reason the current density model is:

\[ I_{\text{post segmentation}}(x, y) = \sum_{k=0}^{N_{ss}-1} f(x, y) - \int_{0}^{P_{hy}} \int_{0}^{\frac{2\pi}{N_{ss}(k+1)}R_{s}} f(x, y) \, dx \, dy \]  

(4.8)

4.5.3  Manufacturing

The rotor yoke begins as a piece of sheet steel which is rolled, welded and then machined to the correct dimensions. To implement segmentation, thin incisions need to be made into the rotor. Two methods can be used for this:
• The incisions can be made before rolling, and as we can see from Fig. 4-10, that this causes the gap to partially close which is good for machine performance as flux density will improve.
• The incisions can be made after rolling and machining, which can lead to greater construction accuracy.

The options for cutting the steel are by using wire erosion or a thin cutting blade. Either of these options can be used depending what is best for the application. The important consideration is to ensure that the cutting width is as small as possible so as to interfere with machine performance as little as possible.

Fig. 4-10: The diagram shows the solid yoke segments positioned on the magnet side of the solid rotor yoke.

4.5.4 Limitations
Cutting thin incisions into the solid rotor yoke is aimed at increasing machine efficiency, but one must observe some disadvantages of implementing PRYS:
• Flux density reduction – the effect of cutting small gaps into the rotor yoke effectively increases the air in the magnetic circuit, causing a reduction in flux density. Magnets need to be made stronger or thicker to mitigate this effect.
• Cost – The time and cost of cutting and extra tool materials adds to the manufacturing cost.

4.6 Optimal Magnet Segmentation
It is a misconception that one can theoretically calculate a generic function to quantify the effect of segmentation in any type of machine. It is clear from most research that the effect of segmentation acts to reduce the eddy currents in a solid and therefore reduces loss. But quantification of the relationship between the degree of segmentation and the degree of loss reduction is not readily defined. Equation (3.59) outlines the principle of how segmentation
helps to reduce eddy currents flowing in a solid; however one can examine it to see how it performs for specific harmonics.

### 4.6.1 Static Segmentation Model

A quantity called relative harmonic segmentation pitch (RHSP) is defined for a given harmonic as the harmonic wavelength divided by the number of segments that that harmonic is divided into. The relative degree of segmentation ($N_{rs}$) is defined as the number of segments that the given harmonic is divided into.

$$\tau_{RHSP}(\mu, N_{rs}) = \frac{2\pi}{\mu N_{rs}}$$  \hspace{1cm} (4.9)

![Diagram of current density and magnet pitch](image)

**Fig. 4-11:** The figure demonstrates graphical examples of relative harmonic segmentation pitch.

In Fig. 4-12, there is magnet at the top and below it are two eddy current harmonic waveforms that make up the total eddy current waveform:

$$J(x) = J_1(x) + J_2(x) = f_1 \sin \left( \frac{\mu x}{\tau_p} \right) + f_2 \sin \left( \frac{2 \mu x}{\tau_p} \right)$$  \hspace{1cm} (4.10)

From equation 4.10, one can see that the current density comprises two harmonics which have the same wavelength and twice the wavelength as the pitch of the magnet respectively. Note in Fig. 4-12 that both of these harmonics are free to flow within in the magnet as the forward and return paths of the eddy currents are uninhibited by the geometry of the magnet.

In Fig. 4-13, the magnet is divided into two segments, which doubles the relative harmonic segmentation pitch of the magnet. The constraint in equation (3.58) is enforced, ensuring that each magnet segment has a zero residual current flowing within it.
Fig. 4-12: A magnet is shown with two eddy current harmonic waveforms that flow within it. The result is that the relative segmentation pitch for \( J_1 \) is two, and therefore has been affected by the segmentation, while \( J_2 \) has a relative segmentation pitch of one and is still allowed to flow uninhibited.

\[
J_{\text{Total}}(x) = \left| J_1 \sin \left( \frac{\mu x}{\tau_p} \right) - \int_0^{\tau_p} J_1 \sin \left( \frac{\mu x}{\tau_p} \right) \, dx \right|^{\tau_p/2}_0 \\
+ \left| \int_0^{\tau_p} J_1 \sin \left( \frac{\mu x}{\tau_p} \right) \, dx + J_2 \sin \left( \frac{2 \cdot \mu x}{\tau_p} \right) \right|^{\tau_p/2}_0
\]

(4.11)

This critical degree of segmentation where segmentation becomes effective can be defined as when the relative segmentation pitch reaches two. This critical point can be defined as the Nyquist Critical Segmentation Pitch.

\[
\tau_{\text{Nyquist}} = \frac{1}{2} \lambda_{\text{harmonic}}
\]

(4.12)

where \( \lambda_{\text{harmonic}} \) is the wavelength of the harmonic in question.

Fig. 4-14 shows a model whereby a solid is defined and a current density wavelength \( P \) is shown that is half the Nyquist wavelength for that solid. For further investigation into how the loss reduction behaves around the Nyquist Point, the \( S \) value is used which is equal to \( N_{\text{ss}} \). The \( S \) dimension defines the relative degree of segmentation, e.g. when \( S=2 \), the original solid is divided into two segments making the segmentation pitch equal to \( P/2 \).
Fig. 4-13: A segmented magnet and two eddy current harmonic waveforms that flow within it.

Fig. 4-14: The segmentation model shows a solid of relative segmentation pitch $S$, with a current density waveform of period $P$.

When $S=1$, the eddy current harmonic can flow freely and uninhibited within the bounds of the solid. As relative segmentation pitch ($S$) is changed from 1 to 10, so the wavelength changes to reflect the constraint imposed by equation (3.58),
Chapter 4  Partial Segmentation of Magnets and Rotor Yoke

\[ J_s(x) = \sum_{n=1}^{S} \left| \sin \left( \frac{\mu x}{P} \right) \right| \left( \frac{n^p}{(n-1)^p} \right) - \int \left( \frac{n^p}{(n-1)^p} \right) \sin \left( \frac{\mu x}{P} \right) dx \]  

The model produces a normalized power loss curve as \( S \) becomes a higher multiple of the segment's harmonic order.

\[ P_s(x) = \frac{J_s^2(x)}{n_s^p} \]  

The plot in Fig. 4-15 shows the normalized solid loss with increasing degree of segmentation. The Nyquist critical segmentation pitch is shown, and at this point the loss drops by 81%. It is interesting to note that after the Nyquist point, the law of diminishing returns applies to higher levels of segmentation, with only a further drop of 15% recorded for the next 8 levels of segmentation. There is an anomalous-looking value at \( S=2 \), and this error is addressed in the next section, where the model is extended to include time.

**Fig. 4-15:** The generic plot of normalized solid loss with relative segmentation degree.

The model in 4.5.2 works for harmonics that don't move relative to the segmentation pitch. This is not that useful in a machine rotor where, by definition the eddy currents are being produced by relative movement of field harmonics. These eddy currents are constantly moving relative to the conduction medium and, hence the segments. In order to demonstrate this, one must examine Fig. 4-16, where two current density waveforms are displayed segmented and
unsegmented. The second waveform is identical in wavelength, but is phase shifted compared to the first waveform:

\[ J(x,t) = J_1 + J_2 = J_1 \sin \left( \frac{\mu x}{\tau_p} \right) + J_2 \sin \left( \frac{\mu x}{\tau_p} + \frac{\pi}{2} \right) \]  

(4.15)

Note that in Fig. 4-15, when \( J_2 \) is segmented according to the Nyquist Critical Segmentation Pitch, the waveform is unaffected. At this time instant, the average current density value in each segment is zero, so the segmentation has no effect on the power loss. This prompts the model to be updated to include all possible phase shifts. Equation (4.16) incorporates these shifts into the relative segmentation model

\[ J_s(x) = \sum_{n=1}^{S} \sin \left( \frac{\mu x}{P} + \omega t \right) \left( \int \frac{n^P}{S} - \int \frac{n^P}{(n-1)S} \right) \sin \left( \frac{\mu x}{P} + \omega t \right) dx dt \]  

(4.16)

Implementation of Equation (4.16) also prompts an update to Fig. 4-14, correcting for the erroneous anomaly at \( S=2 \). This updated graph is shown in Fig. 4-16.

![Diagram showing current density waveforms](image)

**Fig. 4-16:** Two sets of current density waveforms flowing in a solid conductor medium are shown. The lefthand set is unsegmented, while the right hand set is segmented.

### 4.6.3 Optimal Segmentation in Design

For the purpose of machine design, Fig. 4-17 can be used to determine the level of segmentation needed to achieve the required efficiency gain for the least cost. In order to do this, one must be
critically aware of which harmonics are principally responsible for the eddy current loss. The rule of thumb using the Nyquist Critical Segmentation Pitch is that one must take the highest order damaging harmonic and choose a relative segmentation degree that is at least twice this wavelength. E.g.: If the highest damaging harmonic is the 28\(^{th}\) harmonic, then the rotor yoke should be designed with at least 56 segments, to reduce the loss by approximately 45\% for the 28\(^{th}\) harmonic. If there are 40 magnets, then they are naturally already ‘segmented’ to a harmonic degree of at least 40 depending on the pitch because they are separate solids. One can segment the magnets into two pieces each giving a harmonic segmentation pitch of 80 which is well above the nyquist value of 56. Lower harmonic orders will experience more severe loss reduction as the relative segmentation pitch will be higher for a lower order harmonic. If higher levels of loss reduction are required, then one must look at a higher segmentation level.

![Graph](https://example.com/graph.png)

**Fig. 4-17:** The relationship between power loss and segmentation for a given harmonic in time.
Chapter 5  Implementation of Analytical Rotor Loss Calculation in FEA

5  Implementation of Analytical Rotor Loss Calculation in FEA

5.1  Introduction

In machine design, it is important to be able to compute results that are accurate and quickly available, especially during optimization. The investigations into the magnet loss calculation yield some advantages and disadvantages of using analytical and finite element methods. A summary of these include:

<table>
<thead>
<tr>
<th>Finite Element Analysis</th>
<th>Analytical</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slow – caused by calculations that involve many fine time steps and with a fine mesh</td>
<td>Fast – due to calculation of few constants which are used in simple equations</td>
</tr>
<tr>
<td>Material saturation is taken into account using the material’s specific BH curve.</td>
<td>Material saturation is ignored and an average permeability value is used.</td>
</tr>
<tr>
<td>Versatile – many different machines can be modelled accurately, with variations in geometry, windings and anything that can be drawn.</td>
<td>Limited – the different windings need to be modelled and with minor variations in geometry, different Fourier coefficients need to be calculated.</td>
</tr>
<tr>
<td>Simple to use – FEA packages are designed to be simple to use and easily learnt.</td>
<td>Complicated – in order to use the analytical model, one needs to understand the theory and be involved in the programming.</td>
</tr>
</tbody>
</table>

Table 5-1: Advantages and disadvantages of using finite element and analytical calculation methods.

The method of used by finite element analysis is essentially a ‘blind’ method of calculation, as it takes a machine geometry, meshes it, computes the areas of excitation and produces results. It has no idea what it is modelling and is incapable of prediction. It is simply an accurate calculator.

Analytical methods are different in that they have intimate knowledge and understanding of what is being modelled, and the ability to predict what will happen in time. However, the calculations rely on an accurate excitation base with which to calibrate the model, which if calculated analytically produces results less accurate than the finite element model.

In order to harness the advantages of both methods, one needs to combine the accuracy and versatility of the finite element method with the speed and prediction knowledge of the analytical calculation. This chapter will focus on how this is to be achieved.

5.2  Model

The fundamental calculation in any FE software package includes determining the magnetic vector potential at each node in the mesh. Therefore, if a line of nodes is defined along the surface of the stator, then the FE package is able to produce a plot of the magnetic vector potential along the stator surface for a given current and winding. This result can be produced with a static calculation and is required at only one time step. These two factors greatly reduce
the calculation time of the FE model, while still producing a result that is accurate and very useful for the next stage in calculation.

\[
h_{FEA(\mu)} = \frac{1}{2\pi} \int h_{FEAx}(x) e^{-inx} dx
\]  

(5.1)

\[
H_{FEA(\mu)} = \sum_{\mu=1}^{\infty} h_{FEA(\mu)} \cos(\omega_s t - \pi \mu x_s),
\]  

(5.2)

The next step is for the analytical model to receive the data generated from the FE model. This is implemented by the stator line function being separated into its Fourier coefficients or harmonics.

There are three major types of harmonics that are present in the magnetic vector potential function, namely the stator winding harmonics, magnet field harmonics and slotting harmonics. It is important to identify each harmonic from the FE model, as this determines its frequency, its decay direction and each plays a certain role with respect to rotor losses. The harmonic definition column in Table 5-2 contains expressions that define the order of each harmonic source. By examining the harmonic profile of the airgap magnetic vector potential function, one can use this harmonic definition column to separate the harmonics based on their source.

The decay direction column is important and requires one to look at the governing differential equation for the magnetic vector potential in free space,

\[
\nabla^2 A - j \omega \mu A = 0
\]  

(5.3)

The general solution of this equation is:

\[
A(x, y) = (C_y e^{ir y} + D_y e^{-ir y}) e^{-ax}
\]  

(5.4)

<table>
<thead>
<tr>
<th>Harmonic source</th>
<th>Harmonic Definition</th>
<th>Harmonic speed relative to rotor</th>
<th>Decay direction</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator winding</td>
<td>( \mu = n \cdot \text{scm}(N_p, N_s) ) ( n = 1, 5, 7, 11, 13 \ldots )</td>
<td>( \omega = \frac{2\pi n L}{60} \left( \frac{N_p}{2} - \mu \right) )</td>
<td>stator surface toward rotor</td>
</tr>
<tr>
<td>Magnet</td>
<td>( \mu = \frac{n \cdot N_p}{2} ) ( n = 1, 3, 5 \ldots )</td>
<td>( \omega = 0 )</td>
<td>magnets toward stator</td>
</tr>
<tr>
<td>Slotting</td>
<td>( \mu = n N_s + \frac{i N_p}{2} ) ( n = 1, 2, 3 \ldots ) ( i = \pm 1, \pm 3, \pm 5 \ldots )</td>
<td>( \omega = \frac{i}{\text{abs}(i)} \frac{2\pi n L N_s R_s}{60} )</td>
<td>stator surface toward rotor</td>
</tr>
</tbody>
</table>

Table 5-2: The table depicts the origins of harmonics expected in the airgap along with their harmonic definitions, speeds and decaying directions, scm stands for smallest common multiple.

If one examines the expression in equation (5.4), depending on the equation constants, the magnetic vector potential can rise or fall exponentially with increasing \( y \) distance. In the case of MMF and slotting harmonics, which contribute to the rotor losses, they both decay with increasing distance from the stator. This explains why increasing the airgap causes a reduction
in rotor losses, as harmful magnetic field harmonics become diminished. The field harmonics caused by the magnet predictably decay as one moves from the rotor to the stator.

In Fig. 5-1, the airgap magnetic vector potential harmonics are seen for a 40 pole, 48 slot single layer, non-overlapping winding machine. Each harmonic is labelled by its origin. There are higher order harmonics; however these are too small to be viewed graphically.

The process of calculating the loss due to these harmonics begins with the identification of each harmonic and its magnitude. As an example, one can begin by looking at the subharmonic, which in this case has the harmonic order of 4. This harmonic order moves against the direction of the rotor, hence it is negative. This harmonic is caused by the stator current, therefore we shall identify its relative frequency to the rotor using stator winding equation from Table 5-2, where \( N_p = 40 \), \( n_s = 150 \) rpm and \( \mu = -4 \).

\[
\omega = \frac{2\pi \times 150}{60} \left(\frac{40}{2} + 4\right) = 376 \text{ rad/s} \quad (5.5)
\]

![Fig. 5-1: Airgap magnetic field harmonics in a 40 pole, 48 slot machine labelled with their source.](image)

The same process is followed for each harmonic to identify them and calculate their relative frequencies. The line magnetic vector potential function on the stator surface (can also be other interfaces) is used to calibrate the model defined in chapter three. This information is passed to the analytical model which ends the involvement of the FEM in the loss calculations.

Usually, when modelling a machine in FEA, only one part of the machine is modelled, and then using boundary conditions with positive or negative periodicity, the program multiplies the result accordingly to scale the results. The analytical model receives the information from the FEM that is not scaled, so it must first condition the information according the periodicity of the
machine, and the multiplication factor. Failure to do this can result in miscalculation of the order of the space harmonics leading to inaccurate prediction.

Combined with the rotor dimensions, number of poles, slots and rotor speed, the analytical model then uses all the theory outlined in chapter 3 to predict the magnetic vector potential throughout the airgap, magnets and rotor yoke. It uses the relative frequencies of each harmonic to generate the current density and power loss functions for each harmonic.

This is a powerful calculation technique as it essentially uses a magnetostatic FE solution to produce a transient result. This time savings for a computation such as this makes optimization for rotor loss in a machine very feasible.
6 Results and Comparison

The previous chapters have outlined the theory behind the contributions put forward by this work. Graphical evidence and results have been purposely omitted in favour of including them here. The chapter will first focus on establishing agreement between analytical and finite element analysis methods in magnetic vector potential, harmonic frequency and eddy current density in full and no load scenarios. It will then describe the results produced by the different rotors with specific focus on single and double-sided partial magnet segmentation and partial rotor segmentation.

6.1 Test Machine

The machine used to investigate the various loss reduction methods presented in this work is an outer rotor, 150rpm, 15kW PM generator designed for a direct drive wind generator application.

![Figure 6-1: The machine used for testing is seen here on the test bench.](image)

The test machine is used in all the analysis in this chapter and was designed with manufacturability as a main design objective. This aim guided the machine design in the following ways:

- Single layer, non-overlapping windings were used as only half the number of coils is required when compared to double layer windings.
- An outer rotor was selected so that the wind turbine blades could simply be bolted onto the shell of the rotor.
- Unequal teeth were selected, so that the coils sides could remain parallel, ensuring that the coils could be preformed and didn't have to be manually wound onto each tooth.
• A solid rotor yoke was used as it is simpler and cheaper than a laminated yoke which also requires a housing hub.

• Open slots are used for the stator windings to allow the insertion of preformed coils.

Figure 6-2: The picture shows two single-layer, non-overlapping coils with the open slotted stator.

These design decisions may make the construction of the machine simpler, but in terms of rotor loss, they are design choices that aid eddy-current inducing field harmonics in the airgap for the following reasons:

• Single-Layer, non overlapping windings contain a large, asynchronous, subharmonic which penetrates deep into the rotor causing eddy currents.

• The solid rotor yoke is an electrically conductive medium for circulating eddy currents.

• Flux density pulsations causing eddy currents are induced by the open slots due to the large permeance variation seen by the rotor.

More machine information is contained in the table below:

<table>
<thead>
<tr>
<th>Machine Parameter</th>
<th>Dimension</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator Inner Diameter</td>
<td>247mm</td>
</tr>
<tr>
<td>Stator Outer diameter</td>
<td>311.5mm</td>
</tr>
<tr>
<td>Rotor outer Diameter</td>
<td>326.75mm</td>
</tr>
<tr>
<td>Stack Length</td>
<td>100mm</td>
</tr>
<tr>
<td>Rotor Yoke Thickness</td>
<td>7.25mm</td>
</tr>
<tr>
<td>Magnet Pitch</td>
<td>0.73%</td>
</tr>
<tr>
<td>Magnet thickness</td>
<td>6 mm</td>
</tr>
</tbody>
</table>
### Table 6-1: Machine Parameters for the machine used for analytical and FEM results analysis and testing

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Speed</td>
<td>150rpm</td>
</tr>
<tr>
<td>Airgap Length</td>
<td>2mm</td>
</tr>
<tr>
<td>Coil Width</td>
<td>18mm</td>
</tr>
<tr>
<td>Tooth Width</td>
<td>11.1mm</td>
</tr>
<tr>
<td>Current (RMS)</td>
<td>23.9A</td>
</tr>
<tr>
<td>Rated Power</td>
<td>15kW</td>
</tr>
<tr>
<td>Torque</td>
<td>1kN.m</td>
</tr>
<tr>
<td>No. of Turns</td>
<td>46</td>
</tr>
<tr>
<td>Slots</td>
<td>48</td>
</tr>
<tr>
<td>Poles</td>
<td>40</td>
</tr>
</tbody>
</table>

The machine was built and tested with 3 separate rotors, each differing by one design change in order to isolate the effect of each. The three rotors were:

- solid magnet solid yoke
- double-sided segmented magnet, solid yoke
- double-sided segmented magnet, laminated yoke

The analytical, FEM and measured results are presented later in the chapter. The work also discusses the method of partial rotor yoke segmentation and presents finite element analysis and analytical results.

### 6.2 Analytical and FEM Calculation Comparison

The first step in presenting the results is to cross-check the analytical and finite element methods to ascertain their level of agreement.

In the analytical calculations, one needs firstly to define the current sheet to represent the stator winding. It was decided to use the finite element method suggested in chapter 5 to define the stator surface current sheet for the following reasons:

- The analytical method effectively defines the winding numerically by laying out the winding slot by slot, which is a method very similar to FEA. However due to local saturation in the teeth of the stator, the FEA will produce a more accurate result.
- This study is not being undertaken to assess different winding methods. This means that one does not need to have an analytical method that calculates the current sheet based on what type of winding is used. Instead, a ‘worst case scenario’ winding is chosen that is known to contain damaging harmonics.
- By using the current sheet extracted from the FEM, one is also testing the methods of eddy current calculation and solid loss proposed in chapter 5.

The calculations can be divided up into full and no load calculations that are treated separately as they have different harmonic origins that cause rotor loss, but first the rotor harmonic speeds are calculated and verified.
6.2.1 Harmonic Speed Calculation

The role of the harmonic speed is significant in calculating the magnetic vector potential and the current density. From (3.26), one can see that it plays a role in the decay of harmonics, and also in determining the magnitude of the current density seen in equation (3.54). Each harmonic, unique in spatial frequency and time frequency, is treated with a separate calculation.

As the frequency calculation is so integral in the analytical calculation process, a comparison between calculated harmonic frequencies with FEA is made. To this end, a magnetic vector potential line function is extracted from a radius from the moving rotor reference frame at each FEA time step. A spatial fourier transform is then performed to isolate each space harmonic for independent monitoring. The time that each spatial harmonic takes to complete a full cycle in the complex plane, determines the time period of that harmonic, allowing one to calculate the frequency. Analytically, the frequency of a particular harmonic is also calculated using equation (3.7) for comparison. The subsection will look at two harmonics extracted from the FEM and investigate their velocity compared with the analytical model.

![Graphs showing harmonic speed calculations](image)

**Fig 6-3:** The graphs show harmonics extracted from the FEA of spatial order 20 (working harmonic) (top left) and 28 (top right), and plotted against each other (bottom), all shown in the complex plane.
The graphs shown in Fig. 6-3 show two harmonics in the complex plane, one harmonic is of order 28 and the other is the working harmonic of order 20. Each harmonic is shown zoomed at the top of the figure, and at the bottom the two harmonics are shown plotted on the same figure. An initial observation (from Fig. 6-3 bottom) is that the 28\textsuperscript{th} harmonic rotates all around the complex plane, hence it moves relative to the rotor reference frame, and the working harmonic stays stationary (synchronous to the rotor). The working harmonic can be seen to be jagged and non-uniform when zoomed in (Fig. 6-3 top left), but when plotted against the 28\textsuperscript{th} harmonic (Fig. 6-3, bottom), one can see that the non-uniformities are relatively small.

Note in Fig. 6-3, the relative size difference between the working harmonic and the asynchronous 28\textsuperscript{th} harmonic.

The FEA program used a time step of 0.000333s and it took 25 time steps for the 28\textsuperscript{th} harmonic to return to its original position in the complex plane which gives the frequency:

\[
\omega_{28,FEM} = 2\pi f = \frac{2\pi}{P} = \frac{2\pi}{0.000333 \times 25} = 754.73 \text{rad/s}^{-1}
\]  

(6.1)

Using equation (3.7), the harmonic speeds are calculated analytically

\[
\omega_{28,\text{Analytical}} = \frac{2\pi n_s}{60} \left( \frac{N_p}{2} - \mu \right) = \frac{2\pi \times 150}{60} (20 - (-28)) = 753.98 \text{rad/s}
\]  

(6.2)

The agreement between the two harmonic speeds is good; the small error could be due to numerical truncation in the finite element analysis. Looking at the working harmonic, one can see from the graph that we expect a zero result for the relative speed:

\[
\omega_{20,\text{Analytical}} = \frac{2\pi n_s}{60} \left( \frac{N_p}{2} - \mu \right) = \frac{2\pi \times 150}{60} (20 - (20)) = 0 \text{rad/s}
\]  

(6.3)

The variation in the working harmonic in Fig. 6-3 (top left) is likely from small solution variations due to changes in node alignment as the rotor moves into different positions relative to each other in time and the effect of slotting.

### 6.2.2 Full Load

The full load calculation is aimed at calculating the magnetic fields due to the stator current winding only, for this reason, the interference of the fields due to the magnets needs to be removed. To achieve this, the magnetization of the magnet is set to zero for both the analytical methods and FEA. The conductivity and permeability properties of the magnets are retained.

#### 6.2.2.1 Magnetic Vector Potential Ignoring Reaction Field

For the first comparison in magnetic vector potential, the reaction fields are ignored for simplicity, this is achieved by setting the conductivities of the magnet and rotor regions in equation (3.27) to zero. The analytical model makes a prediction of the magnetic field in 2-D
from the stator surface throughout the airgap, magnets and rotor. In order to assess the accuracy of the analytical model and its agreement with the FEA, the model is compared at two circumferences. Fig. 6-5 shows a spatial distribution of magnetic vector potential at a fixed time instant on the two lines shown in green and red in Fig. 6-4.

![Diagram of rotor and stator components](image)

**Fig. 6-4:** This figure shows the two circumferences from which the rotor fields are extracted for comparison.

A good agreement is shown between both of the pairs of lines in Fig. 6-5. The error is quantified as 1.01% for both comparisons which can be mostly attributed to the high order spatial harmonics which are ignored for simplicity as they have a negligible contribution to the rotor loss.

![Graph of magnetic vector potential](image)

**Fig. 6-5:** The figure shows a magnetic field comparison between the FE and the analytical prediction at two circumferences on the rotor.

It should be noted at this point that a zero magnetic vector potential boundary condition was stipulated for the outer rotor yoke surface.
6.2.2.2 Magnetic Vector Potential including Reaction Field

Using a similar comparison method to 6.2.2.1, the magnetic vector potential is calculated throughout the airgap and rotor region, except in this case, the calculation is also inclusive of reaction fields due to eddy currents. This is achieved by setting the relative conductivities of the steel and magnets in the yoke as in equation (3.27). Magnetic vector potential line functions are extracted from two lines in the FEA shown in Fig. 6-4 and compared with equivalent analytically calculated results.

The magnetic field prediction in Fig. 6-6 is due to the presence of a current loading on the stator surface. Equations (3.23) to (3.52) are used to define the general solution for the magnetic field. The field prediction in the magnet (Region 2) shows a very good agreement with the finite element solution, which confirms the accuracy of the analytical model. To a lesser extent, an agreement is reached in Fig. 6-7 for the prediction in the yoke (Region 3), however due to the non-uniformity of the permeance of the steel in the rotor yoke, the agreement is not as close as in region 2. The effect of the more significant permeability and conductivity is evident in Fig. 6-7, where large changes to the waveform occur from Fig. 6-6. This strong effect is due to the dominance of the second term in equation (3.27).

![Graph of Magnetic Vector Potential](image)

**Fig. 6-6:** The comparison between analytical and FE calculation of the magnetic vector potential function in the centre of the magnet.

When examining the spatial magnetic vector potential functions, here are some important comparative differences to note:

When eddy currents are considered:
• there exists a magnitude difference and phase angle shift of the harmonics of the spatial magnetic vector potential function between the magnet and the rotor (compare Fig. 6-5, 6-6 and 6-7).
• the difference in magnetic vector potential in the magnets between when eddy currents are, and are not considered, is less than the equivalent scenario in the solid rotor yoke.
• in the case of the analytical solution, the accuracy of the rotor solution becomes reduced compared to when reaction fields are ignored.

![Graph showing comparison between analytical and FE calculation of the magnetic vector potential in the centre of the rotor.](image)

**Fig. 6-7:** Comparison between analytical and FE calculation of the magnetic vector potential in the centre of the rotor.

The reasons for these differences can be explained with equation (3.27). Using the subharmonic (no. 4 in this case), we use the calculation of $\gamma$ as a worked example:

*In the magnet,*

$$f or \ \gamma = 0, \quad \gamma^2 = a^2 = (4 \times \pi)^2 = 157.9 \quad (6.4)$$

$$f or \ \gamma = 6.25 \times 10^5, \quad \gamma^2 = a^2 + j\omega k\mu$$

$$= (4 \times \pi)^2 + j \times 377 \times 6.25 \times 10^5 \times 4\pi \times 10^{-7}$$

$$= 157.9 + 296j \quad (6.5)$$

*In the rotor,*

$$f or \ \gamma = 0, \quad \gamma^2 = a^2 = (4 \times \pi)^2 = 157.9 \quad (6.6)$$
The variable \( \gamma \) determines the magnitude and phase of the decaying harmonic and is the principle determining factor for how the magnetic vector potential changes when eddy current reaction fields are considered. The two main components of \( \gamma \) can be divided into magnitude and phase; the following table considers the changes to both.

\[
\begin{align*}
\text{No eddy currents present} & \quad \text{Eddy currents present} \\
\hline
\text{Phase} & \text{Magnitude} & \text{Phase} & \text{Magnitude} \\
\text{Magnets} & 0 & 157.9 & 61.92^\circ & 335.5 \\
\text{Rotor} & 0 & 157.9 & 89.99^\circ & 995000 \\
\end{align*}
\]

Table 6-2: The table considers the changes to the subharmonic rate of decay \( \gamma \) as one investigates the effect of the eddy current reaction field.

The effect of the reaction fields on the decay rate, \( \gamma \) on the subharmonic in both phase and magnitude can be seen in Table 6-2. The magnet experiences a phase shift of 61.92\(^\circ\), while the magnitude increases by 2.1 times. The rotor experiences a phase shift of nearly 90\(^\circ\) and the magnitude goes up by 6300 times. It must be noted that each harmonic has a different value of \( \gamma \) and therefore a different magnitude and phase change can be expected for each.

The magnet's reaction field has a smaller effect than the rotor's reaction field due to the lower values of permeability and conductivity. In the case of the rotor, the values of permeability and conductivity dictate that the harmonic order is mostly irrelevant compared to the frequency. Evidence of this can be seen in equation (6.7) where the real term which is calculated from the harmonic order is insignificant compared to the complex term which is calculated from the time frequency. For this reason, the skin depth dominates the decay of the reaction field.

There is a degree of disparity between the FEA results and the analytical results in the solid rotor yoke in Fig 6-7. The reason for this is that in the analytical calculations, the permeability is assumed to be a constant value throughout the rotor. This assumption can lead to inaccuracies in the analytical calculations as the permeability varies with magnetic field strength. The FEA avoids this error by calculating the flux density at each node by method of interpolation from a B-H curve lookup table, effectively creating a non-linear plane of permeability within the rotor.

\[
\begin{align*}
\text{for } \kappa &= 6.0 \times 10^6, \quad \gamma^2 = \alpha^2 + j \omega \kappa \mu \\
&= (4 \times \pi)^2 + j \times 377 \times 6.0 \times 10^6 \times 4 \pi \times 10^{-7} \times 350 \\
&= 157.9 + 9.95 \times 10^5 j
\end{align*}
\]
6.2.2.3 Eddy Current Density

The eddy current calculation is the penultimate step to calculating solid loss, and is determined using equation (3.54). In order to compare the FE eddy current functions in the magnet, the magnet pitch is set to 1 to avoid dead conduction spots in the gaps between the magnets. This is purely for comparison purposes, ensuring that the eddy current functions are accurate. The calculations for yoke loss are also computed with equation (3.63), which doesn't rely on the current density function.

![Diagram of machine model with 100% magnet pitch](image)

**Fig. 6-8:** The figure shows the altered machine model with 100% magnet pitch and the blue and red lines depicting where the comparison lines are located.

![Current density graph](image)

**Fig. 6-9:** The current density in the middle of the magnet, predicted by FE and Analytical methods.

The results in Fig. 6-9 show good agreement between the analytical and the FE prediction, although there is a slightly higher absolute prediction from the analytical calculation. An interesting consideration when looking at this small discrepancy between the FE and analytical results in Fig. 6-9, is to look at the calculation technique of the finite element program. Direct calculations such as magnetic vector potential, torque and flux are where FEA is at its most
accurate, as these values are direct results of the mesh, geometry and excitation. However, when time derivatives are involved, values are subjected to division by time steps which, depending on their size, can sometimes lead to inaccuracies. Current density is one such function, so it is possible that in this case, the analytical method is more accurate than the FEM.

The rotor current density plot is heavily dependent on the solid yoke steel’s conductivity and permeability which contribute to the phase and magnitude of the waveform. Fig. 6-10 shows the two predictions of the analytical and FE models. The disparity between the calculations can be mostly ascribed to the non uniform permeability of the rotor yoke.

![Current density comparison graph](image1)

**Fig. 6-10**: Current density comparison of analytical and FEA methods in the rotor centre.

### 6.2.3 No Load

The no-load calculation is a separate calculation to the full load calculation as it uses a different set of harmonics, of different origin moving at different speeds. The no load loss arises due to the slots of the stator causing a permeance variation ‘seen’ by the rotor’s magnetic field. In order for this scenario to be simulated in the FEM, the magnetization of the magnets is set and the current in the windings is turned off.

#### 6.2.3.1 Harmonic Speed Calculation

In order to calculate the magnetic vector potential and the eddy currents induced due to slotting field harmonics, one needs to identify the harmonics and their relative speeds. The speed and definition of each harmonic is governed by:

\[
\omega = \frac{i}{abs(i)} \frac{2\pi n N_s R_s}{60}
\]  

(6.8)
where $\mu = nN_s + i\frac{N_p}{2}$, $n=1,2,3..., \ i=\pm 1,3,5...$

As worked examples, consider two cases:

Case 1: let $n=1, i=-1, N_s=48, N_p=20.$

$$\mu = 1 \times 48 - i \frac{40}{2} = 28$$

$$\omega = \frac{i}{\text{abs}(i)} \frac{2\pi n N_s R_s}{60} = \frac{-1}{1} \times \frac{2\pi \times 1 \times 48 \times 150}{60} = -753.98 \text{ rad.s}^{-1}$$  \hspace{1cm} (6.9)  \hspace{1cm} (6.10)

Case 2: let $n=2, i=1, N_s=48, N_p=20$

$$\mu = 2 \times 48 + i \frac{40}{2} = 116$$

$$\omega = \frac{i}{\text{abs}(i)} \frac{2\pi n N_s R_s}{60} = \frac{1}{1} \times \frac{2\pi \times 2 \times 48 \times 150}{60} = 1507.96 \text{ rad.s}^{-1}$$  \hspace{1cm} (6.11)  \hspace{1cm} (6.12)

The two harmonics described in the case study are shown in Fig. 6-11. These two graphs represent the two harmonics over the same time frame, i.e.: in the time that the 28th harmonic completed one cycle, the 116th harmonic completed two cycles. This confirms the result that the 116th harmonic moves at twice the frequency as the 28th harmonic. The difference in size between the two harmonics demonstrates how the magnitude decreases with harmonic number. The exact FEA calculation of the period of the waveform follows:

$$\omega_{28,FEM} = 2\pi f = \frac{2\pi}{\frac{P}{3.3 \times 10^{-5} \times 250}} = -754.73 \text{ rad.s}^{-1}$$  \hspace{1cm} (6.13)

$$\omega_{116,FEM} = 2\pi f = \frac{2\pi}{\frac{P}{3.3 \times 10^{-5} \times 125}} = 1509.46 \text{ rad.s}^{-1}$$  \hspace{1cm} (6.14)

**Fig. 6-11:** The two graphs show the two case study harmonics extracted from FEA, 28 (left) and 116 (right) in the complex plane.
An observation here is the existence of an overlap in the no load and full load harmonics. There is a 28\textsuperscript{th} harmonic due to the current winding and there is also a 28\textsuperscript{th} harmonic due to the slotting. It is interesting to note that both of these harmonics move at the same frequency, but in different directions.

### 6.2.3.2 Magnetic Vector Potential

The rotor field induces two groups of magnetic vector potential harmonics; those that are induced by the field itself and those that are induced by the interaction of the rotor field with the permeance variation of the stator slots. The magnetic vector potential function extracted from the FEM on the surface of the rotor is used to define these harmonics, but it is important to distinguish between the two sets of harmonics as only the slotting harmonics rotate asynchronously to the rotor. In the software analysis, the static rotor field harmonics are set to zero.

Once again, the magnetic vector potential is verified using two line functions in the machine rotor shown in Fig. 6-4. Again, as in chapter 6.2.2.2, the difference in accuracy between the magnet plots and the plots in the rotor yoke can be observed. As the identical method is used to calculate both of these results, this behaviour can only be explained by the variance in permeability due to steel saturation.

![Graph of Magnetic Vector Potential](image)

**Fig. 6-12:** The comparison between analytical and FEM calculation of the magnetic vector potential function in the centre of the magnet.
Fig. 6-13: The comparison between analytical and FEM calculation of the magnetic vector potential function in the centre of the rotor yoke.

6.2.3.3 Eddy Current Density

The comparisons between FEM and analytical calculations of eddy current density in the centre of the magnet and yoke (from Fig. 6-8) are shown below in Figs. 6-14 and 6-15.

Fig. 6-14: Comparison of FEM and analytical calculations for eddy current density in the centre of the magnet.
From equation (3.54), it is clear that any harmonics that rotate in synchronism with the rotor will not contribute to the eddy current losses. Therefore, the static rotor harmonics are not present in this plot. The two functions in Fig 6-14 and 6-15 are found to be in good agreement, which bodes well for the accuracy of the final solid loss calculation.

6.3 Measurement Procedure and Determination of Constants

The measurement set up consisted of a drive connected to an induction motor, which is connected to the machine through a 15:1 reduction gearbox. The speed of the machine can be assumed to be 150rpm unless otherwise stated and the full load current is 33.8A, resulting in a machine a rating of 15kW.

The no load test was designed to compare the losses in the three rotors, and also use the input power measurements and rotor temperature measurements for three rotors to determine the rotor’s thermal constants, so that temperature results can be used to estimate rotor loss in other experiments.

6.3.1 Torque and Speed Measurement

The torque and speed were measured on the machine shaft from zero to base speed. Due to zero current in the windings, the only competing loss was in the form of eddy currents in the coils which were assumed to be negligible as the coils measured no temperature rise during the various tests, and stator core loss.
The power loss in the rotor due to slotting is calculated as:

\[ P_{NL} = \frac{n_z \cdot \text{Torque}_{\text{measured}} \cdot 2 \cdot \pi}{60} - P_{\text{Loss}} \]  \hspace{1cm} (6.15)

where

\[ P_{\text{Loss}} = P_{\text{wf}} + P_{\text{eddy}} + P_{\text{Stator\_core}} \]  \hspace{1cm} (6.16)

The \( P_{\text{Loss}} \) component is made up of a variety of parasitic losses in the machine including the wind and friction loss, the eddy currents in the windings and the stator core losses. These losses are assumed to be approximately constant for the three rotors during the no-load tests as the conditions were the same and the behaviour of three rotors would have been seen to be identical from the stator side.

### 6.3.2 Temperature Measurement

The thermal loss manifests itself as a material temperature rise on the rotor yoke, so a model was sought in order to accurately map the relationship between the rotor loss and temperature rise. This model is shown in Fig. 6-16. The test conducted was one where the machine was run under no load or full load conditions for a period of time until the temperature stabilised, where a final temperature reading was taken.

![Fig. 6-16: The thermal model used to link measured temperatures on the rotor with power dissipated in the solid material.](http://scholar.sun.ac.za)

Using Newton’s Law of Cooling for the two rotor surfaces, we have:

\[ P_{\text{rotor\_loss}} = \frac{dQ}{dt} \approx h_1 A_1 (T_{\text{measured\_OuterRotor}} - T_{\text{amb\_Outside}}) \]

\[ + h_2 A_2 (T_{\text{measured\_InnerRotor}} - T_{\text{amb\_Stator}}) \]  \hspace{1cm} (6.17)

where, \( h_n \) is the heat transfer coefficient, \( A_n \) is the surface area, \( T_{\text{amb}} \) is the ambient temperature, \( T_{\text{measured}} \) is the measured rotor temperature and \( Q \) is energy.

It should be noted at this point that certain assumptions can be made to simplify the thermal model. This model only considers heat transfer through convection, so radiation is ignored, also the heat transfer through wind in the airgap is also ignored. The ambient outside temperature \( T_{\text{amb\_Outside}} \) was found to be very similar to \( T_{\text{amb\_Stator}} \) and so they can be assumed to be equal.
Also, the measured temperatures on the inside and outside of the rotor yoke were also found to be very much similar, so these terms are also approximated to be equal, viz:

\[ T_{\text{measured, Outer Rotor}} \approx T_{\text{measured, Inner Rotor}} \quad (6.18) \]
\[ T_{\text{amb, Outside}} \approx T_{\text{amb, Stator}} \quad (6.19) \]

These assumptions from (6.17) give:

\[ P_{\text{rotor loss}} = \frac{dQ}{dt} \approx (h_1A_1 + h_2A_2)(T_{\text{measured, Outer Rotor}} - T_{\text{amb, Outside}}) \quad (6.20) \]

and one can define new constants, \( h \) and \( A \) to rewrite equation (6.20) as:

\[ P_{\text{rotor loss}} = \frac{dQ}{dt} \approx hA(T_{\text{measured, Outer Rotor}} - T_{\text{amb, Outside}}) \quad (6.21) \]

where \( hA = h_1A_1 + h_2A_2 \)

This is an important relationship as it tells us that in steady state, the temperature rise of a surface is directly proportional to the amount of power it dissipates as heat. For this reason, tests were conducted over significant time periods in order to allow the temperature to stabilize. From this point \( T_{\text{measured, Outer Rotor}} \) and \( T_{\text{amb, Outside}} \) will be referred to as \( T_{\text{measured}} \) and \( T_{\text{amb}} \).

### 6.3.3 Determination of Constants

The power into the rotor through the shaft is denoted as \( P_{\text{in}} \), and in no load, this power is dissipated as wind and friction losses, stator core losses and rotor loss. For the purposes of this investigation, the stator core loss will be lumped with the wind and friction losses as it is constant for the three rotors in the no load case for a fixed speed.

\[ P_{\text{in}} = \text{Torque} \times \omega = P_{\text{rotor loss}} + P_{\text{Loss}} = hA(T_{\text{measured}} - T_{\text{amb}}) + P_{\text{Loss}} \quad (6.22) \]

The torque and speed results are measured on the input shaft, and \( T_{\text{measured}} \) and \( T_{\text{ambient}} \) are both measured with a temperature sensor, this leaves just two unknowns, \( hA \) and \( P_{\text{Loss}} \). One can assume that the heat transfer coefficient and surface areas are constant throughout the three rotors, and also the wind and friction losses, eddy losses and the stator core losses. So, using the power results measured from the three rotors, and using the measured temperatures from the rotor surface and the ambient temperature, we can formulate the simultaneous equations to solve for \( hA \) and \( P_{\text{Loss}} \):

\[ hA(60.4 - 19.2) + P_{\text{Loss}} = 41.2hA + P_{\text{Loss}} = 807 \quad (6.23) \]
\[ hA(49.6 - 19.2) + P_{\text{Loss}} = 30.4hA + P_{\text{Loss}} = 607 \quad (6.24) \]
\[ hA(33.0 - 19.2) + P_{\text{Loss}} = 13.8hA + P_{\text{Loss}} = 337 \quad (6.25) \]

Solving for \( hA \) and \( P_{\text{Loss}} \) we get average values of 17.8 and 73W for the two variables respectively. These constants are used to calculate the measured results published in the remainder of this results chapter.
When attempting to measure the rotor losses at full load, the technique of measuring the output electrical power compared with the input mechanical power is difficult as there are additional losses in the stator core and the copper wiring. The copper losses can be accurately estimated, but the core loss in the stator is more difficult to determine. For this reason, this temperature technique is used to estimate rotor loss as it maps the relationship between the power lost in the rotor and the measured rotor temperature using the following formula:

\[ P_{\text{rotor loss}} = hA(T_{\text{measured}} - T_{\text{amb}}) - P_{\text{Loss}} \]  

(6.26)

where the constants \( hA \) and \( P_{\text{Loss}} \) are calculated from the three no-load cases. The measured temperature is taken from the outer surface of the rotor yoke once the temperature has stabilized. Various assumptions are made here which should be noted. The heat transfer model is simplified with just one measured temperature point on the outer surface of the rotor yoke. The model assumes that the temperature throughout the rotor yoke and magnets is constant and that the ambient room/atmosphere temperature is approximately equal to the stator surface temperature.

In design, the aim of determining the constants is so that one can use rotor power loss results from FEA with the thermal constants of the machine to make an accurate prediction of the rotor temperature. This allows one to design a rotor according to material temperature limits, with safety factors to ensure the longevity of the machine and its performance. Additionally, one should note that when the machine is installed in a wind generator, the air moving past the machine can act to improve the heat transfer coefficient ensuring that the rotor performs at an even lower temperature than what would be predicted in a laboratory.

### 6.4 Nyquist Critical Segmentation Pitch

Two groups of analytically calculated harmonics are presented to demonstrate the Nyquist Critical Segmentation Pitch. These two groups comprise the current winding harmonics and the slotting harmonics in the machine. For the investigation, the current density across the mid-point of the magnet is examined. The magnet is left unsegmented, and the pitch is made such that each magnet is touching the adjacent magnet to eliminate any dead conduction spots. The model covers 3 adjacent conducting magnets, each conducting region forced to comply with the current balancing constraint in equation (3.57).

The various harmonics present in the two plots of Fig. 6-18 and Fig. 6-19 are shown in the Fig. 6-17. This plot shows in red the harmonics that are present in Fig. 6-18 and in blue the harmonics that are present in Fig. 6-19. The Nyquist Critical Segmentation Frequency is shown in black.

In Fig. 6-18, the harmonics due to slotting are considered. The harmonic profile of the eddy currents due to slotting are varied, but the main contributors are the 28th and 12th harmonics.
In Fig. 6-19, the harmonics due to the current winding are considered. The harmonic profile of the eddy currents due to the current winding is less varied, but the $4^\text{th}$ harmonic is the major contributor with the $28^\text{th}$ harmonic being secondary.

**Fig. 6-17:** The figure shows the current density magnitudes of the various harmonic orders, colour coded to reflect the origin of each harmonic.

**Fig. 6-18:** The current density plot in the centre of the magnet containing slotting harmonics only. The black lines show the limits of each magnet.
Fig. 6-19: The current density plot in the centre of the magnet containing current winding harmonics only. The black lines show the limits of each magnet.

There are 40 magnets in the machine each with the current constraint of equation (3.57). Using the Nyquist Critical Segmentation Pitch, this means that only eddy current harmonics of order 20 and below are going to experience significant loss reduction. Note that there is very little change between the current density before and after the current balancing constraint is enforced in the no-load case in Fig. 6-17. The reason for this is that the primary contributor is the 28\textsuperscript{th} harmonic, which is almost unaffected by the segmentation pitch which effectively has a harmonic order of 20.

In the case of the winding current harmonics, Fig. 6-17 clearly shows the presence of a dominant 4\textsuperscript{th} harmonic. This harmonic is much larger in wavelength than the segmentation pitch, thereby having a high relative segmentation degree. According to Fig. 4-16, a relatively large change to the current density post segmentation is expected. This significant change in the current density waveform is reflected in Fig. 6-19.

6.5 Solid Magnets and Solid Yoke Rotor

A machine with parameters listed in chapter 6.1 was built and tested with a rotor with a solid yoke and solid magnets. This machine induces the most rotor losses of all the machines tested and was used as a ‘worst case scenario’ machine. The solid rotor and solid magnets both produced excessive heat preventing the machine being run at full load, for fear that the magnets would become damaged due to overheating. The rotor reached a temperature of 98\degree at 14kW.
Fig. 6-20: A close up view of the solid, surface mounted magnets and the solid yoke rotor.

In tables 6-3 and 6-4, the no load and full load losses are presented as measured by analytical methods, FEA and physical measurements. The measured results are only listed in the ‘Total’ column as individual magnet and rotor loss components could not be measured separately. The machine was run in no-load and full load scenarios. The losses were measured as a difference between the mechanical power measured on the shaft of the generator and the output electrical power less the copper losses. The temperature is measured on the rotor yoke outer surface.

6.5.1 Rotor Losses due to Slotting

The calculation of no load losses is presented with the three calculation methods of analytical, FEM and measured all shown in Table 6-3. The measured rotor loss results are calculated using the wind and friction estimation from section 6.3.3 and the measured input mechanical power. Individual physical measurements of magnet and rotor yoke loss were not possible, and so only the summed value is included in the table.

<table>
<thead>
<tr>
<th>Method</th>
<th>Magnet Loss (W)</th>
<th>Rotor Yoke Loss (W)</th>
<th>Total</th>
</tr>
</thead>
<tbody>
<tr>
<td>Analytical</td>
<td>202</td>
<td>515</td>
<td>717</td>
</tr>
<tr>
<td>Finite Element</td>
<td>220</td>
<td>535</td>
<td>755</td>
</tr>
<tr>
<td>Measured</td>
<td></td>
<td></td>
<td>733</td>
</tr>
</tbody>
</table>

Table 6-3: Comparison of three methods of calculating solid losses induced in the magnets and rotor yoke at no load, 150rpm.

The no load loss results due to slotting show good agreement throughout as shown in Table 6-3. This is a test that has the least additional forms of loss making it the most pure rotor loss test.
The rotor experienced a temperature rise from 19.2°C to 60°C, which is a significant increase considering that no current was flowing in the windings.

### 6.5.2 Rotor Losses due to Stator Current

The measurement method used for isolating the losses due to stator current in this subsection is based on the theory in section 6.3.3. The rotor was run at full load and temperature readings were taken on the rotor surface. Using the area, heat transfer and wind and friction loss constants generated in the no load tests, the steady state temperatures were used to approximate the loss that was induced in the rotor. One assumption to be taken into account was that the $P_{\text{loss}}$ parameter might not be accurate under full load conditions.

![Table 6-4: Solid losses induced in the magnets and rotor due to the current winding harmonics only.](image)

<table>
<thead>
<tr>
<th>Method</th>
<th>Magnet Solid Loss (W)</th>
<th>Rotor Yoke Solid Loss (W)</th>
<th>Total</th>
</tr>
</thead>
<tbody>
<tr>
<td>Analytical</td>
<td>55</td>
<td>557</td>
<td>615</td>
</tr>
<tr>
<td>Finite Element</td>
<td>42</td>
<td>520</td>
<td>562</td>
</tr>
<tr>
<td>Measured</td>
<td></td>
<td></td>
<td>645</td>
</tr>
</tbody>
</table>

The three methods of calculating the rotor loss due purely due to stator current is shown in Table 6-4. The magnet loss column shows that the losses induced in the magnets account for much less than those in the solid rotor yoke, which makes the yoke a higher priority for loss reduction manufacturing strategies. Note that the magnet loss would be much higher if they were a solid disk and not separate magnets, due to the segmentation effect that they experience being isolated conductors. The loss results returned in that case would be very similar to the yoke loss, but when one enforces the condition in equation (3.57) that the sum of the currents in each magnet must equal zero, the loss in the magnet reduces dramatically.

### 6.6 Partial Magnet Segmentation

#### 6.6.1 Single Sided Partial Magnet Segmentation

In testing, there was no physical machine built that included SS-PMS magnets, only a machine that implemented DS-PMS. For this reason, the results produced in this subchapter on SS-PMS are generated by the FEA and analytical methods only. Essentially, SS-PMS is a diluted version of DS-PMS, and unless there is a large cost differential or a structural concern between the two methods for a specific application, one would expect designers to opt for DS-PMS.

In SS-PMs, the power loss density in the magnet is expected to differ in the portion of the magnet that is segmented, compared to the unsegmented portion. A three dimensional, analytical model of the magnet in $1/5^{th}$ of the machine is shown in Fig. 6-21, comparing the power loss density in the segmented portion of the magnet with that of the unsegmented
portion. The magnet is partially segmented to 50% of its depth in the y direction, ie: 1mm, with 8 segments per magnet pitch. The effect of the segmentation is very clear in the segmented portion, where the power loss density is far lower than in the portion of the magnet that remains solid. It is clear from this diagram and Fig. 6-21 that increasing the penetration of the segment beyond 50% would be effective in further reducing the power loss.

![Fig. 6-21: A 3-D model of the normalized magnet loss density with the magnet x and y dimensions.](image)

![Fig. 6-22: The normalized power loss density vs. magnet depth for a SS-PMS magnet with 50% penetration.](image)

The graph in Fig. 6-22 shows the average normalized power loss density in the SS-PMS magnet for varying degrees of segmentation from 1 segment (no segmentation) through to 8 segments.
It is clear that increasing levels of segmentation greatly reduce the power loss in the segmented half of the magnet. Also note that without segmentation, the half of the magnet closest to the airgap (0-1mm) has a considerably higher power loss density than the other half (1-2mm).

![Graph of magnet loss](image)

**Fig. 6-23**: The overall magnet loss for the specific machine is plotted with increasing levels of SS-PMS, computed with analytical and FEM methods.

The magnet loss graph in Fig. 6-23, quantifies the overall magnet loss in the 15kW machine described in section 6.1. One can clearly see the expected reduction with increasing levels of segmentation. The graph tends toward an imaginary asymptotic line that represents the total loss in the unsegmented half of the magnet. From a design perspective, one can see that there is little additional benefit in segmenting to 8 pieces, as 4 segments achieves very similar results.

### 6.6.2 Double-Sided Partial Magnet Segmentation

One of the rotors used in the machine described in chapter 6.1, implements DS-PMS in its rotor magnets, so for this subsection a comparison of analytical, FEM and machine measured results can be made. The tests are all run in no-load as this eliminates other forms of loss, making the measured results more accurate.

In Fig. 6-24, one can see the front and back halves of a DSPMS magnet are segmented to different degrees so that their segmentation incisions are not aligned. The magnets are glued into place with epoxy glue. Care has to be taken while bringing a magnet toward the rotor, as it experiences strong magnetic forces from the steel yoke and the other magnets. For this reason an aluminium toothed comb was placed inside the rotor yoke to align the magnets and an
aluminium ‘spade’ was used to position the magnets. Clamps were used to keep the magnets in place until the epoxy had dried.

![Image](image1.png)

**Fig. 6-24:** The picture shows the double-sided partially segmented magnets which have been glued onto the rotor surface.

![Image](image2.png)

**Fig. 6-25:** A 3-D analytical model of the magnet loss vs. x and y dimensions.

A three dimensional depiction of normalized magnet loss vs. the x and y dimensions of the magnet is shown in Fig. 6-25. DS-PMS produces a loss density function which is very similar to that of FMS, due to their closely related segmentation profiles.

In the test machine’s case, a DS-PMS magnet with 4 segments at the face and 3 segments at the back (Fig. 6-24) was chosen as this showed a good cost to performance ratio. The measured rotor loss is compared with analytical and FE calculated results in no load in Table 6-5. The results show a fairly good agreement. One can see the benefit of implementing DS-PMS in the
'Loss Improvement' column which shows a reduction of around 200W in the measured results, resulting in a temperature reduction of around 11.2°C.

<table>
<thead>
<tr>
<th></th>
<th>Solid Magnets (W)</th>
<th>DS-PMS Magnets (W)</th>
<th>Loss Improvement (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>FEM</td>
<td>755</td>
<td>591</td>
<td>164</td>
</tr>
<tr>
<td>Analytical</td>
<td>781</td>
<td>554</td>
<td>227</td>
</tr>
<tr>
<td>Measured</td>
<td>807</td>
<td>607</td>
<td>200</td>
</tr>
</tbody>
</table>

Table 6-5: The table displays the total rotor power loss results of the 15kW machine at 150rpm, no load comparing solid magnets and DS-PMS magnets.

The graph in Fig. 6-26 shows how the power loss density changes through the magnet depth with DS-PMS. The figure’s legend displays the degree of segmentation, but note that the yoke-facing half of the magnet contains one less segment to prevent alignment of incisions. This causes the slight discontinuity seen in the power loss density function at 1mm. The power loss density can be seen to be at its highest at the face of the magnet, and exponentially decreases toward the yoke. For this reason, one targets the airgap-facing half of the magnet in SS-PMS and DS-PMS for segmentation as this is where the highest power loss density exists.

![Fig. 6-26: The power loss density vs. magnet depth with DS-PMS implemented.](image-url)
Fig. 6-27 shows the effect of DS-PMS on the overall magnet losses in the machine as predicted by FEM and analytical methods. They agree fairly well, showing a significant loss reduction with increasing number of segments.

![Graph showing effect of DS-PMS on magnet losses](image)

**Fig. 6-27:** The effect of varying degrees of DS-PMS as predicted by Analytical and FEM methods.

### 6.6.3 Comparison of Partial Magnet Segmentation with Full Magnet Segmentation

The comparison between the traditional segmentation methods and the methods proposed in this research work are presented here. Full magnet segmentation contains segments with 100% penetration. The results of implementing this segmentation technique are plotted against the results produced from single sided and double sided magnet segmentation in Fig. 6-28. As expected, FMS returns a typical segmented loss profile, however two anomalies exist in the single-sided and double-sided loss profiles.

In SS-PMS, the magnet loss value returned in the case of 8 segments is 128W, which is 42% of the unsegmented value. For a magnet that only has a segmentation penetration of 50%, the 58% loss reduction is higher than expected. This phenomenon is due to the higher power loss density in the stator-facing half of the magnet. Segmenting this half of the magnet has therefore more of an effect compared to if the back half of the magnet was segmented.

In the plot of DS-PMS, note that the loss profile actually outperforms the results of FMS. This is due to the unequal level of segmentation that exists on the front vs. the back of a DS-PMS magnet. In the case of S=2, the back half of the DS-PMS magnet is segmented with 2 segments, while the front half actually contains 3 segments.
6.7 Partial Rotor Yoke Segmentation

The method of partial rotor yoke segmentation aims to provide an alternative to the laminated rotor yoke for those designers looking to reduce yoke losses.

Fig. 6-29 shows the power loss density in the yoke of a solid magnet which has been partially segmented to various degrees shown by the different lines. Due to the skin depths of the various harmonics present in the solid, the currents only exist in the first ±20% of the solid. I.e. looking at the 4th harmonic, the skin depth can be calculated as:

$$\delta = \frac{2}{\sqrt{\omega \sigma \mu}} = \frac{2}{\sqrt{377 \times 6 \times 10^6 \times 600 \times 4\pi \times 10^{-7}}} = 1.08 \text{ mm}$$

(6.27)

This is the skin depth of the most penetrating of all of the harmonics, which in a rotor yoke of thickness 7.25mm, represents only a 15% penetration depth, which can be seen in Fig. 6-29. The skin depth essentially defines how deep into the conductive region each harmonic flows. Naturally this information is very helpful in determining the optimal degree of segmentation. Fig. 6-30 shows the relationship between the penetration depth of the segmentation incision and the loss results produced. There exists a value approximately twice the skin depth of the dominant harmonic where the loss results produced do not show any more improvement with increasing penetration depth. This is evident in Fig. 6-30 at 30% penetration, which is exactly twice the skin depth produced in equation (6.27).
Chapter 6  Results and Comparison

Fig. 6-29: The power loss density in the steel rotor yoke as a function of distance from the magnet/yoke interface to the outer yoke diameter.

Fig. 6-30: A graph showing the relationship between the solid loss in the yoke and the penetration depth of the segment incisions.

The effect of the number of rotor yoke segments in minimizing the solid loss in the rotor yoke is shown in Fig. 6-31. The agreement between the analytical and FEM methods is slightly compromised by the effect of varying permeability in the steel. However, a good performance return is delivered as one implements higher levels of segmentation. Including 128 incisions around the inner yoke diameter of the machine with a penetration of 17.5% of the 7.25mm yoke thickness shows a solid loss reduction (FE calculated) of 68% in the yoke. This improvement translates into a 24.8°C reduction in the yoke temperature and an overall efficiency improvement of 2.94%.
6.8 Overall Results Comparison

The results produced in this chapter cover many objectives, so this subsection will just present the main FEM results from the machine at full load to see the benefits of the various loss reduction techniques side-by-side. Given the strong agreement that the FE results have had with the measured results, one can be confident of the predictions in Table 6-6. The loss reduction techniques in the table include single and double sided partial magnet segmentation and partial rotor segmentation.

<table>
<thead>
<tr>
<th>Segmentation level</th>
<th>SS-PMS</th>
<th>DS-PMS</th>
<th>PRYS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnet/Yoke Loss</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>229</td>
<td>229</td>
<td>649</td>
</tr>
<tr>
<td>2</td>
<td>180</td>
<td>123</td>
<td>580</td>
</tr>
<tr>
<td>4</td>
<td>138</td>
<td>53</td>
<td>325</td>
</tr>
<tr>
<td>8</td>
<td>118</td>
<td>32</td>
<td>208</td>
</tr>
<tr>
<td>% Loss Reduction</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>21.4</td>
<td>46.3</td>
<td>11</td>
<td></td>
</tr>
<tr>
<td>39.7</td>
<td>76.9</td>
<td>50</td>
<td></td>
</tr>
<tr>
<td>48.4</td>
<td>86</td>
<td>68</td>
<td></td>
</tr>
<tr>
<td>Temperature Reduction</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>2.75</td>
<td>5.95</td>
<td>3.8</td>
<td></td>
</tr>
<tr>
<td>5.11</td>
<td>9.88</td>
<td>18.2</td>
<td></td>
</tr>
<tr>
<td>6.23</td>
<td>11</td>
<td>24.8</td>
<td></td>
</tr>
<tr>
<td>Efficiency improvement</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>0.33</td>
<td>0.71</td>
<td>0.46</td>
<td></td>
</tr>
<tr>
<td>0.61</td>
<td>1.173</td>
<td>2.16</td>
<td></td>
</tr>
<tr>
<td>0.74</td>
<td>1.31</td>
<td>2.94</td>
<td></td>
</tr>
</tbody>
</table>

Table 6-6: Full load results comparison in FEA for the machine with various loss reduction techniques.
Table 6-6 has the various degrees of segmentation plotted with various types of segmentation vs. the loss reduction benefits. The segmentation level refers to the number of segments per magnet in the case of magnet segmentation and number of segments in a 2π mechanical radian angle in the case of rotor segmentation. The row labelled 'Magnet/Yoke Loss' refers to the loss induced in either the magnet of the rotor yoke depending what column it is in. The table also quantifies these losses as percentage reduction compared to the case where the magnets/yoke are solid. The temperature reduction is calculated from equation 6-22, and the efficiency improvement refers to the loss as a quotient of the full load power of 15kW.
7 Conclusions and Recommendations

7.1 Analytical Calculation Methods

7.1.1 Magnet
The analytical calculations of magnetic fields and eddy currents in the magnets have proved to be very accurate as the relative permeability for the material stays constant. Choosing to calibrate the model using FEM proved to be a good method as:

- Accuracy was ensured as the FE model takes saturation and permeability variation into account.
- Intermediate results could be compared with the finite element to ensure consistency.
- Model was adaptable to no load calculation, where a different set of harmonics was collected from the same stator surface location.

Using FEM also helps when calculating the losses due to slotting as the same result can be extracted, except that current in the stator windings is set to zero.

7.1.2 Rotor
The analytical calculations of the rotor fields were less accurate than in the case of the magnet. This is due to variation in permeability of the steel material due to the effects of saturation. However, one could still make very informed design choices using the information provided by the analytical model:

- The level of segmentation can be calculated using knowledge about the highest order, damaging current density harmonic. This is used in conjunction with the nyquist critical segmentation pitch to influence the degree of partial rotor segmentation.
- The optimal segmentation penetration can be calculated by equating it to twice the skin depth of the most deeply penetrating harmonic.

The average flux density can be extracted from a FEM model to calibrate the analytical model if higher degree of accuracy is required. Further work could possibly implement an iterative method for calculating permeability at individual points using a B-H Curve.

7.2 Optimal Segmentation
The effectiveness of segmentation is costly and one must therefore ensure that a level of segmentation is selected that provides good performance returns without over segmenting. This requires the designer to follow a process where the best possible segmentation level is
chosen for a given machine. The optimal segmentation process begins with the designer having intimate knowledge of the harmonic content of the airgap. The highest damaging current density harmonic must be identified and the Nyquist Segmentation principle and Fig. 4-16 can be used to determine an optimal level of segmentation.

One must also weigh up the construction technique used to segment the magnets and yoke and quantify the amount of material lost in the process. This material loss manifests itself as a strength drop in the case of the magnet, and a flux density drop in the case of the rotor yoke. This is an important consideration to take into account as this essentially weighs up segmentation with machine performance and cost.

### 7.3 Partial Magnet Segmentation

The methods of partial magnet segmentation are very effective in reducing the eddy current losses in the solid magnets. They provide a unique solution that combats the construction difficulties of full segmentation while still deriving similar performance benefits.

In design, one can bring down the cost of the machine with partial magnet segmentation. There exists a level of segmentation where the performance improvements are negligible compared to the extra levels of segmentation which decrease magnet strength. Optimization is therefore a critical issue and depends on a few factors including the level of segmentation and magnet strength lost due to cutting and cost.

The comparison between partial and full magnet segmentation clearly demonstrates that with good design and simple construction techniques, one can make great efficiency improvements and magnet temperature reduction with successful implementation of partial magnet segmentation.

### 7.4 Partial Rotor Yoke Segmentation

The solid rotor yoke provides an electrically conduction medium for any parasitic harmonic, and for those machines that have asynchronous harmonics that are large enough in magnitude, significant eddy currents can flow. The laminated yoke is an option which would effectively reduce the rotor yoke losses to zero; however it is more costly due to the cutting of laminations. It also requires an external hub in which to house the laminations adding more weight and cost.

Partial rotor yoke segmentation provides a means to significantly reduce the eddy current losses in the solid steel rotor yoke. The benefit is that:

- PRYS is cheaper than using a laminated yoke with hub.
- One retains the construction benefits of using the yoke for electromagnetic and structural purposes, which make the building of the machine simpler and cheaper.
- The construction can be done without any expensive machinery.
Care must be taken in construction to ensure that the rotor yoke is accurately rolled and machined. Inaccuracies in this process can cause an imbalance in the system and also variation in airgap thickness.

### 7.5 Overall Machine Design

One does not need to understand everything about rotor harmonics in order to design a machine in such a way to reduce rotor losses to an optimal degree. There are two main design suggestions related to magnet and yoke segmentation to be taken from this research, which relate to the degree of segmentation and the incision penetration.

- **Segmentation degree**: The number of the highest significant harmonic order should be doubled to reach the number of segments that should be implemented throughout the whole circumference of the machine.
- **Segmentation penetration**: The magnets should be segmented with DS-PMS with the number of segments chosen according previous design suggestion. The rotor does not have to have a high penetration as the skin effect dictates where the current occurs. For this reason, the skin depth of the harmonic with the thickest skin depth should be doubled to provide the penetration depth.

Using the above design considerations, if one extracts the 'best case' scenario from Table 6-6, one can deduce that implementation of double-sided partial magnet segmentation with four segments coupled with partial rotor segmentation with 128 segments, the performance benefits would be:

- **Efficiency improvement of 4.11%**
- **Temperature reduction of 34.7°C**, which would have the machine rotor running at around 75°C, which is well within the operating temperature range of the magnets.

### 7.6 Machine Design for WG Applications

The design of large wind generators with direct drive drivetrains is becoming an increasingly attractive prospect, but designers are far from being able to pin down the optimal design that will take the industry forward. Some options that face designers are:

- **Surface mounted magnet vs. buried magnet rotor design**: Two issues arise with surface mounted magnets that need to be addressed. Firstly, the large magnets need to be handled and placed accurately which is a challenge due to the large attractive forces at play. Secondly, the epoxy for bonding the magnets to the rotor surface needs to be rated for at least 30 years. This is something that no manufacturers are willing to guarantee yet. Buried rotor designs are more secure for the magnet and provide slightly easier manufacturing options, however the performance suffers due to increased leakage flux.
• Winding design: The concentrated, non-overlapping, single layer winding design with open slots is desirable due to its ease of manufacture. However it requires the use of a more expensive rotor in order to maintain efficiency standards. A distributed, 3-phase winding is another option which enables use of a cheaper rotor, however it is more expensive to manufacture.

• Laminated vs solid yoke rotor: The use of a solid yoke rotor without any form of segmentation is very likely to suffer from high eddy currents due to winding harmonics, slotting harmonics and switching harmonics from the converters regardless of the stator design.

My conclusion is that for WG industry applications one should manufacture the stator as inexpensively as possible. This could mean using single layer, concentrated, non overlapping coils with uneven, open slots. This stator design will give rise to many dangerous harmonics in the airgap and rotor, however if one redesigns a more expensive stator with less spatial winding harmonics, dangerous time harmonics caused by the switching circuitry will still exist. The solution is to design a rotor that can cope with the dangerous spatial and time harmonics without incurring excessive loss. The extra expense that this will incur will be balanced by the cheaper stator and the extra efficiency gained. This rotor can be designed with any one of the options presented in this work for magnet and rotor yoke loss reduction or even possibly a laminated rotor yoke.

Further work should include the effect of partial segmentation of the magnets and rotor yoke on the time harmonics within the current waveform.
8 References


Papers unrelated to PM eddy current losses directly:


“High-performance permanent magnet brushless motors with balanced concentrated windings and similar slot and pole numbers”, Bojan Stumberger, Gorazd S Tumberger, Miralem Hadziselimovic, Anton Hamler, Milen Trlep, Viktor Gorican, Marko Jesenik. Faculty of Electrical Engineering and Computer Science, University of Maribor, Smetanova 17, 2000 Maribor, Slovenia.


“A New Three-Phase Doubly Salient Permanent Magnet Machine for Wind Power Generation”, Ying Fan, Student Member, IEEE, K. T. Chau, Senior Member, IEEE, and Ming Cheng, Senior Member, IEEE.

"Analytical and Numerical Computation of Air-Gap Magnetic Fields in Brushless Motors with Surface Permanent Magnets", Keld Folsach Rasmussen, Member, IEEE, John H. Davies, T. J. E. Miller, Fellow, IEEE, Malcolm Iain McGilp, and Mircea Olaru


