2D & 3D ELECTROMAGNETIC AND MATERIAL LOSS ANALYSIS OF
AN AXIAL FLUX PERMANENT MAGNET MACHINE

by

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ABSTRACT

2D & 3D ELECTROMAGNETIC AND MATERIAL LOSS ANALYSIS OF AN AXIAL FLUX PERMANENT MAGNET MACHINE

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Advances in computing hardware, software and manufacturing have helped revive the academic and commercial interest in a machine topology known as an axial flux machine (AFM). Traditional linear circuit analysis and basic non-linear Finite Element Analysis (FEA) have provided suitable designs for a wide range of AFM prototypes worldwide. This thesis examines the design of a particular single-sided AFM prototype developed at the Charles Darwin University in 1993 for use in a solar powered racing car. A thorough examination of the machine design shows a reasonable agreement between design specifications and measured outcomes after accounting for a 10% decrease in permanent magnet remanence. The unique property of the AFM whereby mechanical flux weakening is achieved via air gap adjustment is investigated, focusing on the resulting change in several important machine parameters. The effect of an air gap distance increase from 1mm to 4mm is confirmed to reduce the machine constant by 36%. 2D Transient FEA shows that the minimum in cogging torque occurs at a magnet arc width of 128.2° at a nominal air gap distance of 2mm. 3D FEA confirms the fact that a significant radial flux component exists in both the rotor and stator under stalled conditions. While it is unlikely that radial flux component will significantly impinge upon the efficiency of the CDU solar car motor, lower pole count machines are likely to suffer. The effect of rotor lamination combined with 3-segment per pole magnet segmentation is shown to have the potential for a rotor eddy current loss reduction of 5.5 watts.
DECLARATION

I hereby declare that the work herein, now submitted as a thesis for the degree of Doctor of Philosophy by research of the Charles Darwin University, is the result of my own investigations, and all references to ideas and work of other researchers have been specifically acknowledged. I hereby certify that the work embodied in this thesis has not already been accepted in substance for any degree, and is not being currently submitted in candidature for any other degree.

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### LIST OF SYMBOLS

- $\alpha_{\text{mech}}$ - mechanical magnet arc width
- $\tau$ - time constant
- $\sigma$ - conductivity
- $\nu$ - kinematic viscosity
- $\rho$ - mass density
- $\mu$ - permeability
- $\theta_e$ - electrical displacement angle
- $\eta_r$ - rated efficiency
- $\omega$ - angular frequency
- $\omega_E$ - electrical angular frequency
- $A_{cu}$ - copper cross-sectional area
- $B$ - flux density
- $b$ - lamination thickness
- $B_{pk}$ - peak flux density
- $C_a$ - anomalous loss constant
- $C_e$ - eddy current loss constant
- $C_h$ - hysteresis loss constant
- $C_M$ - dimensionless moment coefficient
- $d$ - displacement
- $d_{ag}$ - air gap distance measured axially between the permanent magnets and the stator teeth
- $d_{ha}$ - thickness of the stator back-iron (stator axial thickness minus $d_{sd}$)
- $d_{mr}$ - magnet thickness
- $d_{nw}$ - linear magnet width measured over the mid-radius surface
- $d_{pw}$ - linear pole width or pitch measured over the mid-radius surface
- $d_{ri}$ - thickness of the rotor iron (measured axially)
- $d_{sd}$ - depth of winding slots into the stator toroid (measured axially)
- $d_{sw}$ - linear width of the winding slots measured over the mid-radius surface
- $D_l$ - inner machine diameter
- $D_o$ - outer machine diameter
- $E_a$ - armature EMF
- $f$ - frequency
- $f_{sw}$ - switching frequency
- $F$ - force
- $F_{axial}$ - force in the axial direction
- $g$ - gravitational constant of acceleration
- $H_{irr}$ - irreversible hysteresis component
- $I_{dc\text{max}}$ - maximum DC current
- $I_{HYS}$ - hysteresis band current amplitude
- $k_{anom}$ - anomalous loss factor
- $k_{eddy}$ - eddy current loss factor
- $K_{cu\text{-ut}}$ - copper utilisation of fill factor
- $K_{sw}$ - slot width factor
- $L_{t1}$ - line to line inductance
- $L_{q1}$ - quadrature axis inductance
<table>
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<th>Description</th>
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<tr>
<td>$I_{\text{qm}}$</td>
<td>minimum quadrature axis inductance</td>
</tr>
<tr>
<td>$I_{\text{single}}$</td>
<td>single turn inductance</td>
</tr>
<tr>
<td>$m$</td>
<td>number of phases</td>
</tr>
<tr>
<td>$m_t$</td>
<td>total mass</td>
</tr>
<tr>
<td>$M$</td>
<td>frictional moment</td>
</tr>
<tr>
<td>$n$</td>
<td>Steinmetz index</td>
</tr>
<tr>
<td>$n_{\text{pp}}$</td>
<td>number of slots per pole</td>
</tr>
<tr>
<td>$n_{\text{tp}}$</td>
<td>number of turns per phase</td>
</tr>
<tr>
<td>$p$</td>
<td>number of pole pairs</td>
</tr>
<tr>
<td>$P_{\text{v}}$</td>
<td>volumetric core loss density</td>
</tr>
<tr>
<td>$P_{\text{aero}}$</td>
<td>aerodynamic loss</td>
</tr>
<tr>
<td>$P_{\text{anom}}$</td>
<td>anomalous loss</td>
</tr>
<tr>
<td>$P_{\text{bearing}}$</td>
<td>bearing loss</td>
</tr>
<tr>
<td>$P_{\text{bearing-A}}$</td>
<td>axial component of the bearing loss</td>
</tr>
<tr>
<td>$P_{\text{bearing-R}}$</td>
<td>radial component of the bearing loss</td>
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<tr>
<td>$P_{\text{eddy}}$</td>
<td>eddy current loss</td>
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<tr>
<td>$P_{\text{em}}$</td>
<td>electromagnetic loss</td>
</tr>
<tr>
<td>$P_{\text{hys}}$</td>
<td>hysteresis loss</td>
</tr>
<tr>
<td>$P_{\text{spin}}$</td>
<td>spinning loss</td>
</tr>
<tr>
<td>$P_{\text{wetted disk}}$</td>
<td>power used to move fluid on a single side of a rotating disk</td>
</tr>
<tr>
<td>$r$</td>
<td>resistance</td>
</tr>
<tr>
<td>$r_{\text{ds(on)}}$</td>
<td>MOSFET drain to source on-state resistance</td>
</tr>
<tr>
<td>$R$</td>
<td>(outer) disk radius</td>
</tr>
<tr>
<td>$R_{\text{ab,bc,ca}}$</td>
<td>resistance between phase a &amp;b, phase b &amp; c and phase c &amp; a</td>
</tr>
<tr>
<td>$\mathcal{R}_{c}$</td>
<td>Reynolds number</td>
</tr>
<tr>
<td>$R_{\text{L-L}}$</td>
<td>line to line resistance</td>
</tr>
<tr>
<td>$t$</td>
<td>time</td>
</tr>
<tr>
<td>$T_{\text{em}}$</td>
<td>electromagnetic torque</td>
</tr>
<tr>
<td>$T_{\text{max}}$</td>
<td>maximum torque</td>
</tr>
<tr>
<td>$V_{\text{dc}}$</td>
<td>DC bus voltage</td>
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<td>$V_{t}$</td>
<td>terminal voltage</td>
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<td>Abbreviation</td>
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<td>CDU</td>
<td>Charles Darwin University</td>
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<td>2D</td>
<td>2 Dimensional</td>
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<td>Axial Flux Machine</td>
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<tr>
<td>NTU</td>
<td>Northern Territory University</td>
</tr>
<tr>
<td>EMC</td>
<td>Electronic Motor Controller</td>
</tr>
<tr>
<td>CDUSCM</td>
<td>Charles Darwin University Solar Car Motor</td>
</tr>
<tr>
<td>EMF</td>
<td>Electro-Motive Force</td>
</tr>
<tr>
<td>MOSFET</td>
<td>Metal Oxide Semiconductor Field Effect Transistor</td>
</tr>
<tr>
<td>NdFeB</td>
<td>chemical symbol abbreviation for a (permanent magnet) material consisting mainly of Neodymium, Iron &amp; Boron</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Test</td>
</tr>
<tr>
<td>GPIB</td>
<td>General Purpose Interface Bus</td>
</tr>
</tbody>
</table>
Chapter 1

INTRODUCTION

1.1 Background

The Charles Darwin University (CDU), formerly the Northern Territory University, was host to the Northern Territory Centre for Energy Research whose focus was on energy systems research. A large part of this effort was directed towards improving the efficiency of energy conversions. This thesis presents one aspect of this energy conversion research program involving a particular electric machine known as the single-sided, iron-cored, permanent magnet, axial flux, brushless, direct current machine. In particular it is the axial flux architecture of this machine that makes it the subject of this research.

1.1.1 The History of Electric Machines

The invention of electric machines followed shortly after the discovery of a direct relationship between electric currents and magnetic flux. The reaction between non-aligned magnetic fluxes allowed mechanical work to be produced from electrical energy, which is the fundamental function of all electromechanical machines. From this discovery, rotating electric machines were developed and utilised in a plethora of applications. From the earliest rotating machines to the latest designs, the majority of these machines employ a configuration that directs magnetic flux radially across an air gap between a rotating component, the rotor, and the stationary component, the stator. Well known because of this radial air-gap-flux architecture, this type of machine is most commonly referred to as a radial flux machine. Until recently, the radial flux machine has been considered to be the easiest and most practical electrical machine to design and manufacture. The recent advance in computing, manufacturing and material technology is changing the domination of the radial flux machine and shifting the research focus towards more sophisticated machines designed previously by a multiple-prototype, trial and error method. Using two or more discs and having the primary component of the air-gap flux in the direction of the axis of rotation, the axial flux architecture has more seriously developed over the last decade. The Axial Flux Machine (AFM) is now beginning to appear on commercial production lines as a major topological competitor to
the radial flux machine. Several studies have outlined the performance differences between these two major topologies [1-4].

1.1.2 Modern Electrical Machines
Dramatic advances in high power machine design have been possible due to the ongoing development of rare-earth permanent magnet materials and power electronic components [5, 6]. The electronically commutated or brushless machine, which utilises these technologies has emerged as one of the most efficient means of converting electrical energy into mechanical energy and vice versa [7, 8]. The energy savings of the brushless machine when compared to the brushed alternative have already been proven with up to 15% less consumption in low voltage applications [9]. Adding to these advantages, brushless machines capable of producing 1 to 10 kilowatts of mechanical power at efficiencies in the upper 90% region are readily available at reasonable price. One disadvantage of the more common radial flux brushless machine is their poor ability to dissipate armature losses due to the thermally insulative properties of the cylindrical air gap that surrounds much of the rotor surface.

1.2 Properties of the Axial Flux Machine
In broad terms, the machine considered in this research falls into the categories of;

- Three phase brushless machines,
- Axial Flux Machines (AFMs) and
- Direct Drive Machines

This classification brings together the advantages of the individual areas as noted in [9-18].

1.2.1 Three Phase Brushless DC Machines
The brushless machine was developed some years ago with the global goal of energy efficiency and plant reliability. To develop a balanced and continuous rotating stator field with constant flux magnitude, a minimum of two (2) phases is required. A phase count of three (3) has been adopted, as is common practice, in an attempt to reduce harmonics, hence torque ripple, caused by the non-linear behaviour of the magnetic materials. Higher phase-count machines have been developed and utilised by other researchers in applications that require less torque ripple than that obtained from practical 3 phase designs, however generally a three-phase machine is as efficient.
1.2.2 Electronic Commutators and Controllers

As an integral part of any Brushless DC Machine (BDCM), the electronic commutator (EC) is critical for its reliable and efficient operation. In the case where the BDCM acts only as a motor, the EC is generally referred to as an inverter, converting the DC input voltage to a multi-phase AC voltage, though not necessarily sinusoidal. Conversely, when used as a generator only, the EC is generally referred to as a rectifier. Some confusion may become apparent when trying to define a machine with an EC as either a brushless DC machine or a synchronous AC machine. The reason for this problem is the focus of the definition upon either the actual machine in question or the combination of the machine and its EC. For the purpose of this research, the definition is made with respect to the machine component only. That is, a brushless DC machine is defined here as having a trapezoidal-like back EMF whereas a synchronous AC machine exhibits a sinusoidal EMF with low harmonic content.

In many cases where an EC is employed, either current or voltage control is added to the functionality of the commutator producing an electronic machine controller (EMC). By adding either current or voltage control to machines used in applications where load and operating points change, there is generally a significant increase in the overall system efficiency. Ideally, the characteristics of machines and their controllers are matched to provide efficient electromechanical energy conversion. The efforts of other researchers have shown that there are significant efficiency trade-offs in the design and operation of the controller and the machine itself. For example, it has been shown that iron losses in a machine can be reduced by using an ironless stator however this leads to the need for large inductors in the EMC, which are themselves lossy [18]. In contrast to this, the use of an inverter with higher efficiency, such as the trapezoidal current-controlled inverter, the challenge then moves towards designing a trapezoidal flux machine of similarly high efficiency. As core losses are related to the rate-of-change of flux, the higher order harmonics contained in the trapezoidal excitation tend to lead to a less efficient machine. An additional advantage of the brushless DC machine is that it has a higher volumetric power density than their synchronous AC counterparts [1].

1.2.3 The Axial Flux Architecture

The difference between axial flux and radial flux machines is straightforward and is defined by the primary direction the flux traverses the stator to rotor air gap. Even with such a definition it is possible that there exist machines that cross this definition and
hence fall into the category of hybrid flux machines. Physically, the major difference between the dominant topologies of radial and axial flux is the geometry ratio. That is, the ratio of machine diameter to axial length. Axial flux machines have a large geometry ratio and consequently have been sometimes referred to as pancake or disc motors. This geometry favours in-wheel direct-drive traction, automotive motor-generators and other such applications where there are restrictions on the axial length of the machine. Whilst axial flux machines can exhibit high efficiency, the other criteria in which they excel is volumetric power density. This means that more of the physical structure of an axial flux machine can be used to produce power, rather than just provide mechanical support. A further improvement in volumetric power density can be achieved by using both faces of the stator in a configuration commonly referred to as the double-sided machine [1, 15, 16, 18-21]. References to double-sided machine AFMs also include those that employ two stators either side of a single rotor [22, 23]. Using both faces of the stator however is not exclusive to axial flux machines and can be transformed back to the radial flux architecture. A number of variations of the axial flux architecture have been prototyped in the last two decades including the use of switched reluctance and induction principles [24-30]. Campbell was one of the first to seriously consider the axial flux machine [31] while others have drawn comparisons with the axial flux topology since then [1-4, 32]. The ability of the axial flux machine to be built from an axial stack of modular machines has also been practically explored with significant implications to the mass manufacture of a variety of machine ratings [33-36].

In summary, the axial flux architecture in its many forms has been demonstrated as a high efficiency, compact machine particularly suitable for direct drive traction systems.

1.2.4 Applications of the Axial Flux Machine

Due to its characteristically high efficiency and volumetric power density, the axial flux machine (AFM) has been identified as the best candidate for applications including direct-drive traction machines [14, 37, 38], wind turbine generators [27] and motor-generators for internal combustion engines and gas turbines [39]. The central motivation for this research is not only applying the AFM to these applications, but to enhance the performance of the AFM within the criteria that makes it the preferred architecture.
In particular, three attributes have been chosen that produce the optimum design outcomes for these applications. The chosen attributes are:

- System Efficiency,
- Volumetric Power Density and,
- Materials and Manufacturing Cost.

In addition to these major areas, the axial flux machine also has the advantage whereby the magnets may be retained against centrifugal forces using ferromagnetic materials at the rotor circumference without additional magnetic losses. In addition, this magnet retention property does not impede upon the minimum allowable air gap dimension.

1.3 Motivation
Aside from pure scientific research, other considerations such as environmental effects are driving forces for this research and organisations interested in those effects have funded many of aspects of the research program at the former Northern Territory University.

1.3.1 Environmental motivations
The most recent generation of electrical machines has been driven by an increase in the emphasis placed on responsible and efficient use of the world’s limited energy resources. In 1998 it was stated that 90% of Australia’s electrical energy requirements are derived from fossil fuel based power generation and about 60% is consumed by electrical machines [40]. Given these two factors, the significance of machine efficiency improvement is clearly evident. It was predicted in a 1978 U.S. government report on energy efficiency and electric motors [41], that by utilising higher efficiency motors an energy saving of about 35 billion kWh per year could be obtained by 1990. This is equivalent to about 5% of the total U.S. electric power consumption for that year. To add to this, a worldwide increase in oil price over the last few decades has focused a renewed effort in the development and the commercial introduction of renewable energy systems. The capital cost of these systems demand that each electrical machine, whether it be used in power generation or consumption, be specified at the higher end of the efficiency scale.

1.3.2 The NTU Axial Flux Brushless DC Machine
In 1992, Associate Professor Dean Patterson at the former Northern Territory University, having realised the advantages listed above, designed the University’s first, single-sided,
axial-flux, brushless DC machine. Applied to the University’s solar racing car seen in Figure 1, the *Fuji Xerox Desert Rose*, this machine has seen many hours of real world operation as well as laboratory testing.

![Fuji Xerox Desert Rose](image)

This design has changed only slightly since then and is the subject of many simulations and experiments outlined in this thesis. In favour of its performance is the fact that in 1995 the US company, New Generation Motors, signed a license to manufacture the design for the world solar racing car market. Figure 2 shows a photograph of the machine as it is today, referred to herein as the CDU Solar Car Motor (CDUSCM).

### 1.3.2.1 Machine Construction

With any novel machine design comes the difficulty of manufacturing the end result. Figure 3 shows an exploded view of the CDUSCM with the windings omitted for clarity. Figure 4 shows a cross-section view of the assembled motor with the major components labelled.
Figure 2: Photograph of the CDUSCM. Parts of this photo have been deliberately obscured to protect commercial knowledge.

Figure 3: Exploded view of the CDUSCM. The armature windings have been omitted to improve clarity.
As a relatively new construct, the single sided axial flux machine has a number of construction challenges.

Firstly, the stator is formed as a tape-wound toroid of strip electrical steel. The machines used for the testing outlined later in this thesis use a type of electrical steel called Ly-core® 140. Cost effectively creating the necessary slots in the toroidal laminations for the copper windings is the subject of on-going research and to date has best been achieved through a punch and wind technique that has the advantage of being able to create different profile slots. Both the punch and wind technique and the milling process used to manufacture the existing CDUSCM prototypes are likely to lead to additional stresses in the steel resulting in higher hysteresis loss. Post-manufacture annealing to resolve this possibility has not been performed to date.
Secondly, solid copper windings of rectangular cross-section are known to induce lower losses due to a higher slot fill-factor, but require a significant amount of time to construct manually. On the other hand, multi-stranded, round conductors that produce a lower efficiency machine due to a lower fill-factor are more readily mass-manufactured through conventional winding processes.

And thirdly, the dedicated thrust bearing used to efficiently support the large attractive forces between the magnet assembly and the stator, is physically large. The dimensions of this bearing often dictate either the minimum inner diameter or minimum axial length of the machine.

1.3.2.2 The CDU Electronic Motor Controller

Spawned at the same time as the CDU axial flux motor, the CDU EMC consists of a PI speed controller and a hysteresis-band current controller. Motor commutation timing is based upon three hall-effect sensors and a standard six-step switching scheme is used. The combination of these techniques plus the use of reverse conduction through the MOSFET switches allow it to achieve excellent efficiency figures throughout its operating region. In general, the loss components in the motor controller are considered as being MOSFET conduction loss, MOSFET switching losses and a very small stand-by loss. Of relevance to the motor design is the current waveform, determined by the EMC, which passes through the machine. While the hysteresis band is fixed at 10 amps, the switching frequency varies in the manner shown in Figure 5 due to the use of MOSFET reverse conduction with the switching scheme used [42]. This particular switching scheme is favoured as it has a reduced switching loss at low and high speeds that occur frequently for the application considered. In addition, the hysteresis band type current control can be synthesized using very simple and cheap electronic control circuitry. Further details of the switching scheme and the derivation of switching frequency variation are found in [42].
1.3.3 Electromagnetic Analysis

Designing magnetic circuits has historically relied upon fundamentally derived formulae or circuit specific experimentation. The introduction of computer based numerical methods first provided a design confirmation tool, and in the last decade an electromagnetic circuit design tool. Engineers strive for a recipe-like procedure to design magnetic circuits however it still remains an iterative process including design, simulation and analysis to achieve the optimum practical result.

The tools for machine design have historically been for sinusoidal excitation only because the harmonic solution to Maxwell’s equations is far less computationally intensive. Advances in modern computing technology have meant that the iterative numerical analysis requires less time and has become a viable alternative to the near antiquated design, prototype and measurement procedure. As a result, modern computing software has also become more advanced and has reduced its reliance on idealistic assumptions such as linear materials and sinusoidal excitation. Although not concurrently, the simulation of non-linear, anisotropic, three-dimensional structures can now be modelled under non-sinusoidal excitations. The software suite “Maxwell” compiled by the ANSOFT Corporation and “ANSYS Multiphysics” compiled by ANSYS Inc are good examples of such technology and have been used extensively to produce results presented...
in this thesis. In particular, Ansoft Maxwell version 2.0.55 has been used to generate much of the data shown in this thesis.

As trapezoidal flux (back-EMF) machines require rectangular currents to produce minimum torque ripple, the simulation of their operation using sinusoidal assumptions is only an approximation and results in significant torque ripple. Emerging computational tools use time-stepping algorithms that are inherently more accurate but impose a much greater penalty in terms of computation cycles. The numerical computations involved in developing this thesis use both harmonic and time-stepping algorithms to probe further into the behaviour of the electromagnetic interactions within the single-sided axial flux machine.

Static, harmonic and time-stepping algorithms have been exploited extensively relating to the radial flux topology and to a more limited extent the linear machine topology. While the application of finite element analysis to the axial flux machine is not new [13, 16, 22, 26, 43-48], its application has been sparse and has not quantified the three dimensional nature of the magnetic flux through the axial flux architecture. Nor has there been extensive research into minimising losses in all areas of the machine.

1.4 Original Contributions

This thesis contributes substantially to the knowledge of single-sided AFMs through the analysis of the three-dimensional, non-sinusoidal magnetic circuit that forms the basis of their operation. These contributions are in the areas of magnetic circuit analysis, force analysis, machine loss analysis and non-sinusoidal excitation induced loss analysis.

The original contributions of the author of this thesis are:

- An extensive experimental characterisation of the single sided AFM with respect to air gap distance.
- The investigation of the electromagnetic losses caused by hysteresis band current modulation.
- The investigation of the radial flux component found in the single-sided AFM.
1.5 Thesis Overview

Beyond this introductory chapter, Chapter 2 presents a summary of CDUSCM design including specification, materials and 2D FEA verification of the existing CDUSCM design. Chapter 3 follows examining measured machine parameters and compares them with the design outcomes. The use of transient and multi-static 2D simulations to investigate material losses, cogging torque and air gap distance variation is the subject of Chapter 4. Chapter 5 draws upon anecdotal evidence from Chapters 3 and 4 to present the nature of trans-laminar or radial flux component found in the CDUSCM. Two methods for loss reduction are detailed in Chapter 6. Finally, chapter 7 concludes and details areas for further research.

Appendix A contains a list of properties of the axial flux machine being researched, Appendix B details the method used to calculate the local gravitational constant for experiments and Appendix C contains copies of the papers published by the author of this thesis during his candidature [49, 50].
2.1 Introduction

Contemporary design philosophy is based upon the iterative process shown in Figure 6(a). Over recent decades this process has moved towards that shown in Figure 6(b) where a second iterative loop, including an analysis process, dominates the time required to derive a new machine design. This analysis loop has dramatically reduced the overall process cost by reducing the number of prototypes required to achieve a successful design.

The contemporary design process uses analytical equations that predict design outputs however their accuracy is limited by their simplifying assumptions. With the increasing use of computer-based numerical analysis, several assumptions have been withdrawn and the modern numerical analysis algorithms are quite complex and potentially quite accurate.

This chapter examines the early development of the CDUSCM and utilises some of the modern techniques to analyse its characteristics and outline the areas of interest addressed in later chapters.
2.2 Solar Car Motor Specification

The specification of the AFM design is, as with many machines, driven by the constraints and priorities of its application. In a competitive solar racing car, the efficiency, reliability, volume and weight of any component are of prime importance. By choosing a direct-drive strategy, the loss normally associated with a speed reduction device can be transferred in terms of additional mass to the direct drive machine. This increase in machine weight translates to an increase in copper and steel volumes thereby improving the overall drive-train efficiency. By using this strategy the design challenge shifts towards developing a machine that exhibits high torque at low speeds and is capable of a wide operational speed range. Details of the performance criteria and design process of the original machine can be found in [26].
Summarising the design process, the performance requirements were factored in a spreadsheet-based solution where some other factors could be monitored such as manufacturability (minimum dimensions) and the required specification of the brushless motor controller. Two-dimensional finite element analysis was also used to verify flux densities in various points in the geometry. Also monitored during the analytical part of the process were simple iron losses and the mass of the motor. Several revisions to the performance criteria have been made since this time, in particular the nominal speed. Advances in battery and solar cell technology over the last decade mean that vehicle speeds in excess of 100 km/h are now achievable and reasonable. The current properties of the machine with very similar physical components as that used in 1993 are detailed in Appendix A. In particular, the increment in nominal battery voltage to 55 V has allowed for increased vehicle speed up to 92 km/h with the air gap distance at 2mm.

As a desired performance characteristic of a road vehicle, a relatively high ratio of peak to nominal torque demands a moderate thermal mass or forced ventilation. The significant thermal mass found in the CDUSCM design is therefore a great advantage to its application. In addition, the physical arrangement of the axial flux machine allows for mechanical flux weakening by altering the air gap distance, which has the effect of tweaking the machine constant. These factors allow for some variation in the machine performance to suit changes in vehicle, road and weather conditions.

2.3 Construction Materials

Four different materials, not including air, are used to construct the magnetic circuit of the CDU solar racing car motor, one for each major component of the machine as described in the following sub-sections.

2.3.1 Stator Materials

The stator consists of the copper windings and the stator iron. The enamelled copper windings are a rectangular cross section with average outer dimensions of 2.25 x 5.60 mm and after accounting for the enamel thickness and corner fillets result in a fill factor of 70%.
The stator iron is constructed of low loss non-grain oriented silicon steel known as Lycore® 140, a product previously produced by BHP Steel\(^1\). The 0.35mm thick lamination material was chosen largely for its 2.2 watts per kg core loss at an induction of 1 Tesla at 100 Hz. A long strip of the lamination material was used to wind a toroid with the required inner and outer diameters. The lamination factor is assumed to be 98% as specified by the manufacturer. This lamination factor is in general agreement with measurements taken from the prototype machine. Because the machine was assembled prior to this work, the lamination factor measurement is based upon the dimensions of the prototype machine and the mass of a bare stator made to the same specification as the prototype machine. The manufacturer’s specification is used because the measurement accuracy is questionable. The lamination factor measurement inaccuracy is due to the stator volume measurement being based upon the assumption of a perfectly circular toroid and that the machining of the stator slots does not cause lamination deformation.

2.3.2 Rotor Materials

The rotor part of the circuit consists of the rare earth Neodymium Iron Boron (NdFeB) magnets and the rotor back iron. Due to the largely DC nature of the magnetic flux in the rotor, the rotor back iron consists of a solid piece of iron machined to shape. Two types of iron have been used in the CDUSCM rotor, mild steel and 3% cast silicon steel. Other researchers at CDU have found the difference in material loss of the two rotors to be within experimental error margins in tests performed to date. The silicon steel rotor has been used throughout all experiments presented in this thesis unless otherwise specified. The properties of rare earth magnets have evolved greatly since their discovery. More importantly, since they were adopted in volume production their cost has reduced immensely. The coercivity of commercial grade materials has reached 1290 kA/m and energy products of 480 kJ/m\(^3\) are also available [51]. The magnet material used in the CDUSCM is an N27 grade of the NdFeB material that exhibits a coercivity of 965 kA/m and an energy product of 270 kJ/m\(^3\).

\(^{1}\) BHP Steel was renamed to BlueScope Steel Limited in November 2003. LY-CORE® is a registered trademark of BlueScope Steel Limited.
2.3.3 Summary of Material Properties

The material properties used in the finite element analysis presented in this thesis are summarised in Table 1, Figure 7 and Figure 8. The skin depth of the magnetic materials is presented in Figure 9 and illustrates the different levels of penetration of alternating fields into those materials.

Table 1: Summary of material properties used in 2D and 3D finite element analysis.

<table>
<thead>
<tr>
<th>Material</th>
<th>Mass Density (kg/m³)</th>
<th>Electrical Conductivity (kS/m)</th>
<th>Thermal Conductivity (W/m K)</th>
<th>Relative Magnetic Permeability</th>
<th>Magnetic Coercivity (kA/m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air</td>
<td>1.29</td>
<td>0</td>
<td>0.025</td>
<td>1</td>
<td>-</td>
</tr>
<tr>
<td>Copper</td>
<td>8960 **</td>
<td>58,000 **</td>
<td>392.9 **</td>
<td>0.999991 **</td>
<td>-</td>
</tr>
<tr>
<td>Mild Steel</td>
<td>7849 **</td>
<td>10,000 **</td>
<td>61.0 **</td>
<td>non-linear, refer Figure 7 **</td>
<td>-</td>
</tr>
<tr>
<td>NdFeB Magnets</td>
<td>7438 **</td>
<td>625 **</td>
<td></td>
<td>1.09978 *</td>
<td>965</td>
</tr>
<tr>
<td>Ly-core ® 140</td>
<td>7650</td>
<td>2,000 **</td>
<td>30 **</td>
<td>non-linear, refer Figure 8.</td>
<td>-</td>
</tr>
</tbody>
</table>

* Typical recoil permeability specified.
** Property of typical material, not from material specification.

Figure 7: B-H curve used to simulate mild steel components of the axial flux machine.
Ly-core 140 ® B-H Curve

![B-H Curve](image)

Figure 8: B-H curve used to simulate Ly-core ® 140 components of the axial flux machine.

Material Skin Depth

![Material Skin Depth](image)

Figure 9: Skin depth values of the materials used in the CDUSCM.
2.4 Two Dimensional Machine Design

The iterative design & analysis process can be somewhat simplified by reducing the number of dimensions of the problem. In a traditional fashion this is done by reducing the three dimensional geometry to a two dimensional plane.

2.4.1 Design Plane and Mechanical Parameters

In radial flux machines it is common to consider a two-dimensional (2D) plane perpendicular to the axis of rotation that is representative over the axial length of the machine. This is deemed acceptable on the basis that the axial length of the machine is large when compared to the outer radius of the machine. The orthogonal equivalent for the axial flux machine is a cylindrical surface coaxial with the axis of rotation. With the exception of the winding slots, the geometry of the AFM is constant with respect to radius between the inner and outer diameters. This fact and the fact that both Faraday’s law and Lorentz’s force equation are linear with respect to radius allow us to consider the cylindrical surface as a representative surface of the machine. This 2D representation is not perfect because the winding slots are parallel-sided as will be discussed in Chapter 5.

As shown in Figure 10, the cylindrical surface can be considered in the \( \theta,z \) coordinate system, though it is easier to transform this into the rectangular x,y plane for analysis purposes. Similar to radial flux machines, the symmetrical nature of the geometry is used to reduce the area under consideration to that of one pole. The resultant plane under consideration is shown in Figure 11 with lines of odd symmetry occurring at \( x = \pm \frac{1}{2} d_{pw} \).
Figure 10: Illustration of the representative 2D surface where \( r = \frac{1}{2}(D_0 - D_t) \).

Figure 11: Two-dimensional representation of the AFM with odd symmetry occurring at \( x = \pm \frac{1}{2} d_{pw} \).
The basic geometric design parameters arising from the CDUSCM 2D model, as illustrated in Figure 11 are detailed in Table 2.

Table 2: Design parameters effecting 2D AFM model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Variable</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pole width at mid-radius</td>
<td>$d_{pw}$</td>
<td>mm</td>
</tr>
<tr>
<td>Stator back-iron thickness</td>
<td>$d_{bi}$</td>
<td>mm</td>
</tr>
<tr>
<td>Winding slot depth</td>
<td>$d_{sd}$</td>
<td>mm</td>
</tr>
<tr>
<td>Air gap thickness</td>
<td>$d_{ag}$</td>
<td>mm</td>
</tr>
<tr>
<td>Magnet thickness</td>
<td>$d_{mt}$</td>
<td>mm</td>
</tr>
<tr>
<td>Rotor iron thickness</td>
<td>$d_{ri}$</td>
<td>mm</td>
</tr>
<tr>
<td>Winding slot width</td>
<td>$d_{sw}$</td>
<td>mm</td>
</tr>
<tr>
<td>Magnet width</td>
<td>$d_{mw}$</td>
<td>mm</td>
</tr>
</tbody>
</table>

The other geometric parameters that are required to calculate the machine characteristics or 3-dimensionalise the design outcomes are detailed in Table 3.

Table 3: Design parameters required to calculate global machine characteristics.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Variable</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Machine outer diameter (excluding winding end turns)</td>
<td>$D_o$</td>
<td>mm</td>
</tr>
<tr>
<td>Machine inner diameter (excluding winding end turns)</td>
<td>$D_i$</td>
<td>mm</td>
</tr>
<tr>
<td>Number of pole pairs</td>
<td>$p$</td>
<td>-</td>
</tr>
<tr>
<td>Number of phases</td>
<td>$m$</td>
<td>-</td>
</tr>
<tr>
<td>Number of winding turns per phase</td>
<td>$w$</td>
<td>-</td>
</tr>
<tr>
<td>Number of slots per pole per phase</td>
<td>$n_{spp}$</td>
<td>-</td>
</tr>
</tbody>
</table>

There are also other geometric parameters such as slot profile, magnet shape and tooth shape that can be considered to improve certain design outcomes. These are difficult to include in the general design process because the number of possible outcomes becomes too great. Instead, these parameters can be considered once the overall design is more or less complete. The potential effects of these geometric parameters are discussed further in chapters 4 and 5.
The last parameter of importance is the maximum operating temperature, $T_{\text{max}}$, of the permanent magnets. This temperature is specified as 80° C for the material used in the solar car motor. This is the upper limit of operation and limits the maximum loss of the machine.

### 2.4.2 Electrical Parameters

To constrain the design process a number of electrical parameters are determined either directly from the design specification, or indirectly through the use of an electronic controller.

These electrical parameters are listed in table 4.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Variable</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal DC voltage</td>
<td>$V_{dc}$</td>
<td>volts</td>
</tr>
<tr>
<td>Maximum DC current</td>
<td>$I_{d\text{c(max)}}$</td>
<td>amps</td>
</tr>
<tr>
<td>Minimum inductance (line to line)</td>
<td>$L_{q\text{min}}$</td>
<td>henrys</td>
</tr>
</tbody>
</table>

### 2.4.3 Design Outcomes

Though the priorities of design outcomes may differ between applications there exists a fundamental set of design outcomes that are scrutinised against the original specification to determine the success of a design. These outcomes are listed in Table 5.

<table>
<thead>
<tr>
<th>Outcome</th>
<th>Variable</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electromagnetic torque</td>
<td>$T_{\text{em}}$</td>
<td>Nm</td>
</tr>
<tr>
<td>Base speed</td>
<td>$\omega_E$</td>
<td>rad.s$^{-1}$</td>
</tr>
<tr>
<td>Rated efficiency</td>
<td>$\eta_r$</td>
<td>percent</td>
</tr>
<tr>
<td>Resistance (line to line)</td>
<td>$R_{l-l}$</td>
<td>ohms</td>
</tr>
<tr>
<td>q-axis inductance</td>
<td>$L_q$</td>
<td>Henrys</td>
</tr>
<tr>
<td>Total mass</td>
<td>$m_t$</td>
<td>kg</td>
</tr>
<tr>
<td>Total volume</td>
<td>$V_t$</td>
<td>m$^3$</td>
</tr>
</tbody>
</table>

Cost is also a major concern, though when manufactured in large numbers is often directly related to the mass of the major components.
2.4.4 Design & Analysis Process

The design and analysis steps identified in Figure 6(b) often use a spreadsheet-based environment coupled with other analysis tools including circuit simulation and finite element analysis. A number of linked spreadsheets and 2D magnetostatic FEA are the basis of the original CDUSCM design [26].

More complex spreadsheets are now used to generate non-optimised machine designs for other applications. The optimisation is then performed using 2D and or 3D FEA.

2.4.5 Inspection of Contemporary Design using 2D Finite Element Analysis

The rapid development of computing hardware and hence software in the mid 1990’s led to what is now a readily available engineering tool – finite element analysis. When based on simplified forms of Maxwell’s equations, these tools can produce relatively accurate solutions of complex magnetic circuits. Nowadays, solutions can be obtained for non-linear, static, harmonic, transient, anisotropic and coupled field environments, though not necessarily all at once.

At the time of original design, the best FEA solutions available were for non-linear static or harmonic, two-dimensional problems.

2.4.5.1 Flux Density Verification

Critical to the design process is the assumed air gap flux density. Once an initial value is obtained from linear magnetic circuit analysis, it can be verified by a magnetostatic solution to the 2D single pole model. The CDUSCM geometry and automatically generated mesh for this model is shown in Figure 12. The flux density plotted over the solution space is shown in Figure 13. From this solution, the average flux density over specific areas of interest can be obtained as listed in table 6, and the mid-air gap flux density magnitude can be plotted, as shown in Figure 14.

<table>
<thead>
<tr>
<th>Area</th>
<th>Average Flux Density (Tesla)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rotor Iron</td>
<td>0.91</td>
</tr>
<tr>
<td>Air Gap</td>
<td>0.73</td>
</tr>
<tr>
<td>Stator Tooth</td>
<td>1.09*</td>
</tr>
<tr>
<td>Stator Back Iron</td>
<td>0.59</td>
</tr>
</tbody>
</table>

* Calculated in tooth directly under permanent magnet
Figure 12: 2D geometry and finite element mesh of density typical of that used for magnetostatic simulations.

Figure 13: Shaded flux density plot over the 2D region. Flux density units are Tesla.
Figure 14: Mid-air gap flux density.

The obtained average air gap flux density per pole of 0.73 T is within 2% of previous laboratory experiments [26] and within 3% of that obtained from linear circuit analysis assuming permeability values of non-linear materials at the 0.8 T operating point.

2.4.5.2 Axial Force & Radial Torque Verification

As a critical performance specification, the peak torque of the machine is also verified. This modelling scenario also allows for checking armature-induced saturation and demagnetisation. The 2D magnetostatic solution is based on the same geometry shown in Figure 12. The difference in this case is that two of the three conductors modelled are defined as current sources with a uniform current density and a magnitude equivalent to the rated current determined from the design process. The force solution is in the form of a total force vector that is dominated by the attractive or y-component. This vector is broken into circumferential (x) and axial (y) directions to determine the torque component and the axial component supported by the axial thrust bearing.

The force solution, derived using the preferred [52] virtual-work method, takes longer than the basic field solution to converge within acceptable limits [53]. This ultimately leads to a much finer mesh for the final solution, as shown in Figure 15. The most common maximum armature current used in solar racing cars in Australia is 600 A/slot. This is due to the Federal highway maximum road grade of 6% and the corresponding vehicle mass.
Taking the FEA results for this maximum current and accounting for symmetry, the peak active torque from a 600A/slot current density is 58.8 Nm and the axial force at this current is 4990 N. For comparison, the static axial force at a zero armature current is 4137 N.

Figure 16 shows the variation in peak torque and torque constant generated by different slot current densities. It shows that significant saturation occurs above the rated 600A per slot. As expected, the spreadsheet-based design value of 0.55 Nm/A is slightly less than the 2D FEA computed value of 0.58 Nm/A as the computed constant doesn’t account for the torque variation per cycle.
Peak Torque & Torque Constant vs Armature Current
(2mm Air Gap, 128.2 ø-degree magnet width)

Figure 16: Peak torque and torque constant computed by magnetostatic 2D FEA.

2.4.5.3 Demagnetisation

Using the same model as that for force calculation, the safe working area prior to demagnetisation can be estimated. Using winding current as a parameter, multiple solutions of increasing current are used to identify the point at which the flux density at a point in the magnets approaches zero. The flux density of the solution space at this point is shown in Figure 17. Currents greater than the level identified will induce negative flux densities and if sufficient cause permanent demagnetisation. Although it is possible to induce negative flux densities and not cause permanent demagnetisation, the margin between the negative flux density and zero allows for localised geometric variations and provides a certain level of EMC fault tolerance. For example, the scenario where current limiting fails and the large DC bus capacitor discharges through the motor. The current at which regions of the permanent magnets in the CDUSCM reach zero flux (found using 2D static FEA) is 630 amps.
2.4.5.4 Armature Inductance

Armature inductance is calculated, using equation (1), from the modelled single turn inductance determined by 2D magnetostatic FEA.

\[ L_{l-l} = 2p(n_{pp})^2(D_o - D_i)L_{single} \]  

(1)

Where \( L_{single} \) is the modelled single turn inductance per metre determined by numerical analysis. The modelled single turn inductance and calculated line-to-line inductances are 4.215 µH/m and 182.1 µH respectively. This value of inductance has proven to be suitable for the low voltage EMCs developed in this field [54].
3.1 The First Generation Prototypes
Extensive on-road testing of the first fixed air-gap prototype was carried out during the 1993 World Solar Challenge, travelling from Darwin to Adelaide, a distance of 3010 kms. Since this time 3 other machines based on the same design have been assembled including one variable air gap machine.

Having performed fundamental 2D analysis on the 1993 design, a thorough program of testing was undertaken to measure the actual outcomes of both the final design and the manufacturing process. Laboratory testing of the variable air gap prototype has been carried out and the results of these tests are presented in the following sections of this chapter.

3.2 Static Tests
3.2.1 Mechanical Dimensions
The mechanical dimensions of the machine are reported here for completeness and later reference to manufacturing tolerances. Table 7 summarises the measured values of the major dimensions of the machine including measured manufacturing tolerances.
Table 7: Mechanical dimensions of the variable air gap CDUSCM.

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Machine envelope diameter</td>
<td>320</td>
<td>mm</td>
</tr>
<tr>
<td>Machine envelope thickness</td>
<td>63</td>
<td>mm</td>
</tr>
<tr>
<td>Stator outer diameter</td>
<td>260 ±0.5</td>
<td>mm</td>
</tr>
<tr>
<td>Stator inner diameter</td>
<td>160 ±0.5</td>
<td>mm</td>
</tr>
<tr>
<td>Stator thickness</td>
<td>34.9 ±0.1</td>
<td>mm</td>
</tr>
<tr>
<td>Stator slot depth</td>
<td>18.75 ± 0.25</td>
<td>mm</td>
</tr>
<tr>
<td>Stator slot width</td>
<td>6.5 ±0.5/0.1</td>
<td>mm</td>
</tr>
<tr>
<td>Rotor outer diameter</td>
<td>260 ±0.2</td>
<td>mm</td>
</tr>
<tr>
<td>Rotor inner diameter</td>
<td>160 ±0.2</td>
<td>mm</td>
</tr>
<tr>
<td>Rotor thickness</td>
<td>11.1 ±0.05</td>
<td>mm</td>
</tr>
<tr>
<td>Magnet arc width</td>
<td>21.5 ±0.5</td>
<td>degrees</td>
</tr>
<tr>
<td>Magnet thickness</td>
<td>4.1 ±0.05</td>
<td>mm</td>
</tr>
</tbody>
</table>

* Dimension includes mechanical support structure.

There are 4 dimensions that vary significantly from the design values. These are rotor thickness, slot width, slot depth and magnet thickness. These measured variations ought to lead to slightly higher flux densities in the magnetic circuit.

### 3.2.2 DC Armature Resistance

The armature resistance often causes a significant part of the operational loss of a permanent magnet machine. It also determines the speed at which the maximum torque reduces below the rated torque.

#### 3.2.2.1 Test Apparatus

The DC armature resistance measurement was undertaken using a 15A DC power supply to inject current into the windings and two HP 34401A digital multimeters to measure the actual current and observed voltage drop across the windings. Each of the three line-to-line armature resistances were measured at 4 different current levels utilising brief measurements and rest periods to help prevent temperature variation during the tests carried out at 24 ±1°C. While the temperature dependence of the armature resistance is required to calculate it at a given operating temperature, its consideration here is overly complex. As an example, theoretically, a 10% increase in armature resistance will occur for every 25 °C rise above the initial temperature.
3.2.2.2 Results

The averaged results for each line-to-line resistance are reported in Table 8.

Table 8: Measured line-to-line resistances for the variable air gap AFM.

<table>
<thead>
<tr>
<th>Measurement</th>
<th>Symbol</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistance line a to line b</td>
<td>$R_{ab}$</td>
<td>0.002885</td>
<td>ohm</td>
</tr>
<tr>
<td>Resistance line b to line c</td>
<td>$R_{bc}$</td>
<td>0.002868</td>
<td>ohm</td>
</tr>
<tr>
<td>Resistance line c to line a</td>
<td>$R_{ca}$</td>
<td>0.002857</td>
<td>ohm</td>
</tr>
</tbody>
</table>

These values are, as expected, above the 25.4 mΩ value determined using the analytical spreadsheet approach. The deviation from the predicted values is explained by the spreadsheet's use of an “end-turn length factor” based only on geometry. The measured resistance includes the termination of the windings that is not accounted for in the spreadsheet analysis. For reference purposes, the “end-turn length factor” is defined as the ratio of the total length of copper material to the length of that which is contained in the slot.

The average line-to-line resistance is 28.7 mΩ and has been used to determine an empirical end-turn length factor of 2.94 rather than the value of 2.6 determined by measured geometry. This empirical factor will also include any deviations of the average conductor cross-sectional area from that derived from measured values.

3.2.3 Armature Inductance versus Position

The existence of significant armature inductance variation versus rotor position is useful, though not essential [55], to enable electrical flux weakening techniques. In addition, the absolute inductance and its variation are important as they are a primary factor in determining the maximum operating frequency of hysteresis band current controllers [54].

3.2.3.1 Test Apparatus

This test was performed using an “LCR Databridge” to measure the line-to-line inductance and an indexing head borrowed from a milling machine to vary rotor position as shown in Figure 18. The inductance was measured using an excitation frequency of 1 kHz and the rotor position varied in increments of 10 electrical degrees ($10^\circ_e$). The frequency of 1 kHz was chosen as it is a common operating frequency found by combining the CDUSCM, its hysteresis-band current controller and the solar racing car.
application. This frequency is also readily modelled by the FEA software without the need for overly-large FEA meshes.

![Armature inductance test apparatus.](image)

**Figure 18: Armature inductance test apparatus.**

### 3.2.3.2 Results

The measurement results are presented in Figure 19. Utilising a 2.0 mm air gap, the average actual inductance of 201.9 $\mu$H compares favourably with the 2D FEA result of 182.1 $\mu$H. The 10% difference can partially be accounted for through the inclusion of end-turn inductance contributions because a significant portion of the end-turn copper is in close proximity to the stator core. These inductance contributions cannot be accounted for in the 2D FEA or analytical analysis generally. The measurement offset due to the inductance of the test leads was measured as less than 0.5 $\mu$H, though this is bordering on the limit of accuracy of the LCR Databridge. Results presented under section 3.3.2 will also indicate that the flux density in the machine is lower than expected, which would also increase the inductance.
The relatively small variation in inductance is typical of a surface mounted permanent magnet machine. The measured inductance variation is 9.3% of the average inductance. This variation is due to permeability changes in the soft magnetic materials due to high levels of magnetisation and the permeability of the magnet material being almost 10% higher than the surrounding air.

### 3.2.4 Static Torque versus Position

Commonly referred to as cogging torque or reluctance torque, the net torque produced by a motor at various positions with zero armature current is undesirable for three reasons. Firstly it contributes to mechanical vibration during operation. Secondly it dictates a minimum electromagnetic torque required to start the machine. And thirdly, it is a symptom of time-varying flux within the machine leading to an additional electromagnetic loss.

#### 3.2.4.1 Test Apparatus

The static torque test measures the torque produced under static conditions where magnetic forces occur to reduce the potential energy in the magnetic circuit. The apparatus used in this test consists of a custom-made strain gauge based force transducer, an indexing head and a strain indicator. The force transducer measured the force applied by a torque arm mounted on the stator. The rotor was mounted in the chuck jaws of the
indexing head and rotated at 0.333° increments while the linear displacement of the torque arm was also measured to compensate for transducer flexure. The apparatus used is shown in Figure 20. The force transducer was calibrated using a series of weights and a pulley. The outcome of the calibration demonstrated a sensitivity of 4.06 mN /με and a maximum error of 1.2%.

![Figure 20: Static torque test apparatus.](image)

### 3.2.4.2 Results

The ability to obtain data about the unstable detent position is made difficult by the hysteresis characteristic of the magnetic materials used in the motor and the relatively small spring constant offered by the aluminium transducer arm. In contrast, the peak cogging torque is measured to an accuracy of ± 0.06 Nm. To help identify hysteresis and measure data about the unstable detent position, the rotor position was varied in both the forwards and reverse directions. The results of this test, shown in Figure 21, exhibit a wave shape that is roughly sinusoidal, and has a fundamental frequency that is 6 times the fundamental electrical frequency. The measured torque is very close to being periodic over 60° even though some deviation ought to occur due to the magnet arc width being 8.2° larger than the nominal 120°. Also apparent is the presence of hysteresis torque identified by the angular shift between the forward and reverse direction measurements.
The static torque versus position waveform shown in Figure 21 shows a good level of peak minimisation when compared to the base case of a $120^\circ$ magnet width where the peak torque calculated from 2D static FEA is almost 10 times greater. The optimisation of cogging torque using angular magnet width as a parameter is discussed further in chapter 4.

The measured cogging torque waveform demonstrates good agreement with the 2D multi-magnetostatic FEA results shown in Figure 22. The FEA predicted amplitude is within 10% of the measured data though the wave shape is somewhat different. Some variation is expected due to the high sensitivity of cogging torque upon the manufactured geometry and hysteresis. In addition, the low stiffness in the force transducer used in measurements is likely to lead to an error in position that increases with force magnitude. The FEA analysis results show some small variability due to meshing differences between each magnetostatic model. It will be shown in section 4.3 that the transient algorithm implemented in ANSOFT’s Maxwell 2D is less susceptible to apparent errors caused by differences in the successive static meshes.
3.3 Dynamic Tests

A series of tests are described in the following sections where the machine was driven at various constant speeds to attain voltages, torques and efficiencies within the normal operating range of the CDUSCM.

3.3.1 Test Apparatus

The major components used in the dynamic tests are part of an in-line dynamometer as pictured in Figure 23. The apparatus used was common to all of the dynamic tests.
The components of the dynamometer that were used are:

- Siemens 2GA1482 Eddy Current Brake
- Siemens 2GA1461 Universal Machine (operated as a DC motor)
- Torque Transducer – type dependant on test
- Device Under Test – the CDUSCM
- 48V, 120 A Custom CDU Brushless Motor Controller (detailed in [54])
- Power Ten Inc 4680D-10050, 5 kW DC Power Supply

Two torque transducers with different measurement ranges were used. For tests under load, a Lebow 500 in.lb unit was used and for the no load tests, a Staiger Mohilo 10Nm transducer was used. Table 9 presents a summary of the transducer specifications.
Table 9: Torque transducer specification summary.

<table>
<thead>
<tr>
<th>Manufacturer</th>
<th>Model</th>
<th>Capacity</th>
<th>Overall Accuracy</th>
<th>Measurement Method</th>
<th>Signal Tx Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lebow</td>
<td>1257</td>
<td>500 in.lb**</td>
<td>1.0 %</td>
<td>full bridge strain gauge</td>
<td>silver slip rings</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>angular displacement</td>
<td>rotary transformers</td>
</tr>
<tr>
<td>Staiger Mohilo</td>
<td>0411/03HE10W</td>
<td>10 Nm</td>
<td>0.5%</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

* The signal transmission method signifies the method by which the torque signal is passed from the internal rotating shaft to the stationary transducer body.

** 500 in.lb is equivalent to 56.35 Nm.

3.3.2 Armature Voltage versus Position & Speed

In this test, the CDUSCM was driven at various speeds by the DC machine. The brushless motor controller was disconnected.

The armature voltage or back-EMF was captured using a Tektronix TDS210 digital oscilloscope connected via GPIB to a computer running National Instruments LabView™ software.

3.3.2.1 Results

Figure 24 shows the line-to-line voltage, $V_{ab}$, versus position (displacement origin illustrated in Figure 11.) measurement results of tests carried out between 100 and 1000 rpm. Figure 25 illustrates the relationship between the average of the 3-phase rectified back-EMF versus rotor speed.

The average 3-phase rectified back-EMF measured at 1000 rpm was 44.8 volts and is only 79% of the spreadsheet analysis value of 56.8 volts. Similarly, the machine constant derived from measured back-EMFs is much lower than expected at 0.427 Nm/A. These differences are large and cannot be explained through manufacturing tolerances or measurement inaccuracies alone. To explain this difference, there must be a difference in the material characteristics. The characteristic most likely to change over time is that of the permanent magnet.
Figure 24: Measured back-EMF versus rotor position at various speeds.

Figure 25: Average of the measured back-EMF versus rotor speed including a linear trendline.

The difference between the spreadsheet analysis results and the measured voltage suggests that the air gap flux density is actually about 0.57 T.

The difference between measured and actual results can be verified by determining the machine constant from the results of a torque versus armature current test. To provide a
point of later comparison, the peak line-to-line voltage at 1000 rpm was measured to be 49.2 volts, which corresponds to the peak-volts per radian per second value of 0.47.

3.3.3 Torque versus Current

With the brushless motor controller energised, the average armature current was measured using a 100 amp Tektronix A6303 current probe and the recommended A502 current probe amplifier. The torque was measured using the Lebow 500 in.lb torque transducer. The test was performed with the motor stalled in its peak torque position immediately after calibration.

3.3.3.1 500 in.lb Torque Transducer Calibration

The measurement of torque up to the expected maximum of 55Nm required the 500 in.lb Lebow torque transducer. The transducer, connected to LabView, was calibrated using a series of weights up to 9.8 kg at a torque arm length of 464 mm. The maximum deviation of the transducer output from the average of both runs, was equivalent to ±0.46 Nm. The calibration data is plotted against the linear calibration equation in Figure 26.

![500 in.lb Torque Transducer Characteristic](image)

Figure 26: 500 in.lb torque transducer calibration characteristic.

3.3.3.2 Results

The results of the torque versus current test shown in Figure 27, exhibit good linearity and only a small amount of saturation above 80 amps. The torque versus armature current
characteristic corresponds to a measured machine constant of 0.455 Nm/A up to 80A and reduces to 0.437 Nm/A at 100 amps noting that the test was performed using the peak torque position. This machine constant is in good agreement with the peak-voltage constant determined in the previous section and confirms a much lower than expected flux density in the machine.

![Graph of Measured Torque Constant vs Average Armature Current](image)

**Figure 27:** Measured torque versus armature current under stalled conditions with a 2mm air gap distance.

### 3.3.4 Efficiency versus Speed & Torque

This experiment provides a valuable tool in examining the entire operating region for efficiency peaks and troughs. The results affect both research focus and solar racing car strategy.

#### 3.3.4.1 Measurement Method

The measurement of the combined motor/controller efficiency is based upon the input DC voltage and current compared to the output shaft speed and torque. To account for motor controller losses, the calorimetrically measured efficiency of the motor controller, detailed in [54], is subtracted according to the operating speed and output torque. DC Input voltage and current are measured with an accuracy of ± 0.05 V and ± 0.05 A respectively. Output shaft torque and speed are measured with an accuracy of ± 0.28 Nm and ± 2.5 rpm respectively.
3.3.4.2 Results

The measured efficiency map of the AFM with the air gap set at 2mm is shown in Figure 28. This map has some significant measurement deficiencies and, in that particular test, the temperature of the EMC heat sink limited the maximum speed at which measurements were taken to 800 rpm.

![Measured Motor Efficiency Map](image)

Figure 28: Measured motor efficiency versus torque and speed with a 2mm air gap.

It is important to note that the measurement of shaft torque is particularly uncertain. After the series of measurements were taken, a torque measurement offset was observed. This offset was approximately 0.5 Nm after measuring it at several shaft positions with no shaft load present. By simply applying this offset as a constant to the collected data it is apparent that the transducer gain also differs from that obtained by calibration as efficiency figures greater than 100% are calculated in the higher speed region of the dataset. The origin of this uncertainty is believed to lie with the silver slip-ring contacts in the transducer that, when later examined, show severe signs of wear. Anecdotal evidence collected from past users of the transducer suggests that the errors noted above were typical. Time and financial constraints did not allow the repetition of this experiment with another torque transducer to reduce the apparent error.
In addition to the uncertainty noted at the time of the experiment, later work measuring the “spinning loss” of the machine, described in sections 3.3.5 and 3.4.5, allows for the calculation of loss versus speed and torque. Using the measured values for spinning loss, armature resistance and torque constant, an efficiency map can be calculated and is presented in Figure 29. The calculated map only factors in copper loss, bearing loss and the magnetic loss caused by the rotation of the machine, it doesn’t include the magnetic loss caused by armature excitation. That mechanism is discussed in more detail in chapter 4. It is important at this point to define the two types of current that are observed to flow in the machine windings. That which flows in and out of the machine terminals, and that which flows in ‘eddies’ within the windings due to the changing magnetic fields. These current components are referred to as terminal current and winding eddy current respectively. Winding eddy currents are discussed in more detail in section 4.4.2.3.1.

Of the differences between the measured and calculated efficiency maps, the greatest is the trend seen in the measured characteristic above 600 rpm, where the efficiency drops off significantly with speed and is almost independent of the applied torque. To clarify the differences, a plot of the difference between measured and calculated is shown in Figure 30.
It is believed that a substantial part of the difference shown in Figure 30 is a result of the torque measurement error and hence the calculated dataset is considered to more accurate.

3.3.5 Spinning Loss
The energy required to turn the shaft of a machine by an external source is often referred to as the spinning loss. This loss includes bearing friction, windage and magnetic losses in the rotor and stator. The test used to measure the spinning loss is a ‘no-load driven test’. The DC machine on the dynamometer was used to drive the CDUSCM through the Staiger Mohilo torque transducer. With torque measured to an accuracy of ±0.5% and speed to an accuracy of ±1%, the resulting loss measurements have a theoretical accuracy of 1.505%. The measured results of the spinning loss test with a 2.0mm air gap are shown in Figure 31. The experimental data measured in both the forwards and reverse directions shows that the measurement error is about 1.6% at 1000rpm.
The loss at this air gap distance can be approximated by the relationship in equation (2).

\[ P_{spin} = 2.46 \times 10^{-3} \omega^2 + 3.72 \times 10^{-1} \omega \]  

Operating under rated conditions this loss represents about 1.4% of the input power.

Bearing loss, windage loss and the magnetic losses will be described in more detail in chapter 4. It is important to note here that the magnetic field change inducing the magnetic loss measured in this test is caused by two factors, the rotation of the permanent magnet field and the reluctance change caused by presence of the winding slots.

### 3.3.6 Bearing & Windage Loss Separation

To check the assumption that the windage and bearing loss components are negligible, two experiments were performed using an AFM rotor with no magnets and a rotor constructed with balsa wood pieces in place of the rare earth magnets. The experimental apparatus noted in section 3.3.5 was used to undertake these tests with the aim of measuring the spinning loss in the presence of zero magnetic potential. Without altering the air gap adjustment mechanism from the nominal 2mm position, the two modified rotors were assembled with the stator and driven by the DC machine while the shaft torque and speed were measured. Figure 32 shows the measured results that exhibit a strong constant-torque type loss. This loss can be approximated by a constant torque of
104 mNm and is most likely to be a result of the torque transducer bearing seal on the motor side of the transducer. Though not specified in the transducer manufacturer’s documentation, data from the SKF General Catalogue suggests that the moment produced from the bearing seal ought to be about 95 mNm assuming that a seal is present on both sides of the bearing.

The measured windage and radial bearing loss component can be approximated by equation (3).

\[ P_{\text{bearing-}R+aero} = 1.133 \times 10^{-6} \omega^3 - 6.236 \times 10^{-4} \omega^2 + 1.693 \times 10^{-1} \omega \]  

3.4 Machine Characteristics versus Air Gap Variation

A major feature of the single-sided AFM is the possibility of air gap adjustment without the need for machine disassembly. The company responsible for commercialising the 1993 design, New Generation Motors Corporation, has even developed a mechanism to perform this adjustment ‘on-the-fly’ [56]. This field weakening technique not only affects the machine constant and hence efficiency, but also other parameters such as cogging torque, inductance, and bearing losses.
3.4.1 Air Gap Adjustment Mechanism

The adjustment on the CDUSCM prototype is only possible when the machine is off-line and by way of a short hex-headed shaft at the rear of the motor. The short shaft drives a 13:79 gear reduction whose output turns an internally threaded sleeve of 1mm pitch. The threaded sleeve then supports the rotor bearing assembly. Overall, an adjustment of 0.1646 mm per turn is delivered by the mechanism that is pictured in Figure 33. The remainder of this chapter presents the results of experimental work to evaluate the sensitivity of this parameter upon the characteristics of the machine. The limits of the adjustment are bound by the axial limits of the adjusting mechanism. The operating range of the air gap is between 1.0 mm and 4.0 mm.

![Figure 33: Photo showing the air gap adjustment mechanism. Parts of this photo have been deliberately obscured to protect commercial knowledge.](image)

3.4.2 Armature Inductance

Laboratory experiments have shown that the sensitivity of the average armature inductance due to air gap distance change to be quite small compared to that of the inductance variation over 180°. Figure 34 shows both the absolute inductance averaged over 180° and the inductance variation observed over the same displacement.
These results suggest that significant magnetic saturation is present at air gap distances below 2.0 mm because the rate of increase of average inductance decreases with air gap distance. The alternating component of the inductance is seen to increase as the air gap distance is reduced, though the rate of increase of this component with air gap distance is observed to reduce at smaller air gap distances.

Indicative tests have shown that the armature inductance reduces at higher frequencies. The skin effect of the materials used in construction would suggest that this change in inductance is expected. As a quantitative example, the average inductance measured at 400 kHz with a 2.0 mm air gap is 55μH. In contrast to the results at 1 kHz, the average inductance at 400 kHz is observed to increase as the air gap distance is increased. Though this frequency is an extreme value, there is the suggestion that the inductance driven by the higher frequency modulation of some EMCs will be less than the measured results presented here. This reduction in inductance with frequency will affect the operating frequency or current ripple of EMCs with higher modulation frequencies.

Further discussed in Chapter 5, the modulation frequency has a significant effect upon the electromagnetic losses in the machine. This would suggest that an increase in air gap distance not only decreases the average magnetic field in the machine and increases
armature current for the same power output, but also increases electromagnetic loss through an increase in armature modulation frequency.

3.4.3 Armature Voltage at 1000 rpm

The average of the 3-phase rectified armature voltage or back-EMF was measured at 1000 rpm with a range of air gap distances. The EMF constant was then calculated and compared. The results shown in Figure 35 show a linear sensitivity of approximately $-0.061 \text{ V.s.rad}^{-1}\text{.mm}^{-1}$ for air gap distances less than 3mm.

![Measured EMF Constant vs Air Gap Distance](image)

*Figure 35: EMF Constant versus air gap distance derived from the measured average 3-phase armature voltage at 1000 rpm.*

3.4.4 Spinning Loss

The viability of the mechanical flux weakening technique in a high efficiency application relies upon the magnetic losses decreasing as the copper losses increase. The spinning loss test is one method of determining the change in magnetic losses versus the air gap distance. Figure 36 shows the spinning loss measured at four different air gaps and in both the forwards and reverse directions.
Figure 36: Spinning loss measurements at 1.0, 2.0, 3.0 & 4.0 mm.

This dataset allows an empirical relationship to be derived for spinning loss at any speed or air gap distance within this range. Equation (4) describes this relationship within 2 watts of the experimental data.

\[
P_{spin} = \left( 3.538 \times 10^{-4} d_{ag}^2 - 2.789 \times 10^{-3} d_{ag} + 6.681 \times 10^{-3} \right) \omega^2 + \\
\left( 1.234 \times 10^{-2} d_{ag}^2 - 1.375 \times 10^{-1} d_{ag} + 6.064 \times 10^{-1} \right) \omega 
\] (4)

3.5 Summary

The laboratory testing of the variable air gap CDUSCM has demonstrated that there is a significant difference between the current machine characteristics and its original performance. From the initial tests carried out in 1995 [26], delivering results within 6% of the design specification and the results presented here, there is little doubt that the characteristic of the permanent magnet material used in the machine has changed since the machine was manufactured. The air gap flux density is 78.5 % of what is was when measured in 1995 [26].

The measured parameters of the variable air gap CDUSCM are summarised in
Table 10 along with their sensitivity to air gap distance where applicable.
It has been shown that the energy required to spin the rotor of the machine is dominated by the loss in the axial bearing and the magnetic loss incurred by the moving permanent magnet field. In addition, the loss incurred by radial bearing friction and windage is less than 0.26% under rated conditions.

When predicting the total loss in the machine by adding the spinning loss and copper losses, a significant deviation from the measured efficiency exists above 600 rpm and is relatively independent from torque. This deviation is believed to be a result of an increase in magnetic loss due to armature currents and is discussed further in chapter 4.

Table 10: Summary of Measured CDUSCM Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Nominal Value @ 2.0mm Air Gap Distance</th>
<th>Sensitivity to Air Gap Distance</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC Armature Resistance</td>
<td>28.7 mΩ</td>
<td>n/a</td>
</tr>
<tr>
<td>Average Armature Inductance @ 1kHz</td>
<td>201.9 μH</td>
<td>-4.9 uH/mm</td>
</tr>
<tr>
<td>Average Air Gap Flux Density</td>
<td>0.565 T</td>
<td>-0.081 T/mm</td>
</tr>
<tr>
<td>Maximum Static Torque</td>
<td>0.83 Nm</td>
<td>0.06 Nm/mm</td>
</tr>
<tr>
<td>Machine Constant</td>
<td>0.427 V.s/rad</td>
<td>-0.061 V.s/rad.mm</td>
</tr>
</tbody>
</table>
Chapter 4

DETAILED DESIGN ANALYSIS BY 2D FINITE ELEMENT ANALYSIS

4.1 2D Analysis Introduction
Since the original design was derived, a number of more recent numerical analysis techniques have been refined and introduced into commercial software. These tools have allowed an unprecedented look into the behaviour of the electromagnetic circuit in the AFM. In particular, transient analysis allows the modelling of non-sinusoidal excitation and also the non-sinusoidal behaviour of magnetic fields due to magnetic saturation and geometry.

4.2 2D Modelling Assumptions
By using the 2D transient analysis capability, the assumptions that the electromagnetic circuit is static or that it behaves in a harmonic manner can be removed. The remaining assumptions still required include:

- A mid-radius cylinder is a good 2D representation of the 3D geometry.
- Magnetic flux lies purely within the 2D plane modelled.
- Winding end-effects are negligible.
- Stator laminations are loss-less. (This is discussed further in section 4.4.2)
- No flux exits or enters the surface of the machine envelope in the z-direction.

4.3 Cogging Torque
It is possible to calculate the zero-speed cogging or a static torque waveform of the machine using multiple 2D magnetostatic models each representing a slightly different rotor position. The results are useful though the process is time consuming as each geometric set up needs to be meshed, refined and solved separately. The initial set-up of the transient solution involves the generation and refinement of one mesh used at every time-step of the transient solution. In addition, each successive time-step uses the solution of the previous step to generate the next solution.
As part of the transient problem set-up, the force or torque on the ‘moving’ geometry in a problem can be nominated for calculation at each time-step. Figure 37 contrasts the results from the transient analysis of the CDUSCM and the measured results from section 3.2.4. The transient solution set-up, models a motor speed of 10 rpm, which corresponds to an electrical frequency of 1 Hz. This low frequency results in very little eddy current losses that also affect the torque calculated from the FEA.

![Image of Transient Cogging Torque at 10 rpm](image)

**Figure 37: Cogging torque calculated from transient analysis compared to measured results.**

By observation, the FEA results show a strong $12^{\text{th}}$ harmonic whereas the measured data exhibits a dominant $6^{\text{th}}$ harmonic. The results of the transient analysis also show a $6^{\text{th}}$ order sub-harmonic that is a result of FEA model geometry, which can be removed by further refining the FEA mesh. However, overall the agreement in amplitude is very good.

The occurrence of torque amplitude peaks is the result of the interaction of the permanent magnets and the stator slots. The $12^{\text{th}}$ harmonic is generated by separate interactions of the leading and trailing edges of the magnets, which have an arc width unequal to a whole multiple of the slot pitch. For a machine with a $120^\circ_{\text{e}}$ magnet arc width, the magnet edges coincide with every $3^{\text{rd}}$ slot and have a cumulative effect at the $6^{\text{th}}$ harmonic frequency. By increasing the magnet arc width from $120^\circ_{\text{e}}$ to $128.2^\circ_{\text{e}}$, the $6^{\text{th}}$ harmonic caused by the leading edges can be used to partially cancel the $6^{\text{th}}$ harmonic
caused by the trailing edges. This minimisation of peak cogging torque is useful in applications that require low torque ripple or for generators such as wind turbine generators, which require a low starting torque.

A further explanation for the difference in the waveforms in Figure 37 can be derived from examining the effect of magnet arc width upon the cogging torque waveform. Figure 38 shows the results of four transient FEA simulations using different magnet arc widths expressed in electrical degrees.

![Figure 38: Cogging torque calculated by 2D transient simulation at various magnet arc widths.](image)

Of particular note is the sensitivity of the torque waveform to the arc width. The arc width difference of 1.8°e between the optimised 128.2°e and 130°e leads to more than a doubling of peak torque amplitude. The 1.8°e difference is equivalent to the dimensional tolerances of 0.12, 0.20 and 0.25mm at the inner, mid and outer radii respectively. Examining this sensitivity with respect to the measured tolerance of the manufactured slots from section 3.2.1, it is possible to conclude that the measured wave shape is a result of a distribution of magnet arc widths in the vicinity of the optimised value and the imperfect placement of individual magnets.

The examination and reduction of cogging torque using the rotor/magnet skewing has not been attempted due to the manufacturing complexity and the fact that it not likely to
reduce electromagnetic losses. Rather, it is likely to lead to the time-distribution of losses and perhaps an increase in average loss due to the electromagnetic loss non-linearities discussed in the next sections of this Chapter. A reduction of cogging torque beyond that achieved by arc-width optimisation is also somewhat academic for the application considered.

4.4 Energy Losses

The major loss components found in an electrical machine can be broken into the broad categories of mechanical and electromagnetic as depicted in Figure 39. These loss components are discussed separately over the remainder of this chapter.

![Figure 39: The major loss components in an electrical machine](image)

(Joule effect losses are also known as copper losses).

4.4.1 Mechanical Losses

The quantification of mechanical losses in the machine is important in ‘loss accounting’ related to laboratory experiments. It has been assumed that the mechanical losses can be broken into the independent mechanisms of bearing friction loss and rotor windage (hydraulic) loss. The dedicated thrust bearing in the AFM is aimed at reducing bearing losses, but they are still significant when compared to the magnetic loss components discussed later in this chapter. Until now, the windage loss component has been ignored.

4.4.1.1 Bearing Friction Loss

The single-sided AFM contains two radial bearings and one thrust bearing. For the purpose of accounting for losses in experiments, the radial bearings are assumed to carry the weight of the rotor, torque transducer and coupling-half, where as the thrust bearing is assumed to carry only the attractive force between the stator and the rotor.
This derivation of frictional bearing loss presented here is based upon the analytical formulae and constants presented in the SKM General Catalogue [57].

The moment due to friction of a bearing is given by:

\[ M = 0.5 \times \mu \times F \times d \]  

where \( M \) is the frictional moment in N.mm
\( \mu \) is the bearing factor,
\( F \) is the bearing load in Newtons and,
\( d \) is the bearing bore diameter in mm.

And the power loss due to this frictional moment is

\[ P_{\text{bearing}} = 1.05 \times 10^{-4} \times M \times n \]  

where \( n \) is the rotational speed in rpm.

The characteristics of the two bearing types used in the CDUSCM are presented in table 11.

<table>
<thead>
<tr>
<th>SKF Designation</th>
<th>Type</th>
<th>( \mu )</th>
<th>( d ) (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>51107</td>
<td>Thrust, ball</td>
<td>0.0013</td>
<td>35</td>
</tr>
<tr>
<td>16006</td>
<td>Deep groove, ball</td>
<td>0.0015</td>
<td>30</td>
</tr>
</tbody>
</table>

The load applied to the deep groove bearings is assumed to be only the radial component of the shaft load. For the dynamic testing described in chapter 3, this is the force due to the mass of the torque transducer and associated coupling hardware. The radial load is therefore \( 9.7826 \text{ m.s}^{-2} \times 7.774 \text{ kg} \), which equals 76.05 N.

Combining this load, equations (5) & (6) and the constants from table 11, the total loss due to the radial bearings can be derived as per equation (7).

\[ P_{\text{bearing-}R} = 1.80 \times 10^{-4} \times n \]  

The axial load used in loss calculations is the average of that calculated by 2D multimagnetostatic FEA. Table 12 summarises the average axial force for different air gaps.
Table 12: Average Axial Force values at different Air Gap Distances.

<table>
<thead>
<tr>
<th>Air Gap Distance (mm)</th>
<th>Average Axial Force (N)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.0</td>
<td>5504</td>
</tr>
<tr>
<td>2.0</td>
<td>4132</td>
</tr>
<tr>
<td>3.0</td>
<td>3019</td>
</tr>
<tr>
<td>4.0</td>
<td>2293</td>
</tr>
</tbody>
</table>

Using the Least Squares Regression method with the data in table 12, equation (8) describes the average axial force versus air gap distance.

\[
F_{\text{axial}} = 161.2d_{ag}^2 - 1880.6d_{ag} + 7229.3 \quad (8)
\]

By combining equations (5), (6) & (8), the loss due to the axial bearing can be related to the air gap and rotational speed as shown in equation (9).

\[
P_{\text{bearing-\text{A}}} = 2.389 \times 10^{-4} \times (1.612d_{ag}^2 - 18.806d_{ag} + 72.293) \cdot n \quad (9)
\]

The total bearing loss is obtained from the addition of the axial and radial bearing losses and can be represented by a surface dependant on the air gap and rotational speed shown in Figure 40.

![Total Bearing Loss versus Speed & Air Gap Distance](image)

Figure 40: Total bearing loss related to rotational speed and air gap distance.

At an air gap distance of 2mm the total bearing moment is 0.095 Nm. The relative contributions of the radial and axial components at this air gap distance are 1.8% and 98.2% respectively. Before drawing any conclusions on this result, it is worth analytically...
evaluating the windage loss component, as it is difficult to separate it from the bearing loss component in experiments.

4.4.1.2 Windage

The loss contribution from the air movement induced by the rotor is not accurately predictable from analytical equations though an approximation can be made. On the inside of the machine, the square edges of the magnets and surface friction on the rotor act to rotate the air volume in the air gap. The induced centrifugal force on the air volume causes a radial pressure difference though no net air flow can occur as there is no entry point for air into the centre of the machine. This condition ought to produce a lower windage loss than that which occurs on the outside surface where airflow is relatively free. On the outside of the machine, the rotor surface can be approximated by a finite diameter disk rotating in a fluid as described in Schlichting’s definitive boundary layer theory text [58].

Summarising the procedure described in [58], the moment, $M$, for a disc wetted only on one side is described by equation (10) assuming a purely laminar flow.

$$ M = 2C_M \rho \omega^2 R^5 \quad \text{(10)} $$

where $C_M$ is the dimensionless moment coefficient; $C_M = \frac{3.87}{\Re}$,

$\Re$ is the Reynolds number defined by the disk radius and tip velocity; $\Re = \frac{\omega R}{\nu}$,

$\rho$ is the air mass density,

$\nu$ is the kinematic viscosity of the fluid,

$\omega$ is the angular frequency of the disk and,

$R$ is the radius of the disk.

The values of mass density and kinematic viscosity for air at 25°C used in calculations are 1.1871 kg/m$^3$ and 1.5967 x 10$^{-5}$ m$^2$/s respectively.

From equation (10), the power used to move the fluid can be determined by (11).

$$ P_{\text{wetted disk}} = \frac{7.74}{\sqrt{\frac{R^2 \omega}{\nu}}} \rho \omega^3 R^5 = 7.74 \sqrt{\nu} \rho \omega^{2.5} R^4 \quad \text{(per disk side)} \quad \text{(11)} $$
Substituting the fluid constants for air and the outside radius of the CDUSCM rotor, equation (12) describes approximate loss for the outside surface of the CDUSCM rotor.

\[ P_{\text{aero (outside)}} = 1.0485 \times 10^{-5} \omega^{2.5} \]  

Even if this loss were doubled to take into account the potential windage internal to the air gap, the windage loss at 1000 rpm would be due to a shaft torque of 22.5 mNm.

4.4.1.3 Mechanical Loss Summary

Bringing together the results from sections 4.4.1.1, 4.4.1.2 and 3.3.6, Figure 41 shows the calculated bearing and windage losses against the measured radial bearing and windage losses. Of the measured loss, most is attributed to the bearing seals on the torque transducer and not the CDUSCM bearing and windage. The measurement resolution of the Staiger Mohilo torque transducer is insufficient to make definitive statements about the variation in loss versus speed, though the overall trend does not seem to be effected by an aerodynamic type loss as equation (11) might suggest. This leads to the conclusion that equation (11) is likely to be an overestimate of the aerodynamic losses throughout the measurement range. Additional uncertainty exists in the windage calculations as the Reynolds number calculated for this fluid dynamics system suggests that the fluid flow is in the transitional region between laminar flow and turbulent flow.

Analytically, the total loss due to the radial bearings and aerodynamic loss is in the order of 1W at 1000 rpm.
4.4.2 Electromagnetic Losses

Electromagnetic losses induced in the CDUSCM are found primarily in the stator and rotor iron. Additional electromagnetic loss is caused in the permanent magnets and armature current causes ohmic loss in the copper windings.

4.4.2.1 Low Frequency Magnetic Loss Modelling Theory

There are two broad categories of structures found in the magnetic circuit of the AFM, bulk and laminated. Bulk structures are those whose X, Y & Z dimensions are generally within the same order of magnitude, whereas laminated structures are those that have one dimension less than the other two, by an order of magnitude or more. In addition, laminated structures are used to reduce eddy current losses where medium frequency, high induction fields are found.

The modelling of low to medium frequency magnetic losses has typically been focused on laminated structures because of their use in transformers and medium to high-speed machines. The use of bulk materials in magnetic circuits is generally limited to either very low frequency or DC magnetic circuits such as those excited by permanent magnets. The loss found in those low frequency circuits is largely observed to be negligible with respect
to other parts of the machine. It is for these reasons that the calculation of magnetic loss in bulk (or non-laminated) materials is rarely reported in the literature.

Regardless of the physical structure, magnetic losses are generally separated into the three loss components, hysteresis, eddy current and anomalous loss. The separation of these loss components can be considered for both types of structures independently.

### 4.4.2.1.1 Laminated Structures

The analytical calculation of electromagnetic losses in laminated ferromagnetic materials has been the focus of research for over a century, however certain aspects such as hysteresis loss modelling still attracts the attention of academics. The calculation of electromagnetic losses is somewhat varied in its approach including efforts based upon the Steinmetz equation [59-61], the Jiles-Atherton model [62-65], and the Preisach model [66-68]. More recently, the difficulty in modelling transient hysteresis seems to have been addressed [61], and a solution to modelling both alternating and rotating magnetic fields has been proposed by Guo et. al [69]. Since the academic community is still yet to agree upon an accurate material loss model, it is futile to attempt to calculate accurate loss totals purely from commercial FEA software. However, this uncertainty does not prevent the approximation of loss nor does it prevent the examination of loss variation caused by design parameter changes.

It is generally accepted that the loss in a ferromagnetic material is proportional to the rate of change of flux. The Steinmetz-based time domain formulation for core loss, given by equation (13), from [61] is a good example of this.

\[
p(t) = H_{irr} \frac{dB}{dt} + \frac{1}{2\pi^2} k_{\text{eddy}} \left( \frac{dB}{dt} \right)^2 + \frac{1}{C_a} k_{\text{anom}} \left| \frac{dB}{dt} \right|^{1.5} \tag{13}
\]

where \( H_{irr} \) is the irreversible hysteresis component,
\( C_a \) is a constant evaluated to 8.763363 and
\( k_{\text{eddy}} \) and \( k_{\text{anom}} \) are constants obtainable from either experimental data or manufacturers material loss data under sinusoidal excitation.

The three terms on the right-hand side of equation (13) are due to hysteresis, eddy current and anomalous (or excess) losses respectively. This loss separation model has been tested with success by other authors [61, 70] for ferromagnetic laminations in motors, including
Lycore-140™, and a similar form of the equation is used here to model the losses in the stator laminations of the CDUSCM.

The 3D modelling of laminated structures using the finite element method is cumbersome due to the relatively small lamination thickness compared to the majority of the model geometry and also due to the effective anisotropic conductivity offered by the laminated structure. The former of these issues causes difficulties in preparing effective meshes and obtaining numerical convergence, the later issue prevents the accurate accounting of losses in 2D models due to FEA limitations. For these reasons, the preferred method of loss accounting in laminated structures is still through the use of empirically based analytical methods such as that described earlier in this section. Laminated structures can be modelled in the 2D environment by assuming that their conductivity is zero.

4.4.2.1.2 Bulk Materials

The analytical loss determination for laminated structures relies heavily upon published loss measurements, usually by manufacturers, for a particular lamination thickness. This dependence prevents the complete generalisation of those methods to bulk materials. From equation (13), it can be concluded that all of the loss components are related to the rate-of-change of flux density. It can therefore be said that if a machine component is subject to a lower rate of change of flux density then the reduction in total bulk material loss should be observable by monitoring an individual loss component, if the geometry of that bulk material remains constant. This is important as 2D FEA provides an accurate method of calculating eddy current loss in bulk materials.

The uncertainty in calculating the total magnetic loss is exacerbated by the lack of information pertaining to the loss characteristics of bulk materials under pre-magnetisation. Pre-magnetisation is a condition effecting losses in the rotor where the majority of the bulk materials exist in the CDUSCM. The effect of pre-magnetization, at least in ferrite materials, has been shown to significantly increase magnetic losses [71]. This uncertainty particularly effects the determination of hysteresis loss component. The calculation of the eddy current loss component on the other hand is unaffected by pre-magnetisation as this loss mechanism relies largely on geometry and the conductivity of the material, which are known for a given circumstance.

The available 2D FEA software only fails in determining eddy currents where the conductivity is anisotropic in the z-direction. This anisotropy, typical of laminated cores,
cannot be accounted for in the 2D environment when modelling the plane of lamination. In the environment of laminated structures, it is most beneficial to use 2D FEA to establish the flux trajectories and use the available analytical techniques noted in the previous section.

4.4.2.2 Flux Trajectories

To obtain a proper understanding of the magnetic fields within the machine, the seven points depicted in Figure 42 were selected at which to observe flux vectors while rotor movement was modelled.

![Figure 42: 2D Model locations for magnetic flux trajectory observation.](image)

The flux trajectories for these points were derived from 2D magnetostatic FEA and are given for the stator points and rotor points in Figure 43 and Figure 44 respectively.
Figure 43: Flux trajectories (a), (b), (c), (d), (e) and (f) for the positions A, B, C, D, E & F respectively. Units of flux density are Telsa.
Of particular interest is the effect of the stator slots upon the normally elliptical loci found in the stator (Figure 43).

Due to mesh density variations, the trajectories of points B, F, G & H are of a lesser accuracy than those at the other points. Knowing the nature and magnitude of the flux at these representative points, the losses in each part of the machine can be considered with a more complete knowledge.

4.4.2.3 Stator

Stator Losses are assumed to be composed of Joule effect losses due to current flow in the copper windings and core losses in the stator iron due to magnetic field changes. The change in the magnetic field in the stator is a complex interaction contributed to by permanent magnet field movement, changing armature currents and reluctance variations caused by the presence of the winding slots and rotor movement. Each of these magnetic field change mechanisms causes variations at different rates with respect to time. With the rotor moving at 1000 rpm the permanent magnets induce a fundamental magnetic frequency in the stator of 100 Hz. The frequency at which the slots cause reluctance variations is therefore 600 Hz at the same speed. The armature inductance, the applied DC voltage from the EMC and the rotor speed dictate the rate of change of current and hence rate of change of the magnetic field due to that current. The nominal DC voltage of 48 volts and the average measured inductance of 202$\mu$H have the potential to induce a maximum rate of change of current of 237.6 kA/s. The actual rate of change of current varies with rotor speed due to both the rising back-EMF as well as the switching scheme of the EMC employed with the CDUSCM as mentioned in chapter 1. While it is difficult

Figure 44: Flux trajectories (a) and (b) for positions G & H respectively. Units of flux density are Telsa.
to inspect all of these mechanisms together due to material non-linearity and the limitations of the available FEA solvers, it is possible to gauge their relevance upon machine loss individually.

4.4.2.3.1  Copper Loss

For direct current, Ohm’s Law and the measured line-to-line armature resistance are all that is needed to calculate the total copper loss. For time varying electric currents, the formation of eddy currents in the large conductors used in the CDUSCM produce a significant distortion in the conductor current density and hence increases the copper loss. Analytically, the determination of the eddy currents in the rectangular copper conductors under non-sinusoidal excitation is difficult, though Hanselman has derived a solution for the eddy currents in square-section, slot-bound conductors under sinusoidal excitation [72].

At 100 Hz, the skin depth of copper reaches 6.7mm, which is half of the maximum dimension of the conductor cross-section. To evaluate the eddy current loss within the windings, 2D harmonic and 2D transient analyses were performed using an FEA model with 6 rectangular conductors in the central slot as shown in Figure 45.

For the harmonic analyses, each conductor is modelled carrying a current amplitude of 5 amps. The peak power loss density resulting from the analyses at 100 Hz, 1kHz and 5 kHz are shown in Figure 46 & Figure 47 respectively. Post processing of the harmonic results gives rise to average conduction loss figures of 0.081, 0.936 and 2.278 watts per meter slot length at the same respective frequencies. Contrasting these numbers with the terminal current induced power loss of 0.080 W/m it can be concluded that the effect of current change at the commutation frequency of up to 100 Hz is negligible, while the modulation of current over several hundred hertz has a significant effect upon the effective armature resistance and resultant power loss. The results of the harmonic analyses also agree generally with Hanselman’s analytical work for square section conductors of an equivalent area.
Figure 45: 2D FEA model showing the 6 conductors in the central slot with an optimised mesh.

Figure 46: Power density plot showing the 6 conductors per slot with a current amplitude of 5 amps at 100 Hz
To study the more realistic non-sinusoidal current effects upon eddy currents, several 2D transient analyses were performed using the same geometry as the harmonic analysis. The selection of the current waveforms is based upon a common operating point of the CDUSCM and its EMC, which is an average current of 20 amps and a switching frequency of 1 kHz. This operating point corresponds to a vehicle speed of about 70 km/h. Two alternating current components were modelled, the first a 10A peak-to-peak symmetrical triangular current waveform and the second a 10 A peak-to-peak sinusoidal current waveform. Both alternating components were modelled with a frequency of 1 kHz. In addition, the alternating component was modelled with and without a DC offset of 20 amps. Table 13 presents a summary of the average power lost due to both eddy currents and the terminal current calculated from the 2D transient data and the terminal current loss.
Table 13: Summary of copper loss per metre slot length (W/m) under various current waveforms.

<table>
<thead>
<tr>
<th>AC Component</th>
<th>2D Transient Calculated Loss</th>
<th>Analytical Terminal Current Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>DC Component (amps)</td>
<td>DC Component (amps)</td>
</tr>
<tr>
<td></td>
<td>0</td>
<td>20</td>
</tr>
<tr>
<td>10A&lt;sub&gt;p-p&lt;/sub&gt; Sinusoidal</td>
<td>1.381</td>
<td>0.080</td>
</tr>
<tr>
<td>10A&lt;sub&gt;p-p&lt;/sub&gt; Triangular</td>
<td>0.918</td>
<td>0.053</td>
</tr>
</tbody>
</table>

Figure 48 and Figure 49 show the total transient loss in the conductors for the current waveforms without and with a DC offset respectively. Figure 50 shows the resulting power density plots near the time of peak loss for each of the current waveforms analysed.

![Transient FEA Copper Loss](image)

*Figure 48: Copper loss versus time derived from 2D transient FEA showing losses under both sinusoidal and triangular currents.*
Transient FEA Copper Loss
Averages: Sine 3.926 W/m, Triangular 3.478 W/m
(CDUSCM w/ 2.8 x 5.8mm conductors)

Figure 49: Copper loss versus time derived from 2D transient FEA showing losses under both sinusoidal and triangular currents with a 20 A DC offset.
In summary the increase in copper loss, above that produced by the terminal current, due to eddy currents is significant at about 34% assuming a 10A current band offset by 20A and switching at a frequency of 1 kHz. At the same operating point, 2D transient FEA predicts a total copper loss of 10.8 W, though this assumes that the eddy current
density remains constant over the length of the conductor including the significant length of the winding end-turn where the conductors are not bound by the slot. Further work is required to ascertain the variation that the end-turn current distribution will create, though the loss values presented here are most likely to be a worst-case scenario.

4.4.2.3.2 Lycore\textsuperscript{TM} 140 Loss

To allow comparison with the spinning loss test results in chapter 3, the armature-induced field is ignored until further discussed in section 4.4.2.5.

The magnetic flux behaviour in the stator iron is dominated by a rotational flux component. This component is distorted by the trapezoidal flux seen in the stator teeth and is also reduced in the y-component due to the higher reluctance slots. The stator is constructed of a laminated material whose bulk properties, in particular conductivity, are difficult to represent faithfully in a 2D model and the available software. Given this fact, it is preferable to use the empirical method described in section 4.4.2.1.1.

A commonly accepted method for loss estimation under sinusoidal flux density is the “Watts per kilo” or “loss separation” method. This method has been also shown to produce acceptable results for non-sinusoidal flux waveforms [70]. The loss separation method assumes the material loss relationship is of the form shown in equation (14).

\[ P_{em} = P_{hys} + P_{eddy} + P_{anom} \]  

(14)

where \( P_{hys} \) is the hysteresis loss,

\( P_{eddy} \) is the classical eddy current loss and

\( P_{anom} \) is the loss generally attributed to domain wall motion.

It is accepted that the hysteresis loss approximately obeys the Steinmetz law;

\[ P_{hys} = C_h B_{pk}^n f \]  

(15)

where \( C_h \) and \( n \) are empirical constants dependant upon material characteristics,

\( f \) is the induction frequency and

\( B_{pk} \) is the peak flux density induced in the material.

And that the classical eddy current loss in thin laminations can be calculated by;

\[ P_{eddy} = C_e \left( fB_{pk} \right)^2 \]

- 1.
where $C_e = \frac{\pi^2 b^2 \sigma}{6 \rho}$ is the coefficient of eddy current loss,

- $b$ is the lamination thickness,
- $\sigma$ is the bulk material conductivity and
- $\rho$ is the bulk material mass density.

Finally, the anomalous loss can be given by;

$$P_{anom} = C_a \left( f B_{pk} \right)^{1/2}$$

where $C_a$ is a theoretically constant coefficient determined by empirical methods.

These relationships lead to a loss per cycle characteristic similar to that shown in Figure 51.

Figure 51: Material loss component separation for Lycore 140.

To account for non-sinusoidal fields, both $B_x(t)$ and $B_y(t)$ can be written as a Fourier series and treated independently. This relies on the principle of superposition that in turn assumes that the relationship between core loss and flux density is linear, which equations (15), (16) & (17) show isn’t strictly the case. The Fourier series expansion can however be used to give a better approximation than from using the fundamental component alone.
The major limitation of this method is the need for empirical constants for the material under consideration. In addition, other researchers have suggested that the capability of the model to account properly for rotational hysteresis could be improved. Zhu, et. al. have derived the appropriate constants for the 0.35mm grade of the Lycore-140™ material [70] as summarised in table 14.

Table 14: Loss Coefficients of Lycore-140 obtained by Zhu, et. al. [70]

<table>
<thead>
<tr>
<th>$C_h$</th>
<th>$n$</th>
<th>$C_e$</th>
<th>$C_a$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.0192</td>
<td>1.79</td>
<td>$6.54 \times 10^{-5}$</td>
<td>$6.38 \times 10^{-4}$</td>
</tr>
</tbody>
</table>

4.4.2.4 Spinning loss estimation

4.4.2.4.1 Stator spinning loss

The flux trajectories shown in Figure 43 & Figure 44 suggest that it is difficult to use a peak flux density common for all sections of the Lycore volume. On the other hand, the commercial nature of the software available doesn’t allow the integration of this material loss information to calculate the loss contribution for each element in the finite element mesh. To allow an identification of losses in the machine with some accuracy, the Lycore volume is divided up into 2 volumes, the tooth volume and the back-iron volume. For the tooth and back-iron volumes it is assumed that the change in flux density due to rotor rotation is sinusoidal with a peak induction noted in table 15. A higher frequency component is also included in the loss calculations for the slot induced flux variation also noted in table 15.

Table 15: Values used for Lycore-140 spinning loss calculations.

<table>
<thead>
<tr>
<th>Volume</th>
<th>Bpk-x (T)</th>
<th>Bpk-y (T)</th>
<th>Volume ($x 10^{-4}$ m$^3$)</th>
<th>Mass (kg)*</th>
<th>Frequency (Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator Teeth</td>
<td>0.4</td>
<td>0.9</td>
<td>3.994</td>
<td>3.055</td>
<td>100</td>
</tr>
<tr>
<td>Stator Back-iron</td>
<td>0.4</td>
<td>0.75</td>
<td>5.608</td>
<td>4.290</td>
<td>100</td>
</tr>
<tr>
<td>Entire Stator</td>
<td>0.25</td>
<td>0</td>
<td>9.598</td>
<td>7.345</td>
<td>600</td>
</tr>
</tbody>
</table>

* mass is calculated using material density under ISO standard conditions.

With no armature current and a machine speed of 1000 rpm, the loss separation method and the values in table 15 results in a total Lycore loss of 38.9 watts.
4.4.2.4.2  Rotor spinning loss

The CDUSCM rotor is comprised of 12 sector shaped magnets bonded to a steel rotor. The variable air-gap machine uses a silicon steel rotor, whilst the other existing prototypes and the FEA models use a mild steel rotor. These simple components are considered bulk materials because they are single piece entities. In the 2D modelling environment, it is possible for significant currents to flow in the 3rd dimension that is modelled well by the 2D FEA.

The magnetic fields present in the rotor iron and permanent magnets under normal operation consist of a high DC field component with two alternating components superimposed. The alternating components are contributions from the moving stator slots and the non-synchronous armature component. The field variation due to the stator slots is observed in Figure 44 and suggests a flux density variation of up to 0.3 Tesla peak-to-peak. The variation due to the stator slots is six times the fundamental frequency and is therefore 600 Hz when the machine is operating at the nominal 1000 rpm. A significant difference ought to arise between FEA studies and laboratory measurements due to the difference in material properties of the silicon iron used in the machine tested and the mild steel material modelling in the FEA. The addition of silicon to the steel typically reduces its conductivity by a factor of up to 5 while the corresponding increase in hysteresis loss varies up to a factor of 3. Not knowing the actual properties of the silicon steel introduces an uncertainty between the experimental and simulated conditions. Unpublished tests\(^2\) prior to 1998 demonstrated a near zero difference between the spinning-loss tests of a machine with a cast silicon steel rotor and a machine with a mild steel rotor. This result would suggest that the respective increase and decrease of hysteresis loss and eddy current loss were approximately equal or that the magnitude of the rotor loss (over the speed range tested) and hence the difference was relatively small and perhaps immeasurable with the equipment used. To help quantify the loss in the rotor, a 2D transient FEA was performed using the same model as that shown in Figure 12. The results of the transient FEA modelling a rotor speed of 1000rpm shown in Figure 52, has one major peak and one minor peak at a frequency corresponding to the slot frequency. Four points in time representing the peaks and troughs of this waveform were chosen and the post processing of the eddy current loss density at these points reveals the plots shown in Figure 53.

---

\(^2\) The results of these tests have not been published previously or presently due to commercial design sensitivity.
Integrating the eddy current loss densities at $t=0.0056$ s, $t=0.0063$ s, $t=0.0077$ s and $t=0.0119$ s separates the loss contributions from the magnets and the rotor iron as detailed in table 16. The observed spinning loss suggests an approximate average loss per magnet of 0.73 watts and an average rotor iron loss of 4.1 watts.

Simple convection heat transfer analysis suggests that 12.6 W of loss would result in a rotor temperature rise of less than 2°C above ambient temperature reaching a steady state temperature after approximately 9 minutes of operation under these conditions. This temperature rise is not of concern for the CDUSCM as the ambient operating temperature rarely exceeds 65°C. This analysis assumes a rotor-to-air heat transfer coefficient of 25 W/m²K for both sides of the rotor.

Figure 52: Transient rotor loss at 1000 rpm calculated from 2D FEA.
Figure 53: Power loss density at (a) $t = 0.0056$ seconds, (b) $t = 0.0063$ seconds, (d) $t = 0.0077$ seconds and (e) $t = 0.0119$ seconds plotted against scale (c). Units are W/m slot length.

Table 16: Rotor Spinning Loss Composition at $t=0.0056$ s, $t=0.0063$ s, $t=0.0077$ s, and $t=0.0119$ s.

<table>
<thead>
<tr>
<th>Time (seconds)</th>
<th>Power Loss (W)</th>
<th>Time (seconds)</th>
<th>Power Loss (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Rotor Iron</td>
<td></td>
<td>Permanent Magnets</td>
</tr>
<tr>
<td>0.0056</td>
<td>2.45</td>
<td>0.0063</td>
<td>3.08</td>
</tr>
<tr>
<td>0.0077</td>
<td>4.14</td>
<td>0.0077</td>
<td>8.33</td>
</tr>
<tr>
<td>0.0119</td>
<td>5.52</td>
<td></td>
<td>9.14</td>
</tr>
</tbody>
</table>

It is clear that the decrease in machine efficiency due to the spinning loss is small. It could be speculated that the spinning loss in the rotor could be as high as 20 watts by including hysteresis and anomalous losses.

4.4.2.5 Hysteresis Band Induced Losses

The modulation of armature current performed by the EMC is often neglected in accounting for electromagnetic losses in machines. The frequency of modulation is often much higher than the maximum fundamental or commutation frequency in direct drive machines. This is not to suggest that the loss is strictly frequency dependant as the armature current modulation is generally of a triangular nature. With respect to the hysteresis-band current control employed by the CDU EMC, the symmetry of the current modulation and the rate-of-change of current are determined by five factors, the applied DC voltage, the instantaneous back-EMF, the instantaneous armature inductance, the MOSFET on-state resistance, $r_{ds(on)}$, and the instantaneous (effective) armature resistance. Given that the electromagnetic losses are derived from the rate-of-change of flux, it can be deduced that the magnetic loss component derived from armature current...
modulation in the CDUSCM ought to be related to the same five factors. Figure 54 illustrates a typical current waveform produced by the combination of the hysteresis band current controller and the CDUSCM and includes the approximate expressions for the rate-of-change of current. The constant dead-time of the controller switching scheme used is at most 0.5% of the switching period and hence can be disregarded.

Over the period of conduction, the line-to-line inductance changes by less than a few percent and hence it is considered constant for the purpose of determining the rate-of-change of current. In addition, the DC source voltage is assumed constant by way of a large DC input capacitor and the low inductance, low resistance busbar connections throughout the EMC. The total MOSFET on-resistance is approximately 1/3 of the line-to-line armature resistance and by way of a large heat sink is assumed constant. These assumptions leave only the back-EMF and effective armature resistance to effect the rate-of-change of current during operation.

If the armature resistance is considered as constant and we assume that the effective series time constant \( \tau = \frac{L}{R_1} \), is much greater than the switching period, \( T \), then equation (18) describes the average switching frequency, \( f_{sw} \).

\[
\frac{di}{dt} \approx \frac{V_{DC} - E_a}{L} \]

\[
\frac{di}{dt} \approx -\frac{E_a}{L}
\]

Figure 54: Typical armature current waveform induced by the hysteresis-band current controller showing asymmetry and strong linearity.
\[ f_{sw} = \frac{E_a \cdot V_{DC} - E_a^2}{V_{DC} \cdot L \cdot I_{HYS}} \]  

(18)

where \( V_{DC} \) is the driving DC voltage, 
\( L \) is the line-to-line inductance, 
\( I_{HYS} \) is the hysteresis current band and, 
\( E_a \) is the back-EMF.

Further analysis, depicted in Figure 55, shows that the average \( \frac{di}{dt} \) and even the RMS \( \frac{di}{dt} \) exhibit similar characteristics to the switching frequency versus motor speed. This characteristic does little to explain the difference between the calculated and measure efficiency of the machine in section 3.3.4. It does however confirm that the eddy current loss due to current modulation in the motor coincides with the MOSFET switching loss in the EMC that is proportional to switching frequency. To determine the relative magnitude of this loss a transient FEA model was set up using the geometry shown in Figure 56. To remove losses due to reluctance variation, the modelled mechanical speed was kept to near-zero. The armature current was set to a symmetric, triangular wave shape of amplitude 10 A\(_{p-p}\) and at a frequency of 7 kHz. It is expected that this scenario will provide the maximum loss in the rotor corresponding to a speed of 500 rpm.

![Armature Current Modulation Characteristics versus Motor Speed](image-url)

**Figure 55:** Current modulation characteristics of the CDU EMC combined with the CDUSCM.
The results of the transient FEA, shown in Figure 57, indicate that the loss in the rotor and copper due to armature current modulation in significant at an average of 19.3 watts. A break down of the losses at times corresponding to the peaks and troughs of this waveform is given in table 17. The corresponding power loss density plots obtained from post processing are given in Figure 58. The high frequency nature of the armature induced field gives rise to the significant magnitude of the resultant loss even though the magnitude of the field variation is quite small. The flux density variation predicted by the 2D FEA is approximately 0.01 T\text{p-p}. With a high conductivity, permeability and volume the mild steel in the rotor exhibits the smallest skin depth and noticeably the highest total eddy current loss. The skin depth of mild steel at 7 kHz is 0.08 mm. The majority of the loss in the rotor-iron occurs within 0.5 mm of the air-gap surface of the rotor. The loss found in the magnets occurs within 6mm of the leading and trailing edges.

While there is some uncertainty in what the value of the hysteresis and anomalous loss components are, the eddy current loss is itself substantial enough to warrant further consideration.
Figure 57: Transient power loss in the CDUSCM rotor and copper due to a time varying armature current.

Table 17: Armature current induced loss breakdown at \( t = 0.0100175, 0.0100675, 0.0100875 \) & \( 0.0101425 \).

<table>
<thead>
<tr>
<th>time (seconds)</th>
<th>Motor Component Loss (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Copper</td>
</tr>
<tr>
<td>0.0100175</td>
<td>5.9</td>
</tr>
<tr>
<td>0.0100675</td>
<td>12.27</td>
</tr>
<tr>
<td>0.0100875</td>
<td>7.54</td>
</tr>
<tr>
<td>0.0101425</td>
<td>8.20</td>
</tr>
</tbody>
</table>

The rotor loss result is contrasted by the calculated Lycore loss using the loss separation method described in section 4.4.2.3.2. The Lycore-140 loss resulting from a \( 0.005_p \) T variation at 7 kHz is 1.63 watts and indicates the eddy current reducing performance of the laminations in the stator compared to the bulk materials in the rotor.

The overall picture shows that the major loss resulting from the modulation of armature current arises in the rotor iron. By assuming that the eddy current loss varies with the RMS \( \frac{di}{dt} \), an approximation of loss versus motor speed can be made without repeated transient simulations. The outcome of this approximation is illustrated in Figure 59. The copper loss shown in Figure 59 is inclusive of the instantaneous ( \( i(t)^2 \times R \) ) loss. The addition of the eddy current loss in the copper and the terminal current loss results in a
total copper loss that varies largely with average motor current (hence torque) but also significantly with speed as shown in Figure 60.

Figure 58: Power loss density plots showing rotor and copper losses due to time-varying armature current at (a) $t = 0.0100175$, (b) $t = 0.0100675$, (c) $t = 0.0100875$, & (d) $t = 0.0101425$ against scale (e). Units are W/m.
Predicted Armature Current Modulation Losses

Figure 59: Summary of Losses due to armature current modulation.

Figure 60: CDUSCM Copper Loss versus Torque & Speed.

4.5 Summary
The cogging torque of the CDUSCM has been examined experimentally and through the use of static and transient FEA. The CDUSCM exhibits a magnet arc-width dependant
cogging torque whose magnitude is well predicted by transient 2D FEA. The optimum magnet arc-width found for the CDUSCM is 128.2°.

The power lost in the conversion of electrical power to mechanical power in the CDUSCM has been examined theoretically and experimentally. The loss under normal operation is dominated by the speed-dependant spinning loss and the current dependant copper loss. The spinning loss is predicted to within 3% of the measured values at 1000 rpm, though significant uncertainty still exists in the laboratory torque measurements and the exact properties of the rotor materials used in the CDUSCM prototypes. The breakdown of measured spinning loss at 1000 RPM is shown in Figure 61.

The hysteresis-band current control technique imposes an additional loss in the machine that is speed dependent and significant where the modulation frequency exceeds 1 kHz. The main contribution to this additional loss comes from the eddy current loss in the rotor and copper. The addition of this loss mechanism reduces the machine efficiency, particularly at half-speed, though the resulting efficiency versus speed profile remains an increasing monotonic function as indicated in Figure 62.

![Pie Chart](#)

**Total Measured Spinning Loss = 67.7 W**

Figure 61: Predicted breakdown of measured CDUSCM spinning loss at 1000 RPM.
The peak predicted efficiency of the CDUSCM is 95.4% when operating at 1000 RPM and delivering a shaft torque of 25 Nm\(^3\).

Figure 62: CDUSCM Motor Efficiency including Armature Induced Losses.
5.1 Validity of 2D Assumptions
The 2D FEA presented in this and earlier chapters rely on a number of assumptions listed in section 4.2. The validity of the assumption that the 2D representation is sufficiently accurate is questionable. This is so for two reasons, firstly the copper windings are of a fixed width and lead to a stator slot whose width is not a function of radius. And secondly, the number of magnetic poles is small enough such that the flux travelling between magnets deviates significantly from the circumferential 2D surface. These causes for concern are investigated in this chapter.

The significance of flux traversing into and out of the 2D plane has particular effect eddy current losses in the stator laminations and also by creating a non-uniform flux density gradient between the inner and outer diameters of the machine. Any flux density gradient with respect to machine radius will increase the overall loss due to the non-linear nature of the loss mechanisms.

5.2 Radial Flux Gradients & Inter-laminar Flux
3D static FEA allows a basic investigation of the non-circumferential nature of the flux present in the CDUSCM under quasi-static conditions. Figure 63 shows the 3D wire-frame model of one pole of the CDUSCM that is used for the 3D static modelling of the CDUSCM in this section.

The Permanent Magnet (PM) field in the rotor is influenced by the fact that the flux produced by each magnet is a function of radius. That is, there is a greater proportion of flux towards the outer radius than at the inner radius. In contrast to this the rotor cross-section for the flux to traverse between magnets to constant with respect to radius. The fact that the path of least reluctance is not circumferential, but a straight line between successive poles reduces the tendency for higher flux densities at the outer radius. The superposition of these two effects is given by the flux density plot shown in Figure 64.
This is a plot upon the surface in the x-y plane midway between the external and air gap surfaces of the rotor.

Figure 63: 3D Wire-frame model of a single pole of the CDUSCM.

Figure 64: Static flux density in the central x-y plane of the rotor disc. Units are Tesla.

A flux density gradient from inner diameter to outer diameter is in fact not monotonic. Figure 65 shows the flux density in the rotor plotted along a radial line found in the plane used in Figure 64, midway between the magnets where the flux density is greatest. In the
rotor position used, the radial line is aligned with the centre of the middle slot. The flux density along this line peaks at a radius $\frac{3}{4}$ the distance from the inner radius to the outer radius and reduces most rapidly towards the inner radius. In part, the reduced flux density at $r = \frac{D}{2}$ is due to the rotor iron having a smaller inner radius to allow the rotor carrier to support the attractive axial force. The field in the rotor makes use of this extra volume and generally shows a reduced flux density magnitude with respect to that predicted by 2D FEA (Figure 13). The higher level of induction is present through the entire rotor thickness as visualised by the flux density magnitude in the Y-Z plane shown in Figure 66.

![Figure 65: Radial flux density gradient in the CDUSCM rotor midway between magnets.](image)

In contrast to the rotor, Figure 66 shows a much more uniform flux density under the winding slot in the same plane. This would suggest that there is radial flux component either side of the Y-Z plane. Also of interest is the flux density in the air gap as shown in Figure 67. The air gap flux shows reasonable uniformity versus radius with the exception near the magnet centreline. This provides a further suggestion that the radial flux component occurs in the magnetic materials.
Figure 66: Flux density magnitude, in Tesla, plotted in the Y-Z plane, coincident with the magnet gap and winding slot centrelines.

Figure 67: Flux density magnitude, in Tesla, plotted in the air gap 1mm from the stator surface.

To visualise the radial component of flux a number of vector plots were created along planes in the model. Figure 68 to Figure 74 display these plots that contain only the radial component of the flux vector and utilise different scales to fully illustrate the radial flux
flow. The vector plots utilise vector arrows whose size and colour are scaled according to the flux density magnitude.

Figure 68: Radial flux vector plot in the air gap, 1mm above the stator surface. Units are Tesla.

Figure 69: Radial flux vector plot in the rotor over the X-Y plane that lies centrally between the external surface and the air gap surface. Units are Tesla.
Figure 70: Radial flux vector plot over the radial plane coincident with the magnet centreline. Units are Tesla.

Figure 71: Radial flux vector plot over the radial plane coincident with the centreline of the stator tooth under the magnet edge. Units are Tesla.
Figure 72: Radial flux vector plot in the radial plane coincident with the stator winding slot and inter-magnet gap. Units are Tesla.

Figure 73: Radial flux vector plot over the X-Y plane set at $\frac{1}{2}d_{sd}$ below the air gap surface of the stator. Units are Tesla.
It is clear that there is very little radial flux in the air gap though a significant component exists in both the rotor and the stator. The peak radial component in the rotor is about 3 ½ times larger than that in the stator and results from the rotor iron having a small inside diameter than the magnet and stator annulus. The radial field in the rotor is less of a concern than that in the stator because of the potential of the radial field to increase eddy currents in the laminations. The majority of the radial flux in both the rotor and the stator lies towards the inner diameter. A peak in the radial component exists in the stator underneath the magnets and peaks at a Z coordinate near the base of each tooth or bottom of the winding slots. Another smaller magnitude radial component can be seen in the stator teeth very near to the air gap surface underneath the magnets. The peak in radial flux density is 0.138 T in the stator and 0.48 T in the rotor. The radial flux density figures given for the stator will be greater than the actual values as the permeability of the stator is actually anisotropic. The space between consecutive laminations, consisting of an insulating layer and air, reduces the relative radial permeability to approximately 98% of that in the circumferential plane. The 3D static model does not model this anisotropy.

The static model represents an approximation to very low speed operation. Having found the field behaviour at this point, the dynamic field behaviour needs to be considered. As each material in the machine exhibits frequency dependent field propagation, the transition from quasi-static to transient behaviour is dependant on the material and
Due to lamination, the stator construction performs efficiently in carrying time-varying magnetic flux in the plane of lamination (circumferential). The loss caused by radial or inter-laminar flux, however, is much higher as the geometry dimensions in this plane are large compared to the skin depth of the Lycore material. For example, the skin depth of Lycore 140 at 10 Hz is less than 5.8mm, whereas the stator back-iron thickness is 17mm. Under dynamic conditions, the presence of inter-laminar flux is reduced due to the increase in radial permeability of laminations at the lamination to inter-laminar-gap boundary. Essentially, the radial flux component found in the stator laminations shifts from the inter-laminar-gaps to the rotor-stator air gap. It is proposed that the 3D static scenario modelled earlier in this chapter presents a worst-case scenario.

The true 3D behaviour of the CDUSCM stator is difficult to model numerically because of the large volume and high number of laminations (approximately 140) coupled with finite computing power available. More recent and rather expensive releases of commercial software are able to model transient fields in three dimensions. This would suggest that the investigation of 3D fields in a finite length lamination stack due to permanent magnet movement is now possible and will reveal the exact nature of dynamic inter-laminar flux.

5.3 Conclusions
3D static FEA reveals that a significant radial component of flux exists in both the rotor iron and the stator. The cause of the radial flux component is a combination of:

- A constant thickness rotor back-iron
- Magnet induced flux that increases with radius
- Constant width winding slots leading to a tooth-width-per-pole-pitch that increases with radius
- A rotor back-iron inner diameter that is smaller than the permanent magnet inner diameter (for mounting purposes).

The rotor and stator experience peak radial flux density components of 0.48 T and 0.138 T respectively. This is a worst case scenario due to the isotropic modelling of the stator material and the probable effect that time-varying fields will have upon the flux.
distribution. The latest 3D transient analysis should give a good quantitative assessment of the dynamic inter-laminar flux.
Chapter 6

REDUCTION OF ELECTROMAGNETIC LOSSES

6.1 Introduction

There are only a couple of areas in the CDUSCM where practical measures may be taken to further reduce energy losses. In general, hysteresis losses can be reduced by the reduction of the intrinsic coercivity of a material, with a consequent reduction in the area contained within the hysteresis loop. And, eddy current losses can be reduced by decreasing the electrical conductivity of the material and by laminating the material, which has an influence on overall conductivity and is important because of skin effects at higher frequency.

6.2 Geometry Modifications

A number of proposals are presented in the remainder of this chapter that aim to improve the overall efficiency of the CDUSCM without major design alterations or additions to the mass of the machine. In the CDUSCM, there are some areas that exhibit higher loss under the different loss mechanisms detailed in chapter 4. The areas that are targeted for improvement are:

- The air gap surface of the rotor
- The permanent magnets

6.2.1 Rotor Lamination

As seen in sections 4.4.2.4 and 4.4.2.5, the eddy current loss in the rotor iron is significant, particularly under the high frequency excitation of the modulated armature currents. It is not easy to construct a rotor entirely from laminated materials due to high mechanical stresses imposed by the magnetic forces and general concerns about vibrations combined with the strong desire for minimal mass. Given that the high frequency fields only penetrate a few tenths of a millimetre at 5 kHz, it is probable that only the air gap surface of the rotor needs to be laminated. With this in mind, a 2D transient FEA model was constructed modelling a 0.35 mm Lycore-140 lamination on the surface of the rotor as
shown in Figure 75. This lamination thickness was chosen, as it is the same material as that used for the stator.

Two transient analyses were performed to evaluate the usefulness of the surface lamination by modelling the spinning loss and the armature induced loss in the same manner as described in sections 4.4.2.4.2 & 4.4.2.5. The transient spinning loss with the surface lamination compared to that of the “as-built” model is illustrated in Figure 76. The effectiveness of the lamination in reducing rotor eddy currents due to rotor rotation at 1000 rpm is small with an average loss reduction of 1.9 watts. The effect of the surface lamination is less significant when subject to fields produced by armature current modulation. This is shown in Figure 77 where an average rotor eddy current loss reduction of 1.3 watts is achieved between the non-laminated rotor and single-lamination rotor.

Figure 75: 2D transient FEA model of the CDUSCM incorporating 1 x 0.2 mm lamination on the air gap surface of the rotor.
Rotor Eddy Current Loss
(zero armature current, 1000 RPM)

Figure 76: 2D transient FEA rotor spinning loss results comparing a surface laminated rotor with the “as-built” CDUSCM.

Armature Modulation Induced Rotor Eddy Current Loss
(0 rpm, 20+- 5A at 7kHz)

Figure 77: 2D transient FEA rotor eddy current loss results comparing a surface laminated rotor with the “as-built” CDUSCM subject to armature current modulation induced fields.

6.2.2 Magnet Segmentation

Several flat pieces or segments of magnet material are sometimes used in radial flux machines to construct a “curved” pole, though more often it is done to reduce eddy currents in the magnet poles. The eddy currents can occur as a result of the asynchronous
stator MMF [73] or because of large reluctance variations due to open slots. Both of these conditions are present in the CDUSCM and hence magnet segmentation is a potential loss reduction technique.

At rated speed, the slot frequency is 600 Hz and the skin depth of NdFeB reduces to 25.4mm. Given that the maximum width and radial length of the magnets is about 50mm, it should be sufficient to break each magnet pole into 2, 3 or 4 segments and achieve a significant reduction. This is also based upon the findings of Howe [73] who reported a 90% reduction in eddy current loss by using 4 segments per pole in a high speed radial flux machine. 2D FEA models of the CDUSCM with 1, 2, 3 and 4 segments per pole were constructed to determine the most effective number of segments. The maximum number of segments is four due to brittle nature of the material and the manufacturing complexity for numbers beyond four.

Figure 78 shows the 3 finite element models used for the 2D transient simulation of eddy currents in the rotor of the CDUSCM with 1, 2, 3 and 4 segments per pole. The models include a 0.1 mm gap between segments through the total magnet arc width is the same for each model. The transient set-up assumes a steady state speed of 1000 rpm and a 2mm air gap distance. The calculated transient power losses of these models are compared in Figure 79.

The average loss versus the number of segments per pole, in Figure 80, shows a rather flat characteristic with a minimum at 3 segments per pole. The difference in rotor eddy current loss between using 1 segment per pole (the “as-built” scenario) and 3 segments per pole is 1.37 watts. This equates to 11% of the estimated as-built rotor eddy current loss of 12.6 watts.

The fact that the loss versus the number of magnet segments is not a monotonically decreasing function is in contrast to that found by Atallah et. al [74]. This deviation is largely because the transient analysis used for the CDUSCM includes PM magnetisation and hence accounts for the losses induced by stator slotting, whereas, Atallah doesn’t. It is concluded that the magnet segmentation provides a reduction in eddy current losses, but at the same time causes additional loss due to the slotting in the magnet pole (higher harmonics).
The potential spinning loss reduction suggests that 1-segment per pole would be the optimum, though the additional effort required during manufacture may result in the use of 1-segment per pole.

Figure 78: 2D transient FEA models of the CDUSCM using 1, 2, 3 and 4 magnet segments per pole.
Magnet Segmentation Effect Upon Rotor Eddy Current Loss
(zero armature current, 1000 RPM)

Figure 79: 2D transient FEA rotor spinning loss results comparing 2, 3 and 4 magnet segments per pole in the CDUSCM.

Average Rotor Eddy Current Loss vs Magnet Segments Per Pole
(zero armature current, 1000 RPM)

Figure 80: Average CDUSCM rotor spinning loss comparing 1, 2, 3 and 4 magnet segments per pole.

Contrary to a simple loss vs. frequency analysis, the effect of magnet segmentation is less pronounced under the higher frequency armature modulation scenario. Using the same FEA set up as that in section 4.4.2.5 and the 3 segments per pole geometry, the transient rotor eddy current loss can be compared to the single segment per pole scenario as shown in Figure 81. These results indicate an average loss reduction of 4.2 W.
6.3 Conclusions

Other than trivial improvements such as reducing stator lamination thickness and increasing the machine mass, there are few opportunities left to significantly and practically improve upon the CDUSCM. Two geometric modifications have been investigated with the aim of reducing rotor eddy current loss. These modifications being, rotor surface lamination and magnet segmentation have been shown, via 2D FEA, to reduce rotor eddy current losses. Armature modulation induced losses are based upon the machine operating at half its nominal speed due to the characteristic of the EMC employed and spinning losses are based upon the machine operating at its nominal speed of 1000 rpm. Table 18 presents a summary of the average spinning and armature modulation losses for the as-built, single rotor surface lamination and 3-segments per pole scenarios.

Table 18: Summary of rotor eddy current losses including the effect of two loss reduction techniques.

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Avg Spinning Loss</th>
<th>Avg Modulation Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Loss (W)</td>
<td>% Reduction</td>
</tr>
<tr>
<td>As-Built</td>
<td>12.61</td>
<td>-</td>
</tr>
<tr>
<td>Single Rotor Surface Lamination</td>
<td>10.71</td>
<td>15.1%</td>
</tr>
<tr>
<td>3-Segments per Pole</td>
<td>11.24</td>
<td>10.9%</td>
</tr>
</tbody>
</table>
Given the potential loss reductions shown in this chapter, it is recommended that future versions of the CDUSCM incorporate 3-segments per magnet pole. The use of a rotor surface lamination is marginally beneficial at 1000 rpm and would provide a significant benefit at higher speeds. The surface lamination method of loss reduction needs further optimisation.
Chapter 7

CONCLUSIONS

7.1 Conclusions
The application of the single-sided axial-flux brushless DC machine to a direct-drive solar powered racing car is appropriate due to its geometry ratio and its ability to achieve high efficiency, high volumetric power density and good reliability while maintaining an appreciable thermal mass required for traction applications. The characteristic of the CDUSCM meets these requirements and in doing so achieves an efficiency better than 80% throughout 87% of its operating region. The peak efficiency of the CDUSCM is 96.4% measured at 600 rpm and 10 Nm. The laboratory measurements of back-EMF versus speed and torque versus current has shown a 20% decrease in air gap flux since its manufacture in 1993.

The unique mechanical flux weakening property of the single-sided AFM has been experimentally investigated. The investigation shows that over the air gap range of 1mm to 4mm the machine constant changes by 36%. This corresponds to a moderate extension in the maximum speed relative to the speed extension capability of other techniques. 2D Transient FEA has shown that the cogging torque increases rapidly as the air gap reduces below 2mm, though remains relatively constant at air gap distances between 2mm and 4mm.

The electromagnetic loss induced by the presence of time varying armature currents has been investigated and it has been shown that this additional loss varies with speed in a manner similar to, but not exactly the same as, the switching frequency of the EMC. It is proposed that 21.5 watts is lost in addition to the spinning loss when operating at 500 rpm or half speed with an average armature current of 20 amps. Of this loss, a little over 57% is found in the rotor iron and about 23% is found in the copper. The armature induced loss in the magnets accounts for about 12.5% while the remaining 7.5% is located in the stator iron.
The assumption that the magnetic flux lies purely in a 2D circumferential plane has been tested through the use of 3D static FEA. The results from the analysis show that the stator is subject to a peak radial field variation of amplitude 0.138 Tesla. This variation will be reduced slightly by the anisotropic permeability of the stator laminations and further effected under dynamic conditions. The cause of this variation is the combination of non-radially aligned slot sides and a radially constant stator back-iron thickness. While inter-laminar flux in the stator has been shown to be relatively insignificant for 6 pole-pairs, the efficiency of lower pole count machines is expected to suffer from this effect. The rotor’s DC component of flux density is non-uniform in the radial direction and reaches a maximum at approximately 70% of the distance from the inner diameter to the outer diameter. This non-uniformity suggests that a non-uniform rotor thickness would be a more optimum design solution.

Two motor geometry modifications have been investigated with the aim of reducing rotor loss. The modifications have been investigated using 2D transient FEA and aim to reduce eddy current losses due to rotor rotation and that due to armature current modulation. It has been shown that the combined effect of the modifications detailed in Chapter 6 coupled with a rotor speed of 1000 rpm, ought to give rise to an average rotor spinning loss reduction of 3.2 watts. The same rotor modifications and a rotor speed of 500 rpm are expected to lead to a rotor eddy current loss reduction of 5.5 watts when a 7 kHz, symmetrical, triangular current waveform of 20±5A excites the armature. In either case the loss reduction is approximated by a 25% reduction in rotor eddy current losses and suggests a potential machine efficiency improvement of in the order of 0.3%.

7.2 Recommendations for Further Work

With an improved understanding of the magnetic circuit and the extent to which 2D analysis assumptions hold, further efforts are possible to both evaluate and improve upon the single-sided axial flux machine for traction applications.

Of utmost importance is experimental data acquired with reliable accuracy upon a machine with known material characteristics. To date, this has not been done with complete satisfaction and is a primary goal to achieve in the near future. This would involve the construction of a new machine from materials with measured characteristics. This would provide a much-needed source for accurate experimental validation.
With a known base case, several other issues that warrant further investigation could be pursued. These issues fall broadly into the categories of experiment work and finite element analysis.

7.2.1 Experimental Work
A number of experimental tests are recommended to enhance the work presented in this thesis.

The active torque measurements in chapter 3 ought to be repeated with a more reliable torque transducer, preferably one that relies on a non-contact method of measurement between the rotating shaft and the transducer body.

The measurement of static torque provides an important tool for validation and also for gauging hysteresis loss. An apparatus could be constructed whereby static torque can be measured with an absolute minimum of backlash. By measuring the static torque versus position in both directions, a measurement can be made of the hysteresis exhibited in the machine. A high-resolution torque transducer would be required, as the cogging torque would tend to dominate those measurements.

The current manufacturing process of the CDUSCM does not include any post-process annealing to relieve stresses in the steel. It has been reported that stresses from slitting and machining may cause an additional 20% loss in magnetic steels [75].

7.2.2 Finite Element Analysis
One outcome this thesis is that the sensitivity of the machine constant to the air gap distance is quite low. Of interest to designs that need a significant level of flux weakening are the factors that affect the sensitivity of the machine constant to air gap distance. The relative importance of those factors needs to be established so that a specified sensitivity can be achieved where necessary.

Typical operation of a hysteresis band current controller imposes asymmetrical current waveforms upon the armature. The effect of asymmetrical currents upon eddy current losses in the machine is an extension that could be made to the work found in section 4.4.2.5.

Because the windings for the prototype CDUSCMs are first formed and then inserted into the stator slots, the ability to include teeth on the stator is limited. For machines with
higher turn counts and smaller winding cross-sections, the possible presence of stator teeth is likely to reduce the flux variation due to stator slotting. The effect of stator teeth upon rotor and stator losses needs to be quantified for the single-sided AFM, though the results are likely to be similar to those found for radial flux machines.

The radial flux variation observed in the rotor back-iron suggests that a non-uniform rotor thickness would be desirable to achieve the optimum efficiency versus weight performance. A further analysis needs to be completed to obtain the optimum back-iron profile. The effect of difference numbers of pole pairs also needs to be established particularly for low pole count machines.

The most recent FEA software has capabilities in 3D transient analysis and hysteresis modelling. The ability to model laminated structures in 3D under arbitrary excitation will be particularly useful in understanding the dynamic nature of the inter-laminar flux caused by the geometry of the AFM. In addition, it is recommended that further work be completed to quantify the magnitude of the radial flux field with a variable air gap distance. Significant computing resources would be required to process such a detailed problem.

Imperfections in the insulation between the laminations will reduce their effectiveness in reducing eddy currents. This is typified by the mounting arrangement whereby 9 x M5 bolts are used to through-bolt the stator carrier. The presence of these bolts, in direct contact with up to 17 laminations, affects more than 10% of the stator. This potential short circuit between consecutive laminations provides a path for eddy currents between laminations and is likely to reduce their effectiveness. To optimise the structure used to attach the stator laminations to its carrier, the increase in loss due to such attachments needs to be quantified.

One particular disadvantage of the axial flux machine is its tendency to behave like an acoustic diaphragm. The moderate changes in axial force caused by both rotation and armature current commutation/modulation can be amplified by the mechanical structure of the rotor disk. An optimisation of the mechanical structure needs to be done to mitigate this problem, which is of great importance to many consumers.
## Appendix A

**NUMERICAL PROPERTIES OF THE CHARLES DARWIN UNIVERSITY SOLAR CAR MOTOR**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator Outside Diameter</td>
<td>$D_o$</td>
<td>260</td>
<td>mm</td>
</tr>
<tr>
<td>Stator Inside Diameter</td>
<td>$D_i$</td>
<td>160</td>
<td>mm</td>
</tr>
<tr>
<td>Number of Poles</td>
<td>$p$</td>
<td>12</td>
<td>-</td>
</tr>
<tr>
<td>Air Gap Flux Density</td>
<td>$B$</td>
<td>0.72</td>
<td>Tesla</td>
</tr>
<tr>
<td>Air Gap Length</td>
<td>$d_{ag}$</td>
<td>2</td>
<td>mm</td>
</tr>
<tr>
<td>Magnet Mechanical Angle</td>
<td>$\theta_{mm}$</td>
<td>21.37</td>
<td>degrees</td>
</tr>
<tr>
<td>Magnet Thickness</td>
<td>$d_{mt}$</td>
<td>4</td>
<td>mm</td>
</tr>
<tr>
<td>Slot Depth</td>
<td>$d_{sd}$</td>
<td>18</td>
<td>mm</td>
</tr>
<tr>
<td>Slot Width Factor</td>
<td>$k_{sw}$</td>
<td>0.43</td>
<td>-</td>
</tr>
<tr>
<td>Stator Thickness</td>
<td>$d_{st}$</td>
<td>35</td>
<td>mm</td>
</tr>
<tr>
<td>Rotor Thickness</td>
<td>$d_{ri}$</td>
<td>12</td>
<td>mm</td>
</tr>
<tr>
<td>Number of Turns per Phase</td>
<td>$t$</td>
<td>6</td>
<td>-</td>
</tr>
<tr>
<td>Copper Cross-section</td>
<td>$A_{cu}$</td>
<td>3 x 6</td>
<td>mm</td>
</tr>
<tr>
<td>Copper Utilisation Factor</td>
<td>$k_{cu faculte}$</td>
<td>76.6</td>
<td>percent</td>
</tr>
<tr>
<td>Number of Phases</td>
<td>$n$</td>
<td>3</td>
<td>-</td>
</tr>
<tr>
<td>Envelope Volume</td>
<td></td>
<td>0.00641</td>
<td>m$^3$</td>
</tr>
<tr>
<td>Nominal Voltage</td>
<td></td>
<td>55</td>
<td>volts</td>
</tr>
<tr>
<td>Nominal Current</td>
<td></td>
<td>100</td>
<td>amps</td>
</tr>
<tr>
<td>Nominal Torque</td>
<td></td>
<td>55</td>
<td>Nm</td>
</tr>
<tr>
<td>Nominal Speed</td>
<td></td>
<td>948</td>
<td>rpm</td>
</tr>
<tr>
<td>Maximum Efficiency</td>
<td></td>
<td>96.4</td>
<td>percent</td>
</tr>
<tr>
<td>Active Mass</td>
<td></td>
<td>11.74</td>
<td>kg</td>
</tr>
<tr>
<td>Mass Power Density</td>
<td></td>
<td>468.5</td>
<td>W/kg</td>
</tr>
<tr>
<td>Volumetric Power Density</td>
<td></td>
<td>858.0</td>
<td>kW/m$^3$</td>
</tr>
</tbody>
</table>
Appendix B

CALCULATION OF THE LOCAL GRAVITATIONAL ACCELERATION CONSTANT

The localised gravitational acceleration constant can be calculated using equation (19) quoted from [76].

\[
g = 9.7803184(1 + A \sin^2 L - B \sin^2 2L) - 3.086 \times 10^{-6} H \quad (19)
\]

where \( A = 0.0053024 \)
\( B = 0.0000059 \)

\( L \) is the latitude (12° 28' South for Darwin),
\( H \) is the height in metres above mean sea level (30m for Darwin).

For the location of the experiments at the CDU in Darwin the value of 9.7826 m.s\(^{-2}\) has been used.
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