Spring 1995

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Analysis and design considerations of a high-power density, dual air gap, axial-field, brushless, permanent magnet motor

Abstract
Until recently, brush dc motors have been the dominant drive system because they provide easily controlled motor speed over a wide range, rapid acceleration and deceleration, convenient control of position, and lower product cost. Despite these capabilities, the brush dc motor configuration does not satisfy the design requirements for the U.S. Navy’s underwater propulsion applications.

Technical advances in rare-earth permanent magnet materials, in high-power semiconductor transistor technology, and in various rotor position-sensing devices have made using brushless permanent magnet motors a viable alternative. This research investigates brushless permanent magnet motor technology, studying the merits of dual-air gap, axial-field, brushless, permanent magnet motor configuration in terms of power density, efficiency, and noise/vibration levels.

Because the design objectives for underwater motor applications include high-power density, high-performance, and low-noise/vibration, the traditional, simplified equivalent circuit analysis methods to assist in meeting these goals were inadequate. This study presents the development and verification of detailed finite element analysis (FEA) models and lumped parameter circuit models that can calculate back electromotive force waveforms, inductance, cogging torque, energized torque, and eddy current power losses. It is the first thorough quantification of dual air-gap, axial-field, brushless, permanent magnet motor parameters and performance characteristics.

The new methodology introduced in this research not only facilitates the design process of an axial field, brushless, permanent magnet motor but reinforces the idea that the high-power density, high-efficiency, and low-noise/vibration motor is attainable.

Keywords
Physics, Electricity and Magnetism, Engineering, Electronics and Electrical, Engineering, Mechanical, Engineering, Marine and Ocean

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ANALYSIS AND DESIGN CONSIDERATIONS
OF
A HIGH-POWER DENSITY, DUAL AIR GAP, AXIAL-FIELD,
BRUSHLESS, PERMANENT MAGNET MOTOR

BY

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B.S.E.E., University of Massachusetts/Lowell, 1987
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A DISSERTATION

Submitted to the University of New Hampshire
in Partial Fulfillment of
the Requirements for the Degree of

Doctor of Philosophy

in

Engineering

May 1995
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7 March 1995
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DEDICATION

To my father.
ACKNOWLEDGMENTS

The author wishes to express his appreciation to his advisor and dissertation director of his graduate committee, Dr. B. K. Fussell, for his patience, encouragement and expert technical guidance given throughout the development of this research.

Thanks are expressed to the other members of the author's graduate committee, Dr. D. E. Limbert, Dr. R. B. Jerard, Dr. L. G. Kraft, and Dr. K. Sivaprasad, for their encouragement and support. The interaction with all members of the committee was indeed a valuable learning experience.

Special thanks go to Dr. Charles K. Taft, Professor Emeritus, University of New Hampshire, for his friendship and guidance during the period of research.

Special acknowledgment and gratitude are accorded to Dr. D. C. Hanselman, University of Maine, Orono, ME, Mr. J. J. Stupak, Jr., Portland, OR, and Dr. J. Y. Hung, Auburn University, AL, for their exceptional insight, assistance and personal concern for the author's efforts throughout the work.

Gratitude is due Mr. K. Plant, Mr. W. Henrickson and Dr. T. Keim at KAMAN Co., Electromagnetic Division, Hudson, MA, for their generous permission for use of their technical information and test data.

Thanks are expressed to all co-workers, Mr. M. Keshura, Dr. P. Hendricks, Mr. A. Barnett, Mr. J. Raposa, Mr. R. A. Bedingfield, Anne Kuklinski, and Dr. R. Kuklinski of the Naval Undersea Warfare Center, Newport Division, Newport, RI.
Thanks are also go to Mr. Kevin Peters and Mr. Shine Ho of the Dynamic Systems Modeling Lab. at University of New Hampshire. In particular, deeply indebted thank to Mr. Timothy S. Burke of the Eastern Air Devices, Dover, NH, Dr. Steven R. Prina, Dr. Timothy J. Harned, Dr. Steve Huard of the Parker Hannifin Corporation Motor Design Center, Portsmouth, NH for many suggestions, comments and ideas during the course of this study.

The author wants to special thank his wife (Kim Inok), his first son (David), his daughter (Grace), his second son (Joseph), mother (Nayun), and his brothers and sisters for their encouragement and understanding. Their enthusiasm for his work is something for which he will always be grateful.
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LIST OF SYMBOLS

A  Magnetic vector potential
B  Flux density vector
B_r  Residual (remanence) flux density
H  Magnetic field intensity vector
H_c  Coercive magnetic force
H_{dB}  Coercivity of normal B/H curve
H_M  Permanent magnet coercive force
W_M  Magnetic energy
W_{co}  Magnetic co-energy
F  Force
q  Charge
J  Current density
v  Velocity
T_{mn}  Maxwell stress tensor
\phi  Flux
\Phi  Total flux
\mu_r  Relative permeability
\mu_o  Permeability of free space (air)
S  Surface
V  Volume
\lambda  Flux linkage
L  Inductance
J_{PM}  Permanent magnet equivalent current density

xvi
\( I_{PM} \) Permanent magnet length
\( E \) Induced voltage
\( \theta \) Angular displacement
\( \omega \) Angular velocity
\( X \) Reactance
\( M_i \) Magneto motive force (MMF) of winding
\( F_i \) Magneto motive force (MMF) of permanent magnet
\( R \) Reluctance
\( K_T \) Motor torque constant
\( K_E \) Motor back emf constant
\( \sigma \) Electric conductivity
\( \delta \) Skin depth (penetration depth)
\( E_i \) Phase back emf
\( B \) Number of pole pairs
\( R_L \) Slot leakage reluctance
\( s_w \) Slot width
\( T \) Thickness of tooth
\( l_s \) Slot depth
\( \Phi_R \) Air-gap flux
\( R_R \) Air-gap reluctance
\( N \) Number of coil turns
\( \Phi_L \) Slot leakage flux
\( i \) Applied phase current
\( T_i \) \( i^{th}\)-segment cogging torque
\( W_{coi} \) \( i^{th}\)-segment total magnetic co-energy
\( x \) Rotor displacement
\( L_{ri} \) \( i^{th}\)-segment radial thickness or model depth
$N_p$  Total number of poles

$l_{armi}$  $i^{th}$-segment lever arm length
ABSTRACT

ANALYSIS AND DESIGN CONSIDERATIONS OF
A HIGH-POWER DENSITY, DUAL AIR GAP, AXIAL-FIELD,
BRUSHLESS, PERMANENT MAGNET MOTOR

by

Chahee Peter Cho

University of New Hampshire, May, 1995

Until recently, brush dc motors have been the dominant drive system because they provide easily controlled motor speed over a wide range, rapid acceleration and deceleration, convenient control of position, and lower product cost. Despite these capabilities, the brush dc motor configuration does not satisfy the design requirements for the U.S. Navy's underwater propulsion applications.

Technical advances in rare-earth permanent magnet materials, in high-power semiconductor transistor technology, and in various rotor position-sensing devices have made using brushless permanent magnet motors a viable alternative. This research investigates brushless permanent magnet motor technology, studying the merits of dual-air gap, axial-field, brushless, permanent magnet motor configuration in terms of power density, efficiency, and noise/vibration levels.

Because the design objectives for underwater motor applications include high-power density, high-performance, and low-noise/vibration, the traditional, simplified equivalent circuit analysis methods to assist in meeting these goals
were inadequate. This study presents the development and verification of detailed finite element analysis (FEA) models and lumped parameter circuit models that can calculate back electromotive force waveforms, inductance, cogging torque, energized torque, and eddy current power losses. It is the first thorough quantification of dual air-gap, axial-field, brushless, permanent magnet motor parameters and performance characteristics.

The new methodology introduced in this research not only facilitates the design process of an axial field, brushless, permanent magnet motor but reinforces the idea that the high-power density, high-efficiency, and low-noise/vibration motor is attainable.
CHAPTER 1

INTRODUCTION

1.1 ELECTRIC MOTOR

An electric motor is a well-known device that converts electrical energy to mechanical energy using magnetic field linkage. An electric motor consists of two major elements: (1) a fixed stator with current-carrying windings or permanent magnets and (2) a rotating rotor, which provides a magnetic field produced by additional current-carrying windings or attached permanent magnets. Basic motor action is caused by an interaction between the rotor and stator magnetic fields [1,2,3]. A brief history of electric motors is presented in this section.

Because the history of the electric motor/generator has already been adequately chronicled [4], only a brief account is given in this chapter. Many of the earliest electric machines were axial flux machines, but the radial-flux machine, in terms of the historical life of electrical motors, achieved and maintained prominence in a relatively short period of time. The first recorded electric motor is Michael Faraday's axially oriented field, disc-type, eddy current induction motor introduced in 1831, one year after Ampere discovered the nature of electric current and its relationship to magnetism. The first patent for an electric motor was for a radial field motor obtained by an American inventor, Thomas Davenport, in 1837 [4]
Since 1837, many patents for electric motors -- mostly radial field machines -- have been issued. Because there have been more applications for radial field machines, they have developed more rapidly than the axial field machines. Axial machines have been studied for only a few applications, such as the printed circuit motor and the superconducting homopolar machine [5]. The major disadvantage of the axial field motor is the strong magnetic attractive force between the rotor and the stator. This disadvantage, however, can be overcome by using either a double stator with single-rotor configuration or by using an ironless rotor as is in the printed circuit motor.

Modern electric motor advances are characterized by frequent developments and refinements in magnetic materials, integrated circuits, power electronic switching devices, and manufacturing technology, rather than by fundamental changes in operating and control principle. The dramatic improvements in permanent magnet materials and power electronic devices over the last two decades have led to the development of brushless permanent magnet motors that offer significant improvements in power density, efficiency, and noise/vibration reduction. Also, less radiated noise, i.e., no electrical sparks.

1.1.1 Brushless Permanent Magnet Motor Technology

Although brushless, permanent magnet motors are highly efficient, brush direct current (dc) motors with a range of output powers are the dominant drive system used in various applications. Brush dc motors are widely used because they provide easily controlled motor speed over a wide range, rapid acceleration and deceleration, convenient control of position, and lower production cost.
Electronically commutated, brushless permanent magnet motors are, however, becoming prime movers in vehicle propulsion, industrial drives, and actuators as a result of improvements in rare-earth magnet materials, advances in the power electronic devices, and power integrated circuits in the last two decades. Not only have there been gradual improvements in ceramic and other alloys, but the rapid development of rare-earth magnets, such as samarium-cobalt (Sm Co) and neodymium-boron-iron (Nd B Fe) around 1980, have provided designers with a significant increase in available field strength. These new advanced technologies have made possible the development of a high-power density, brushless axial field, permanent magnet motor system that provides a very high torque to inertia ratio.

For example, the power range of a few hundred to a few thousand horsepower has been achieved [6], with obvious advantages over brush dc motor systems. Additionally, many studies investigating the performance characteristics of both ferrite and Sm Co brushless motors have demonstrated the superiority of the rare-earth permanent magnet material in motor design and its desirability for present and future applications.

Figure 1.1 and Figure 1.2 show the basic components of the brushless, permanent magnet motor system. The electrical energy can be a dc source, such as a battery, or an alternating current (ac) source. The power inverter/converter section takes the available dc or ac energy from the electrical energy source and inverts/converts that energy into the shape required by the particular motor configuration. Using this inverted/converted electric energy, the brushless, permanent magnet motor produces mechanical output power, which is sent to the mechanical load. Control of the brushless, permanent magnet motor drive system is provided to maintain stable operation and optimize performance.
Figure 1.1 Block Diagram of a Brushless, Permanent Magnet Motor

Figure 1.2 Graphical Representation of a Brushless, Permanent Magnet Motor
The immediate advantage of eliminating brush-type commutation within a permanent magnet motor is that such a design increases motor bearing life because there is no brush debris or dust that contributes to bearing wear. Additionally, the brushless design offers increased motor speed range because the motor speed is not limited by the arcing of the commutator to brush as it is with brushed motors. The brushless, permanent magnet motor allows motor operation in applications in which low noise/vibration is required — a primary goal for U.S. Navy underwater propulsion applications.

In addition to the initial manufacturing cost of the rare-earth magnetic materials for rotor construction, a handicap for both brushed and brushless motors, there are several other obstacles to overcome in pursuing permanent magnet motor technology. A rotor-shaft, position-sensing device with a more sophisticated electronic inverter and controller circuit is also required. Also, motor commutation must be accomplished electronically, resulting in increased complexity of the design of the motor controller and a commensurate increase in the cost of its implementation. The solutions to these disadvantages are sensorless, brushless, permanent magnet motor designs and inexpensive integrated circuits (IC) for switching. Many researchers and designers are working on the sensorless motor design and compact digital control areas.

1.1.2 Brushless, Permanent Magnet Motor Configurations

There are many motor and magnetic field configurations that can be used for brushless, permanent magnet motor technology. However, to date, it has not been determined which configuration should be used to maximize the power density, efficiency, and quietness of a motor. To thoroughly investigate brushless, permanent magnet motor technology, it is necessary to study the
relative merits of each configuration in terms of the power density, efficiency, and noise/vibration levels.

A. Description of Conventional Radial Field Motor

Figure 1.3 and Figure 1.4 show a typical radial field, brushless, permanent magnet motor configuration. In this motor configuration, primary magnetic flux from the rotor permanent magnet crosses the air gap in a radial direction to interact with the stator phase windings in axial direction orthogonal to each other to maximize Lorentz force production, \( dF = I \, dI \times B \). There are many ways to achieve this force depending upon how the permanent magnets are placed on the rotor. However, this section describes only the configuration in which the flux from the permanent magnet crosses directly into the air gap. The permanent magnet retaining technique, one of many critical design issues, can be accomplished by a bonding or a nonmagnetic retainment ring.

Alnico and ferrite (ceramic) permanent magnet types were the most popular and common materials used in radial field motor configurations until the rare-earth permanent magnet materials were developed. By using these newly developed rare-earth permanent magnets, the weight and volume can be reduced, thus increasing the motor output power density and performance. The possibility of using high-energy density, rare-earth magnets to increase the power density of the motor necessitates quantifying the performance of both Sm Co- and Nd B Fe-type magnets in this configuration. Numerous studies about this motor configuration in conjunction with high-energy density rare-earth permanent magnets have been done [2,3,6].
Figure 1.3 Typical Radial Field, Permanent Magnet Motor

Figure 1.4 Detailed Illustration of Radial Field, Permanent Magnet Motor
B. Description of Axial Field Motor

There are three types of axial field motor configurations: single stator with single rotor (Figure 1.5), double stator with single rotor (Figure 1.6), and single stator with double rotor (Figure 1.8). In most axial field, permanent magnet motor configurations, the primary magnetic flux from the rotor permanent magnet crosses the air gap in the axial direction. The rotor is a disc structure containing axially magnetized permanent magnet blocks. An advantage of the axial field configuration is that the retaining ring, which holds the permanent magnets in the disc structure, is not part of the main magnetic circuit. Consequently, the retaining ring can be made to the required thickness to withstand the rotational stresses without affecting the motor magnetic performance (see Figure 1.7). The stator is a disc structure consisting of windings, slots, and teeth. The stator for an axial field motor configuration is a radial array of current-carrying conductors (motor windings) with appropriate end turns fixed in the stator slots.

There are three methods to achieve the described axial field laminated stator configurations.

1. The simplest way to construct a laminated stator is by winding thin stator laminated magnetic material into a roll and laser-cutting slots through the tape [7]. Though this method is the easiest manufacturing method, it is not the preferred method because magnetic properties are lost during the slotting process.

2. A more intricate method to produce a laminated stator is by slotting the magnetic material tape first and then winding the magnetic tape so that the precut slots are aligned [8]. This method is more difficult because of the
criticality of the slot-to-slot distance. The advantage of this procedure is that the magnetic property lost in the cutting process can be retrieved before winding.

Figure 1.5 Single-Stator, Single-Rotor, Axial Field Motor Configuration

Figure 1.6 Double-Stator, Single-Rotor, Axial Field Motor Configuration
3. A third method to construct a stator, developed by Weh [9], is molding a stator from iron (magnetic material) powder and resin. This type of stator, however, does not have high density; the low relative permeability (μr) of the stator requires more rotor permanent magnet volume in order to produce the
same output torque. The main advantage of this type of stator is the virtual elimination of eddy current losses due to high-frequency pulse-width modulation (PWM) components in the stator teeth and back iron.

Figure 1.8 shows a stator structure with the teeth made of soft iron stampings held between two discs as proposed by Leung [10] for a single-stator with a double rotor, axial field generator. As presented by Capaldi [8], the best power density (motor output power/volume) from an axial field-type motor is achieved with a dual air-gap motor, either a double-stator/single-rotor type or single-stator/double-rotor type.

![Diagram of Double-Rotor with Single-Stator Configuration](image)

**Figure 1.8 Double-Rotor with Single-Stator Configuration**
1.2 LITERATURE SURVEY OF THE AXIAL FIELD MOTOR

The design and analysis of axial field motors has been a subject of research for the last two decades [8,10,11,12,13,14,15,16]. The majority of work has been done on a single-stator with single-rotor, eddy current type or induction motor configuration. A series of papers [8,12,14,15,16] deal with different axial field configurations and their motor performance analysis using simplified analytical formulas based on Maxwell's field equations.

Most of these research studies were conducted to develop new axial field motor concepts and to make comparisons with similar axial field-type or conventional radial-type motors. Therefore, simplified analytical equations were sufficient to support those studies. The objective of this author's research, however, is to develop high-power density, high-efficiency, and low-noise/vibration motors for underwater applications. Unlike other researchers, this author used finite element analysis (FEA) combined with lumped magnetic circuit models to analyze and predict axial field motor parameters and performance characteristics.

A double-stator, single-rotor, variable speed eddy current motor, which is similar to the axial field induction motor, has been built and tested by Nasar [12]. The analysis was performed using an approximate formula based on Maxwell's field equations with many simplifying assumptions. Analytical results of the speed-torque characteristics were then compared with experimental results, and the correlation was poor. Nasar's analysis is applicable to eddy current motors only.

Capaldi [8] studied the double-stator with single-rotor, axial field induction motor and reluctance motor configuration. The potential output
power/weight was compared with that of a conventional radial field induction motor. The air-gap flux was calculated using the conformal transformation method. The purpose of Capaldi's study was to compare axial field induction and reluctance motors with conventional radial field induction motors. Results showed that the power/weight ratio of axial field induction and reluctance motors is twice that of the radial field induction motor [8].

Campbell studied a permanent magnet, brush, dc single-stator motor with a single-rotor configuration, similar to the printed circuit motor [11]. The approximated field in the air gap is calculated using a simple potential equation with the assumptions of an infinitely long magnet pole and a uniform potential over the whole permanent magnet surface. The objective of Campbell's work was to describe a simple, low-cost, and compact air-gap motor; detailed analyses and parameter calculations were not provided.

Boules and Weh first proposed the use of rare-earth Sm Co permanent magnets in the axial field configuration [14]. The axial field synchronous machine proposed by Weh [9] has a rotor consisting of two discs with permanent magnets attached to the surface. A stator is placed between the discs (see Figure 1.8). The permanent magnets facing each other on the two discs are of opposite polarity, and the magnetic flux travels axially from one disc to the other through the stator. Motor constants and design considerations are discussed using an equivalent circuit model, which was developed in terms of the principal machine dimensions. The approximated calculations were based on a field analysis using a linearized model. In this equivalent circuit, the magnet leakage and stator material permeability are not included. An optimum design would only be possible with a more detailed equivalent circuit and with the aid of a computer program.
Amaratunga, Acarnely, and McLaran studied an axial field, permanent magnet generator as a candidate for an aerospace application [15]. An air-gap flux density calculation, using a simple two-dimensional (2-D) FEA was performed. The main objective was to compare output power with a given volume and weight constraints for several different types of generator configurations, such as radial field, axial field, flux-squeezing, and homopolar. A detailed FEA analysis was not carried out. Amaratunga et al. concluded the radial field permanent magnet motor offered the highest power density.

The axial field motor literature reviewed in this study pertains to low-cost and simple manufacturing production commercial motors for which simplified analysis methods were sufficient. The design goals for underwater motor applications, however, are high-power density, high-performance, and low-noise/vibration. The complexity and rigor of these objectives necessitated more advanced and exacting analysis tools, techniques, and design methods.

1.3 SCOPE OF THIS STUDY

The scope of this study is limited to the development and verification of detailed FEA models and lumped parameter magnetic circuit models of a dual air-gap, axial field, brushless, permanent magnet motor with the purpose of making design improvements. Motor design parameters, cogging torque and axial force variation, and eddy current power losses in the rotor permanent magnet are investigated. Methods using FEA with a generic modeling program package developed by this author and lumped parameter magnetic circuit analysis are directly applied to an axial field, dual air-gap, brushless, permanent magnet motor to improve its power density, performance, and noise emission.
The author analyzed a preliminary motor design by Kaman Electromagnetics Corporation (KEC) (formerly EML Research) of Hudson, MA. Both the preliminary and final prototype motors in this study were designed and developed as a joint effort between the Department of Navy, Naval Undersea Warfare Center (NUWC) Division Newport, Newport, RI and KEC.

1.4 ORIGINAL CONTRIBUTIONS IN THIS DISSERTATION

The research work presented in this dissertation is the first thorough investigation quantifying dual air-gap, axial field, permanent magnet motor parameters and performance characteristics using FEA combined with lumped parameter magnetic circuit analysis.

The main contribution of this dissertation is the development and verification of detailed FEA models combined with lumped parameter magnetic models that can calculate back electromotive force (emf) waveforms, inductance, cogging (detent) torque, and energized torque waveforms. The application of these models allows accurate prediction of the motor parameters and the dynamic behavior of such motors without the construction of costly prototypes. The motor model-generated back emf waveform, inductance value, cogging torque, and torque waveforms closely match those obtained in experimental tests performed on prototype motors. This result supports the various assumptions and simplifications used in the derivation of the models. Furthermore, because a generic FEA modeling program package and standard lumped parameter magnetic modeling techniques are used, the same modeling approach is applicable to other axial field, brushless motor sizes and configurations.

In summary, original contributions include:
• Application of the FEA and lumped parameter magnetic circuit analysis to an axial field, two-stator/one-rotor configuration, dual air-gap, high-power density, large horsepower, brushless permanent magnet motor for high-performance and quiet operation.

• Performance of a convergence test of FEA models using a refinement method in conjunction with total magnetic co-energy calculation.

• Calculation of motor parameters and performance characteristics using an FEA and lumped magnetic model. These calculations include flux distribution in the motor, back emf waveforms, cogging and energized torque waveforms, inductance, eddy current power loss calculations.

• Use of a simplified multisegment, 2-D FEA models with a developed computer algorithm to predict the complicated process of calculating cogging torque without using three-dimensional (3-D) FEA models.

• Development of an algorithm to allow incremental movement of a rotor (permanent magnet) geometry to calculate cogging torque and energized torque.

• Quantification of a trade-off study between cogging torque and axial force using the FEA method.

• Performance of a rotor misalignment effect study in a dual air-gap configuration using the FEA method.

• Development of an algorithm to calculate the effect of permanent magnet skewing on cogging torque using SIMULAB (an analysis and data manipulation software tool based on the MATLAB program).
• Derivation of an analytical expression and FEA model to predict eddy current power losses in the permanent magnet material to compare laminated and un laminated material. (This expression and FEA model can also be used to predict stator core eddy current power losses.)

• Development of an axial motor design methodology using integration of FEA and lumped magnetic circuit analysis.

1.5 OVERVIEW OF THE CONTENTS

Chapter 1 contains the introduction and history of brushless, permanent magnet motors and axial field motor configurations; a literature survey of the axial field, brushless, permanent magnet motor, the scope of the study, and the original contributions of this dissertation.

Chapter 2 describes the prototype axial field, brushless, permanent magnet motor configuration and includes design considerations, motor descriptions, prototype motor fabrication procedures, and drive techniques.

Chapter 3 is a detailed investigation of the axial field motor parameters and characteristics calculations using FEA and lumped magnetic circuit models. Different force/torque calculation methods are examined and compared. To facilitate this investigation, the author developed an automatic, generic FEA modeling program package. FEA models are analyzed assuming a quasi-static magnetic field variation. Checking FEA model accuracy using magnetic co-energy calculation is presented. A trade-off between the model accuracy (total number of elements) and computer time is also discussed. Chapter 3 also describes a multisegment modeling technique. Motor parameters that affect the motor output torque and performance, such as back emf waveforms and
inductance values, are identified. The magnetic force/torque in an axial field, dual air-gap motor is calculated from the flux density distribution by FEA using the virtual work method (co-energy method). Finally, the energized torque waveform is calculated under an electrically loaded condition. The results are compared with experimentally measured values from the prototype motor for verification of the models.

Chapter 4 calculates cogging torque considering a wide range of conditions using FEA models by moving a permanent magnet in the rotor along the air gap. The cogging torque is investigated under several different conditions to include staggered and unstaggered stator and skewed vs. unskewed permanent magnet. Chapter 4 also calculates and presents an axial-force variation and the effect of rotor misalignment. Based on this thorough investigation, a critical design trade-off between the cogging torque reduction and the axial force variation is presented. The effect of staggering the stators on the output torque was examined using measurement of back emf waveforms as a function of different speeds, thus, allowing calculation of back emf constant and torque constant for staggered stator and unstaggered stator.

Chapter 5 presents the detailed analysis and results of an axial field, brushless, permanent magnet motor eddy current power loss calculations in the permanent magnet material using an FEA model and a lumped magnetic model expression. The rotor permanent magnet experienced excessive temperature rise and was laminated to reduce eddy current power loss. Results of this study not only compare the eddy current power loss in the laminated vs. unlaminated permanent magnet, but also demonstrate how to prevent possible demagnetization of the permanent magnet.
Chapter 6 discusses a new three-step iterative design methodology for an axial field, brushless, permanent magnet motor based on the results of this study and the prototype motor tests. This chapter describes a slotless, axial field, brushless, permanent magnet motor an alternative design to the prototype slotted stator, axial field motor, brushless, permanent magnet motor.

Chapter 7 provides conclusions and recommendations for further study.
CHAPTER 2

AXIAL FIELD PROTOTYPE MOTOR DESIGN CONSIDERATIONS AND DESCRIPTION

2.1 INTRODUCTION

This chapter presents design considerations and a prototype motor configuration for the brushless, axial field, permanent magnet motor whose attributes include high-power density, high-performance (high-efficiency), and low-noise/vibration. Design constraints, design specifications, and critical motor components are also addressed.

The prototype motor was designed to provide a high-power density, high efficiency, and low-noise/vibration propulsion electric motor for underwater applications. With the U.S. Navy’s recent focus on brushless, permanent magnet electric motor systems for long-range underwater propulsion, the design of high-power density, quiet, and reliable motor systems has become critical in meeting current and future underwater propulsion requirements.

The author analyzed the preliminary KEC design. Research results included an extensive and accurate cogging torque analysis, a trade-off study between cogging torque reduction and axial force variation regarding noise/vibration, an eddy current power loss comparison for un laminated and laminated rotor permanent magnets, and motor design parameters and their characteristics. These results were incorporated into the final baseline design of
the prototype motors. Throughout this dissertation many of the physical dimensions, specific design parameters, and characteristics of the motor are normalized to prevent disclosure of classified information and to protect proprietary interests of the KEC design.

2.2 DESIGN CONSIDERATION

Designing and developing a high-power density, high-performance, and low-noise motor necessitates that: (1) air-gap magnetic flux density between the stator and rotor be maximized, (2) motor operating speed be optimized (using high PWM frequency), and (3) all possible losses be minimized. Because of modern technologies, these criteria are attainable, but within limits. For example, rare-earth permanent magnet materials such as Nd B Fe and Sm Co have an energy product of 30 to 55 mega-gauss oersted (MGOe). Improvements in magnetic core material (stator and rotor lamination materials), multistranded current-carrying wire technology, integrated circuits, power electronics switching device technology such as Metal oxide semiconductor field-effect transistor (MOSFET), insulated gate bipolar transistor (IGBT), and mos-controlled thyristor (MCT) are all technological advances that make it possible to develop a high-power density, high-performance, and low-noise motor. Fast and powerful computers can support detailed magnetic field calculations such as the FEA method or finite difference method (FDM), optimization algorithms, and other computer simulations to maximize the usage of materials and minimize the undesirable power losses.

In conventional brush dc motor design, increasing the magnetic force in the air gap through armature electrical current excitation is a common way to increase motor output torque. This increased current excitation results in an
increase in the conduction losses ($I^2R$) (waste energy), which appears as heat in
the windings and armature. If there are not enough heat sinks or thermal paths
provided, then the winding insulation can be degraded or permanent magnet
damage can occur. Therefore, to increase the magnetic flux by increasing stator
current excitation requires the effective removal of this waste heat from the
motor by means of active or passive thermal management systems such as: (1)
liquid flow around the stator, (2) fins on the outside of the stator, or (3) a built-in
fan on the rotor. Thus, magnetic force loading in conventional brush motors is
typically a compromise between motor weight, operating temperature, and
overall efficiency.

There are three ways to increase motor output power in a brushless,
permanent magnet motor configuration. First, high-energy density permanent
magnets can be used to increase the magnetic force in a motor and thereby
increase its power output without adding to its heat load or increasing its weight.
The increased magnetic force in the air gap, with reduced loss in the permanent
magnet motor, translates into higher power density and higher efficiency.

The second way to increase motor output power with given weight and
volume constraints is to increase the relative operating speed between the rotor
and the stator. However, the resulting high motor speed is not compatible with
many drive applications and therefore requires a speed-reduction mechanism,
such as adding a gearbox to reduce the motor speed and increase the output
torque. The disadvantages of operating at a high speed to increase power output
are: (1) the higher eddy current and hysteresis power losses in rotor permanent
magnets and in the stator lamination material and (2) the requirement for high-
switching frequency of the PWM circuit with additional high-frequency
switching losses. The negative effects that result from adding a speed-reduction
device are increased weight, complexity, losses, and additional noise/vibration. Since vibrations caused by rotating imbalance increase as the square of the rotational speed, precision balancing of the motor and speed-reduction device components becomes crucial in noise-critical applications, such as computer peripherals, medical equipment, color laser printers, and underwater applications [6,10,17]. In addition, the speed reduction device, such as gear meshing, typically produces high-frequency noise components that can be difficult to attenuate.

The third method to increase the output power is to design a motor with a higher magnet pole number, which in turn reduces the flux-per-pole that the back iron material must maintain. Because the stator back iron is usually sized to operate just below the point of magnetic saturation, reducing the flux-per-pole allows the total amount of back iron to be reduced. A drawback of increasing the magnet pole, however, is the increase in eddy current losses in the stator and rotor back iron, permanent magnets, and stator phase winding conductors. Stator back iron and permanent magnet eddy current loss can be reduced using thinner laminations. Through the careful selection of both the current-carrying conductor materials and the winding configuration, the eddy current loss in the windings can also be reduced.

Sections 2.2.1 through 2.2.5 concentrate on the design considerations for an axial field, brushless, permanent magnet motor. Five main subjects are addressed:

1. Motor configurations, their characteristics, and selection criteria;
2. Electromagnetic design concerns including air-gap flux density, operating speed, high-frequency operation, and loss control;
3. Material selection process for the stator, rotor, winding, housing, and power electronic device;
4. Thermal management method; and

2.2.1 Motor Configurations

In order to produce a high-power density, high-performance, and quiet motor, the motor configuration must be carefully selected based on the design specifications, constraints, and availability of technologies and materials. (The major advantages of an axial field motor as compared to a radial motor were discussed in Chapter 1).

Axial field, brushless, permanent magnet motors have three typical configurations: (1) single stator with single rotor, (2) double stator with single rotor, and (3) double rotor with single stator.

The double-stator with single-rotor configuration has these advantages over the other configurations. Precise positioning of the rotor between the two stators produces less axial attractive force than that of the single stator motor. The power loss is less because the rotor back iron is eliminated, which decreases the amount of magnetic material. The primary flux passes axially through each air gap and permanent magnet until it reaches the flux return path of the magnetic material (stator back iron). In this way, the motor torque and power, therefore, can be increased while decreasing the ratio of weight to horsepower.

In contrast, the double-rotor with single-stator configuration has disadvantages with regard to thermal management because the stator is positioned between rotors. In addition, stator attachment to the housing is not
easily accomplished. However, for the double-stator with single-rotor configuration, stator heat management can be accomplished by using a simple heat sink attached on the back surface of the axial field stator. Finally, a large inertia is produced by the double-rotor with single-stator configuration as compared to the double-stator with single-rotor configuration.

As a result of the prototyping process of the double-stator with single-rotor configuration, these additional advantages were identified.

1. The air-gap length can be changed by simply adding or subtracting shimming. Such modification is not possible in radial field motor design without refabrication of the rotor or stator to change the air gap.

2. The stator and rotor can be built independently of each other, thus allowing for, in some cases, the rotor diameter to be increased or decreased without rebuilding the stators.

3. The air gap of the prototype motor was also changed during the test to find the optimum noise/vibration characteristics and optimum output torque.

In conclusion, a double-stator with single-rotor motor configuration was chosen because this configuration meets all design objectives and specifications.

2.2.2 Electromagnetic Design

The electromagnetic design of a double-stator with single-rotor, axial field motor is crucial to achieving the desired high-power density, high-performance, low-noise/vibration motor. The objectives of the electromagnetic design are:
1. maximizing the air-gap flux density by using high-energy permanent magnets with maximization of the available volume and cross-sectional area and minimization of the air-gap length,

2. optimizing physical location and distribution of current-carrying windings to maximize usage of the electric current loading with minimized losses and leakage flux, and

3. minimizing stray forces such as cogging and magnetostriction forces to minimize noise/vibration.

Electromagnetic design can be performed using lumped magnetic circuit equations or the FEA method or a combination of both, depending on the desired accuracy and design goals. There are many different motor design packages available using lumped parameter magnetic circuit equations, and many technical papers discuss how to utilize these equations. Based on the study conducted, most of the available lumped magnetic circuit equations or formulas have advantages and disadvantages, which are a product of the design objectives and constraints imposed when those equations were derived. A design package suitable for a high-power density, high performance, and low-noise/vibration application was unavailable. In addition, most of the design equations are for the radial field motor, not for the axial field motor configuration. Therefore, an objective of this study was to develop a specific tool for electromagnetic design using existing lumped magnetic circuit equations [2]. Hanselman's lumped magnetic circuit equations were transferred to spread sheet programs (MathCAD, EXCEL) for their mathematical manipulation and graphical interaction. Most of the trade-off analysis conducted in this study was performed using lumped magnetic circuit equations with spread sheet programs. Further design refinement and verification was done using FEA method.
A. Air-Gap Flux Density

As discussed, the general electromagnetic design objective was to obtain maximum air-gap flux based on the available permanent magnet materials and given design specification and constraints. The stator winding is arranged to maximize utilization of that flux. Electrical current-conducting losses (I^2R) can be minimized by maximizing the slot area and the packing factor (maximum copper in motor stator). Packing of the copper is limited by the cross-sectional area between slots and teeth. A trade-off study using a simple lumped magnetic model in conjunction with FEA models is required to find the optimum ratio between the slot and tooth area. Precision phase-winding placement and balance must be maintained. Imbalance of the phase winding will result in high harmonic back emf and create a nonuniform reaction field between the rotor permanent magnets and the stator windings. This will result in improper inverter operation and difficulties in controller operation.

A starting design point for the air-gap flux density can be obtained using a lumped magnetic circuit model based on the residual magnetic flux density of the permanent magnet material and the design output torque requirement. Given a starting air-gap flux density, the initial magnetic circuit dimensions can be established based on the saturation levels of the magnetic circuit parameters and the properties of the material.

Air-gap flux density and stator back iron thickness have a direct relationship in terms of the magnetic design. Power density is affected by the thickness of the stator back iron because it contributes to total motor weight. If the air-gap flux density in the back iron of the stator can be increased without saturation, then the stator weight can remain constant. If the air-gap flux density
in the stator back iron cannot be increased, because of magnetic saturation, it is necessary to increase the thickness of the stator back iron in proportion to the increase in air-gap flux density. The weight increase would be minimal, however, because extra stator back iron is only a small portion of the total motor weight.

The stator back iron thickness effect on noise/vibration, however, is more complex. Trade-off considerations involve increasing stator back iron thickness to minimize the noise/vibration levels. Increasing the stator back iron thickness stiffens the stator back iron and reduces the axial deflections accordingly. With the given axial length as a design constraint, the stator back iron thickness can be changed by either of two ways. First, the amount of slot area and therefore total copper could be reduced resulting in higher resistive losses and less efficiency. Since the mass density between copper and back iron (e.g. M19) is nearly the same, the total weight would remain unchanged. Second, if the stator back iron thickness increases, and the slot area is kept constant, the rotor length with permanent magnet axial length must be reduced, thus reducing the total flux in the air gap with resulting efficiency loss. The change in motor efficiency would vary with the relative proportion of the resistive losses versus the eddy current or hysteresis losses. Motor efficiency would be optimum if the resistive losses are dominant (i.e., all other losses are suppressed). The noise/vibration would increase as a square function of air-gap flux density for both cogging and energized torque-produced vibrations because of the magnetic force increase function of air-gap flux density [18].

Air-gap flux density can be increased in two ways, the first of which is using higher energy density permanent magnets in the rotor. The increased air-gap flux density produces increased axial forces on the stators as almost the
square of the function of the air-gap flux. However, the increase in stator back
iron thickness provides a stiffness as the cube of the stator back iron thickness.
Thus, stator axial deflection noise/vibration may decrease. However, there are
torque perturbations (cogging torque) and harmonics superimposed on the
fundamental energized torque that also increase as the square function of air-gap
flux density. The potential total motor efficiency improvement can be obtained
by using higher energy permanent magnet material and increasing the total
number of poles in the rotor to provide the same stator back iron thickness, but
higher air-gap flux density. The trade-off between the number of rotor magnet
poles and motor efficiency is discussed in this chapter.

A second way to increase air-gap flux density is minimizing the air-gap
length. Minimum air-gap length can be calculated using a simple magnet circuit
model with FEA calculation. In axial field motor configurations, the air-gap
length can be adjusted during the motor design and prototyping stage, which is
one of many advantages of the axial field motor configuration. Capability to
change the air gap enables determination of the optimal air-gap length, which
depends not only on the magnetic design, but also on the limitations of the
currently available manufacturing technology. The optimal air-gap length
provides a reasonably safe margin of error in the air gap, which can be
prototyped in the manufacturing facility. This air-gap length can be
compensated for by increasing air-gap flux density through the use of higher
energy density permanent magnet material. This would require using a stronger
rotor structural material that can withstand the rotor warp by axial force
variation. Additionally, perfect rotor alignment would reduce axial force
variation.
B. Motor Operating Speed

Because the axial field motor that was studied had a limited battery energy source, with volume and weight constraints, it was necessary to achieve the required power output \( P = T \omega \), where \( P \) is motor output power, \( T \) is motor torque, and \( \omega \) is rotor shaft speed) by operating the motor drive at the highest speed possible. The overall weight of a motor can be reduced in direct proportion to increases in its speed of operation. However, if the motor speed is greater than that required by the mechanical load, the transmission between the motor drive and load must include a speed-reduction mechanism, e.g., a gearbox. This mechanism has its own loss characteristics and is often a source of acoustic noise/vibration. Therefore, the advantages of high-motor speed operation are counteracted by the added speed-reduction system mass, reduced total motor system efficiency, and increased acoustic noise incurred by the speed-reduction device. In the past, high-speed motors with a gearbox were used because acoustic noise was not a critical factor in most motor system applications. However, in current and future high-performance motor systems, acoustic noise/vibration reduction is an essential design objective. Thus, a gearbox or speed-reduction device cannot be utilized where quiet operation is required.

For high-speed motor designs, rotor structural centrifugal forces and high-switching frequency inverter limitations (see sections 2.3.2 and 2.4) should also be considered along with the stator lamination material. The laminations have high-magnetic permeability and strength (with the possible exception of amorphous magnetic materials), which are typically conflicting requirements. Amorphous magnetic material is a new class of magnetic material and further tests must be conducted in the area of fabrication techniques and the effects of
forces, temperature, etc. Since eddy current and hysteresis losses in stator lamination material are much higher than winding conduction current losses ($I^2R$) in high-speed operation, amorphous material may be beneficial.

In radial field, brushless, permanent magnet motors, retaining permanent magnets in the rotor at the required high peripheral speeds is a major problem. To retain the magnets at high speeds it would be necessary to use a retaining cylinder, which in turn increases the air-gap length, and thus decreases motor efficiency. However, the axial field motor design eliminates the permanent magnet retaining problem and can be used for high-speed operation with acceptable power losses, such as eddy current and hysteresis losses.

Finally, high-speed motor operation requires a means to provide accurate and fast rotor position information to the motor controller; otherwise, the motor operation will be unstable. Because noise/vibration reduction is a primary design objective in this study, the direct drive motor was adopted. In a motor application with a low-noise/vibration requirement as a key design goal, such as underwater applications, laser printers, and medical applications, a motor design consisting of direct drive with high-power density is most effective.

C. Loss Control

Examining the flow of power through an electric motor drive allows insight into the system design and operation. As shown in Figure 2.1, only a portion of the electric power available from the energy source is usable for the motor drive. Every subsystem produces its own loss that adds to the total energy required, and the variables at the interfaces between the subsystems significantly affect the performance of the overall system. The total energy source power output is given by the product of its voltage, $V$, and current, $I$. Part of this power
is converted to mechanical power, which is given by the product of torque T and motor speed, ω.

In general, 50 - 70 percent of motor power loss is proportional to P^2R loss. Eddy current, hysteresis, friction, and windage losses are relatively small. Thus, for the motor drive, the energy source should provide the highest possible voltage so that the required VI product is achieved with the least current. However, as the energy source (battery) voltage increases beyond a certain voltage, its volume, weight, and resistive loss increase rapidly. Therefore, there is an optimum operating point that is dependent upon the selection of an electric energy source and motor drive.

![Diagram of Power Flow in Motor Drive System]

Figure 2.1 Power Flow in Motor Drive System

Since the motor output power, \( P = T \omega \), of the motor drive is the product of torque, \( T \), and speed, \( \omega \), a specific output power can be achieved by producing high torque at low relative speed or low torque at high relative speed.
Fundamental motor design experience shows that when the geometry ratios are the same, the torque is proportional to the motor diameter squared for the radial field motor \( T = D^2 L \) and cubed for the axial field motor \( T = D^3 L \), where \( D \) is the motor diameter and \( L \) is the motor length [3,6]. Thus, as the available diameter increases, the ability to produce high torque increases. In addition, there is a tendency for relative motor losses to decrease as the motor diameter increases. For example, in the axial field motor, as motor diameter increases, the ratio between radial winding length to end-turn length increases resulting in decreased \( I^2R \) loss.

The performance of an axial field, permanent magnet motor can be optimized at a particular speed to achieve high efficiency. Eddy current and hysteresis losses in the magnetic materials are a significant problem when the motor is operated in a high-speed drive. In the brushless, permanent magnet motor, however, the magnitude of the rotor permanent magnet excitation is a constant magnetic field and independent of speed, but frequency changes as a function of speed. Thus, eddy current and hysteresis losses increase rapidly with increasing speed. A partial solution to the problem is to use very thin stator laminations to reduce eddy current losses. In addition to the stator iron losses, there are comparable frequency-dependent losses because of the teeth in the stator. Since the amount of power losses by stator magnetic materials is large, a slotless motor design would help reduce them. In the slotless-motor configuration, high efficiency over a wide speed range using a lightweight motor can be achieved.
2.2.3 Material Selection

This subsection presents information on the materials available for a high-power density, high-performance, and low-noise/vibration motor design and their selection criteria based on the axial field, brushless, permanent magnet motor design and prototyping. In the early stages of the design process, a material survey can be conducted prior to the electromagnetic design or preliminary design and can be performed using general material properties. Many trade-offs exist among different magnetic materials, and it is necessary to define material selection criteria in the early stage of design and prototyping. A specific priority list of factors and their relative importance to one another must be constructed.

A. Stator Material

The axial field motor stator consists of laminations that should be selected for maximum magnetic flux density handling capability and minimum eddy current and hysteresis losses. There are many stator materials available that offer various combinations of reduced eddy current and hysteresis losses, higher flux density saturation level, and lower cost. The stator lamination material selection should be made considering the shape and frequency of the phase current signal generated by the inverter/converter PWM circuit. This stator lamination material selection process results in a trade-off between efficiency, size, weight, and cost of the motor to be designed.

Figure 2.2 depicts the hysteresis loop difference between soft magnetic material and permanent magnetic material. Figure 2.3 shows a typical hysteresis loop for a soft magnetic material [19].

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Figure 2.2 Hysteresis Curve Comparison

Figure 2.3 Hysteresis Loop for Soft Magnetic Material
The saturation flux density, $B_s$, is the flux-carrying capability of the magnetic material. $B_r$, remanence flux, indicates the amount of magnetic retentivity when all external fields are removed, and $H_C$ indicates the amount of external field required to overcome remanence flux. $B_r$ and $H_C$ are usually undesirable properties in stator lamination material and should be minimized, but $B_r$ and $H_C$ are desirable properties in permanent magnetic materials. The permeability $\mu$ is the slope of the $B$ vs. $H$ curve. A higher permeability implies better efficiency in carrying magnetic flux within the same amount of magnetic material.

Amorphous magnetic materials are the best known magnetic materials having minimum core loss and high permeability and are being extensively studied by many universities and laboratories. For electric motor lamination applications, there is significant uncertainty about the use of amorphous material as a magnetic core material. The new amorphous magnetic materials will require many years of feasibility studies and development.

Vanadium permendur stator lamination material has the highest saturation flux density among magnetic core materials, thus reducing the required amount of magnetic material. Vanadium permendur lamination material is high in price, difficult to procure, and sensitive to thermal and mechanical forces. The advantage of using vanadium permendur lamination material is that it does not oxidize and it has a high-corrosion resistance. For these reason, vanadium permendur lamination material is used primarily in high-power density and high-performance applications.

Cold-rolled steel, also called low-carbon steel, saturates at about 20 kG and therefore requires approximately 10 to 15 percent more magnetic material
volume (and mass) than vanadium permendur to carry the same amount of flux. Cold-rolled steel is low in cost, easy to obtain, and readily machined. It must be plated, however, to prevent rusting, a process that can present difficulties in manufacturing to close tolerances.

Stainless steel (for example, 416 SS) saturates at about 17 kG, leading to a 30-percent increase in magnetic core material weight compared to the weight of vanadium permendur; but it is relatively inexpensive, easily obtained, and is easier to machine than vanadium permendur, although not as easily machined as cold-rolled steel. The primary advantage of stainless steel is that it requires no plating or protective coating to prevent oxidation [19,20].

There are many grades of silicon electrical steel with saturation levels of 18kG to 20kG, such as M15, M19, M36, M43, all of which have been especially formulated for many ac electric motor designs, such as induction motors and synchronous motors. The higher M numbers (and, thus, higher core losses) are progressively lower in cost, although only a few percent is saved with each step down in performance [20]. These alloys, when properly heat-treated, have a high permeability, a low coercive force, and a saturation flux density nearly equal to those of cold-rolled steel.

The cost of the magnetic material for stator lamination, relative to the total production cost of the motor system, is also a very important factor in the material selection process. Moreover, the core loss function of flux density and applied field frequency is another factor. The material with sufficient power density and performance to meet the design objectives of this study was found to be vanadium permendur because of its maximum flux density-handling capability and minimum core loss characteristics.
B. Permanent Magnet Materials

Subsection 2.2.3 B discusses permanent magnet materials, characteristics, and basic terms. Figure 2.4 shows a typical permanent magnet material B/H curve. The M/H curve is the intrinsic demagnetization curve. Once a permanent magnet material has been magnetized, it remains magnetized even if its magnetic field intensity (or externally applied magnetic field) is decreased to zero. The magnetic flux density at this point is called the residual magnetic flux density or the magnetic remanence (B_r). Furthermore, when the magnetic field intensity is increased in the opposite polarity along the demagnetization curve, the magnetic flux density eventually becomes zero. This field intensity is called the magnetic coercive force or the coercivity (H_cB). The absolute value of the product of the flux density, B, and the field intensity, H, at each point along the demagnetization curve is called the energy product. The maximum energy product is one of the indexes of the strength of the permanent magnet material.

There are basically three different types of permanent magnets that are used in permanent magnet motors: alnico, ferrite or ceramic, and rare-earth magnet materials. More detailed information on these materials and their applications can be found in References [21,22,23].

The normal second quadrant of B/H curves of these three types of permanent magnets vary widely as shown in Figure 2.5. The alnico magnet provides a fairly high remanence flux density, but a low coercive force. When the coercivity is low and two opposing magnetic poles are in proximity of each other, the magnetic poles can weaken each other and there is a possibility of
Figure 2.4 B/H Curve for Typical Permanent Magnet Material

Figure 2.5 Demagnetization Curves of Permanent Magnet Materials
permanent demagnetization by the opposing field. Therefore, an alnico magnet is used after being magnetized lengthwise. Unlike an alnico magnet, the ferrite magnet has a low flux density but a high coercive force. It is possible to magnetize the ferrite magnet across its width because of this high coercive force. Ferrite magnets are widely used in electric motors because their material and production costs are low.

Rare-earth magnets have both high magnetic remanence and high coercive force. Since the initial cost is high, these permanent magnets are used in applications such as high-performance and high-energy density motor applications. For a given volume, the flux density is twice that of the ferrite, leading to a larger torque production. Figure 2.6 summarizes the evolution of the permanent magnet materials at 20 ºC [22]. The strength and structure of the magnetic field in brushless, permanent magnet motor technology is mainly related to the type and the maximum energy product of the permanent magnets used. As seen in Figure 2.6, Nd B Fe magnetic material is superior to any other magnetic material now on the market. The only disadvantage of using an Nd B Fe magnet, as opposed to an Sm Co magnet, is that the high-energy density Nd B Fe permanent magnets have a maximum operating temperature of 100 to 150 ºC as compared to 200 to 300 ºC for Sm Co, alnico, and ferrite.

Development of the magnetic energy product was studied using different manufacturers' data. The analysis shows a clear trend toward higher energy density close to the theoretical limitation of the Nd B Fe magnetic family. For these reasons Nd B Fe permanent magnet material was chosen for the prototype motors in this study.
C. Current-Carrying Wires

Force produced in electric motors is simplified using the Lorentz formula, $F = I \, dL \times B$, where $F$ is the mechanical force, $I$ is the current flow in conductor, $dL$ is the differential length of current-carrying conductor, and $B$ is the magnetic flux density around the current conductor. Maximum force can be obtained when the current-carrying conductor and magnetic flux field are orthogonal to each other. In a brushless motor, the $B$-field is provided by rotor permanent magnets, and $I \, dL$ is supplied by stator windings. It is important to select the best current-carrying wires for stator windings to minimize current conduction losses ($I^2R$) and eddy current losses.
The common electric-conducting materials used in current-carrying wires for motors and electric machines are copper, aluminum, and copper-clad aluminum conductors. The most widely used conductor is solid copper wire because it is easy to wind and is readily soldered onto lead wires or terminal pins.

Aluminum wire tends to be brittle and difficult to work with. It is also difficult to terminate aluminum wire because it does not solder well. Moreover, solder joints with aluminum wire are subject to corrosion and are unreliable. Cladding aluminum wire with a thin layer of copper can reduce the termination difficulties, but the mechanical brittleness of the wire creates manufacturing problems.

Figure 2.7 shows a variety of wire types, and Figure 2.8 depicts different multistranded wires. For some very high temperature motor applications, a hollow-conductor copper wire can be used so that a cooling fluid can be pumped through the center. Multistranded wire is useful for the high-speed, high-frequency, brushless motor design and is used in the prototype motors studied in the research for this dissertation. The use of multistranded wires, compared to conventional solid copper wire at the same rating, in the advanced high-power density motor design is expected to have these improved results:

1. Lower winding losses,
2. Reduced turn insulation,
3. Improved winding stacking factor, and
4. Easy bending and winding.

Round cross-sectional electric conductor wire is the most common wire shape because it is easy to manufacture, insulate, and wind. Square or
rectangular wire is used in high-power density motor design when maximum utilization of the slot area is mandatory and in layer-wound coils, such as those in voice coil actuators and wound inductors, in which the maximum conductor packing factor is required. Improved motor performance of 7 to 10 percent can be achieved with rectangular wire, compared to round wire, because of the increased slot packing factor of the conductor within the winding volume (Figure 2.9). Rectangular wire, however, cannot be twisted and is prone to dielectric breakdown when wound around corners, resulting in a reduced reliability when compared to round wire. A special winding fixture is also required to insert the wires in stator slots.

![Solid Wires with Insulation](image1.png)

![Hollow Wires with Insulation](image2.png)

Figure 2.7 Different Wire Types
Figure 2.8 Multistranded Wire Types

Figure 2.9 Comparison of Round vs. Rectangular Wire
In high-performance, high-power density motor design, the reduction of conduction current ($I^2R$) loss is a critical design issue. Therefore, a low-resistivity conductor should be used for minimum resistance to current flow, which minimizes $I^2R$ loss. In high-speed or high-frequency motor operation, a trade-off between conduction current loss versus conductor eddy current loss is necessary in most cases. For example, a simple comparison can be made between solid wire and multistranded wire. The solid conductor wire with a round or rectangular shape has maximum cross-sectional area so it has less resistance and results in less conduction current loss compared to the loss with multistranded wire in a low-frequency application. As the frequency increases, the ac resistance of the solid wire increases at a faster increment than that of multistranded wire. Resistance values of multistranded conductor wire and solid wire were experimentally examined in this study. Experimental measurements were done with solid, 9-gage, round conductor and 7x 20-gage, multistranded, conductor wire (New England Wire Co). Figure 2.10 shows the wire ac resistance versus frequency applied using the resistance test data. The results show conduction current loss can be minimized by using multistranded conductor wire, which eliminates the skin effect and thus reduces eddy current losses.

Current-carrying wires should be electrically insulated to prevent turn-to-turn shorts in a slot coil winding. A wide variety of polymeric insulation types with thermal ratings from 105 °C (e.g., Polyurethane) to 220 °C (e.g., Polyimide) are available. Factors affecting the choice of wire insulation include operating temperature, exposure of the winding to chemicals, wire bonding and soldering techniques, and cost.
Figure 2.10 AC Resistance Test Results

Figure 2.11 shows the evolution of different wire types used in the prototype motors to reduce eddy current and conduction current power losses. The first wire used was a common, solid, round wire. The second wire used was a square wire that increase the slot copper-packing factor to reduce the resistance and thus reduce conduction losses. (Increased copper material reduces dc conduction current losses.) The third wire used was a flat, rectangular wire designed to reduce slot leakage flux eddy current losses in the conductor. The fourth type of wire used was multistranded wire whose purpose was to reduce x- and y-directional flux eddy current losses.
Figure 2.11 Evolution of Current-Carrying Wires Used in Prototype Motor

Based on the comparison study and test conducted, the multistranded wire was used in the prototype motor (unit 2). Not only were there ac resistance loss and eddy current loss reductions, but the multistranded wire provided better compact winding in end-turn areas.

D. Rotor Structure

The rotor matrix in axial field motor with a double-stator with single-rotor configuration is a simple disc web where the rotor permanent magnet can be plugged in. Therefore, the rotor structure can be made with an electrical insulator, such as the epoxy/glass laminates currently used for flywheels. The
advantage of using an electrical insulation material is that there is no eddy
current power losses from the stator excitation. To support high-power density,
high-performance, and quiet operation design objectives, a high-strength
material with a nonmagnetic property, such as stainless steel, titanium, or
aluminum is commonly used. The strength of the materials is a primary
consideration because it must withstand centrifugal radial force and
electromagnetic axial forces. When high-energy density, rare-earth, permanent
magnet materials are used, the axial force is also higher. Therefore, rotor matrix
material and size must be selected based on the axial force calculation. In the
prototype motor, the rotor structure was machined using titanium material
because of its high strength and stiffness to weight ratio. Titanium is strong
enough to withstand the effects of vibration and noise.

E. Motor Housing

The efficiency, weight, heat conduction, and noise/vibration of the motor
are affected by the characteristics of the housing material. The most common
materials used for high performance motor housings are aluminum, stainless
steel, and titanium. Housing material should be based on the motor design
specifications and objectives. Material selection is also influenced by fabrication
techniques, availability, and cost. For example, aluminum offers reduced weight
and is a good thermal conductor, but it is subject to corrosion. Table 2.1 ranks
three materials (in order of preference, 1 being most preferable and 3 being least
preferable) for their suitability in satisfying various design objectives.

Aluminum material was selected for the prototype motor housing in this
research based on these considerations: weight, machineability, and thermal
conductivity.
Table 2.1 Ranking of Motor Housing Material

<table>
<thead>
<tr>
<th></th>
<th>Aluminum</th>
<th>Titanium</th>
<th>Stainless steel</th>
</tr>
</thead>
<tbody>
<tr>
<td>Weight reduction</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>Thermal conductivity</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>Manufacturing cost</td>
<td>1</td>
<td>3</td>
<td>2</td>
</tr>
<tr>
<td>Strength to stiffness ratio</td>
<td>3</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>Noise/vibration reduction</td>
<td>3</td>
<td>1</td>
<td>2</td>
</tr>
</tbody>
</table>

F. Power Electronic Devices

In most brushless, permanent magnet motor designs, a solid-state inverter is used to convert the input dc into multiphase, variable frequency ac. The primary advantage of this approach is that it allows the number of poles to be optimized and it allows the stator current drives to be tailored for maximum efficiency. For example, the higher number of magnet poles require less stator back iron, and the exact current shape and frequency matching with back emf provide ripple-free torque (theoretically).

The principal elements in a solid-state inverter are the power electronic switches that turn the current on and off under direction of the motor control logic. For high-power applications, solid state switches are typically bipolar, Darlington, MOSFETs, or IGBTs.

Important characteristics in selecting the switches for a specific application include their voltage and current rating, conduction voltage drop, switching speed, and base drive requirements such as short-circuit protection and gate-drive circuitry. In some applications, the base drive circuitry necessary to control and protect the power switches is complex, including snubber and surge
suppressers to prevent voltage transients from destroying the semiconductor switch devices.

Solid-state switches have evolved rapidly in the past ten years to meet the needs of a wide variety of power circuit applications. New devices such as MOSFETs and IGBTs that are capable of switching high currents at very high frequencies (up to 400 amps at 50 kHz) have become available. IGBTs combine the high-speed and low-gate-drive requirements of FETs with the high-current and low-conduction drop characteristics of Darlington transistors.

To build a compact, light weight inverter with high switching speeds, the IGBTs are the most suitable device and are currently available off-the-shelf. The IGBTs were used in both prototype motors (units 1 and 2).

2.2.4 Thermal Management

All motors must deal with the elimination of the waste heat that is generated by conduction losses in current-carrying windings; eddy current / hysteresis losses in the stator magnetic material; eddy current in the rotor permanent magnet and rotor structure; switching losses in the power inverter/ converter; and controller circuit losses. The first step in thermal management is to minimize the heat generated by each of these sources in the initial design stage and to provide aggressive thermal management procedures to remove the heat.

One of many motor design considerations is the temperature rise in the stator, which is a function of slot and outside dimensions, electric current loading, and stator lamination material. All heat generated inside the motor must be conveyed through the motor frame to ambient. The major heat
conduction path is from the stator current-carrying winding to the stator core steel. However, eddy current losses in the rotor frame and permanent magnet are significant in cases in which high-energy density, rare-earth permanent magnet materials are used. Thus, it is important to consider the rotor eddy current heat generation. The heat generated in the rotor or transferred to the rotor will be conducted through the rotor (permanent magnet) structure and bearings to the motor housing. When selecting the bearings rotor heat generation, therefore, must be considered. Heat is also convected across the air gap from the stator to the rotor, or in some cases from the rotor to the stator. It is important to identify heat sources and paths to control and manage waste heat from the motor. Thorough investigation can be conducted using any FEA thermal analysis package.

End-turn winding thermal management is one of many important issues since its heat generation results in lost energy, which is a significant portion of the total motor heat loss. (In some cases, this loss is 15 to 25 percent of the total heat loss.) It is important to minimize end-turn loss by using shorter end-turn windings or to provide other means to eliminate the heat loss. Different end-turn winding techniques and imbedding in thermally conductive materials are being studied.

The simplest and most widely used method to provide heat rejection from the motor is heat sink fins on the outside of the stator or motor housing. However, these heat sink fins add weight and increase the volume of the motor. Some motors are designed to have rotor blades that can blow the surrounding air through the air gap between the stator and rotor. This method can be used in the motor without noise as a design consideration factor.
The most efficient and lightweight design involves using water to infiltrate the motor. This design, for short-life applications, puts the sink (water) in direct contact with the source (stator or rotor). For longer life applications, the motor can be filled with oil and sealed. The most inefficient extended-life design currently being used is where a complete motor is repackaged inside another housing and then filled with oil. In this case, weight is increased and efficiency reduced due to the extra packaging and the long thermal path.

High-power density and high-performance motor design requires some form of forced-fluid cooling. Many use water-cooling loops that are close-coupled to the stator back iron and inverter switches to provide minimal thermal resistance to heat transfer. In prototype motors a cooling plate with a liquid flow tube is attached on each stator back iron to eliminate the waste heat from stator. There are many design and fabrication issues, such as how and where to attach the cooling plate, to be considered in thermal management systems.

2.2.5 Noise/Vibration

One of the many advantages to brushless, permanent magnet motors is the potential, in a properly designed system, for extremely quiet and smooth operation [6,18]. When considering a motor system design with low-noise/vibration as a performance requirement there are system trade-offs between the electromagnetic design, the PWM controller design, and the structural analysis that have a marked effect on the system noise/vibration performance. Several sources of noise/vibration may be linked to performance requirements that are in opposition to their noise-optimal operating conditions. System-level trade-offs require knowledge of noise/vibration sources over the dynamic range under consideration. Accurate estimates are required regarding
component noise/vibration source level and its means of transmitting acoustic energy to the surrounding medium.

The major sources of noise/vibration in a typical axial field, brushless, permanent magnet motor system are the torque ripples, mechanical imbalance of the rotor, stator lamination material humming, and rotor bearings. Two torque ripples, cogging torque and energized torque, are produced as the rotor permanent magnet passes the stator teeth and phase windings. Magnetostrictive noise similar to transformer hum is also evident in motor vibration. The rotor structure potato-chipping effect becomes a problem only when there is an axial force imbalance.

Noise/vibration from the cogging torque is the result of very powerful permanent magnets producing alternating magnetic pressures through the air gap to the stator structure. This alternating magnetic force between the rotor permanent magnet and stator teeth produces strong noise/vibration at the tooth or slot. The magnitude of this cogging force can be observed by turning off all the stator excitation currents and rotating the rotor by external force. This cogging force is a characteristic of brushless, permanent magnet motors, and its reduction is a major focus of motor-quieting efforts. Techniques were studied to reduce this cogging torque ripple through geometric variations in the magnet and tooth topologies (Chapter 4). Those discrete vibrations that exist are easily identified and would be ideal candidates for active isolation techniques.

Mechanical vibration is caused by imbalances in the rotating components of the motor, rotor, bearings, shaft, and couplings. Any asymmetry in mass about the axis of rotation results in an oscillatory force that is directly proportional to the mass imbalance and its offset from the axis of rotation, which
is proportional to the square of the rotational speed. Such an imbalance results in a once-per-revolution vibration that is applied by the rotating member to its stationary support structure. In addition to this once-per-revolution vibration, higher order harmonics can be excited at integral multiples of the shaft rotation speed. Thus, good manufacturing practice requires precision balancing of all rotating components. Significant additional vibration reductions can also be achieved by limiting the rotational speed.

Cogging and energized torque vibrations have a fundamental frequency based on the repetitive passing of rotor poles and stator slots as the motor rotates (see Chapter 4 for details). Generally, both types of vibration can be treated during design by shaping and skewing the magnetic poles. It is critical to find an optimum permanent magnet skewing ratio using a developed spread-sheet design package/FEA package combination.

There is also a suggested technique to control the variations in output torque by using a combination of PWM circuit and current feedback control to tailor the waveform of the current applied to each phase of the stator at each rotor position. The tailored current generates a stator torque that virtually cancels the effect of cogging torque between the rotor poles and stator teeth. The tailored waveforms, generated from FEA during the design, are stored as digital data in read-only memories in the motor controller. Waveform data are read out to the PWM drives in the inverter as a function of rotor position, which are continuously monitored by means of a shaft-mounted position sensor, such as an encoder or resolver with resolver-to-digital circuit.
2.3 PROTOTYPE MOTOR

The prototype motor is a brushless, permanent magnet, disc-type axial field motor with dual air gaps. A simplified double stator with single rotor, axial field permanent magnet motor is shown in Figure 2.12. Figure 2.13 shows a cutaway view of the prototype motor. Figure 2.14 depicts the prototype motor stator picture with phase windings using multistranded wires. (Multistranded wire offers reduced \( I^2R \) power losses, improved winding stacking, and easier bending and winding.) The magnetic field is supplied by permanent magnets embedded in the rotor as shown in Figure 2.15. The rotor structure is made of a nonmagnetic material such as titanium. The stators are radial arrays of current-carrying conductors with appropriate end turns and connections fixed in the stator slots. The wound stator discs are fixed to the motor cooling plate and to the motor housing (see Figure 2.13).

The motor configuration in Figures 2.12 and 2.13 shows that the air gap between the rotor and stators is a plane perpendicular to the axis of rotation. Permanent magnet flux crosses this air gap in the axial direction. There are 48 permanent magnet poles in the rotor. Each permanent magnet pole covers three teeth on the stator. Since the permanent magnets must be placed with the pole faces at the air-gap surface, the air-gap flux density is proportional to the permanent magnet working flux density. Therefore, using high-remanent flux density magnet materials, such as Nd B Fe magnets is advantageous in generating high torque in the prototype motor configuration.
Figure 2.12 Simplified Axial Field, Permanent Magnet Motor
Figure 2.13 Cutaway View of Axial Field Permanent Magnet Motor
Figure 2.14 Prototype Motor Stator
Figure 2.15 Prototype Motor Rotor with Laminated Permanent Magnets
The rotor thickness in the axial direction is equal to the permanent magnet length, which must be sufficient for the permanent magnet to generate a magneto motive force (MMF) that is capable of producing a large flux across the two air gaps. Permanent magnet materials with high coercivity lead to shorter magnet lengths and hence to smaller rotor volumes and weights.

Because the flux is entirely axial, both in the air gap and in the permanent magnets, this configuration removes the need for magnetic back iron in the rotor to conduct the magnetic flux from pole to pole. This elimination of flux return paths on the rotor and subsequent reduction of the overall rotor volume and weight constitute one of many advantages of the dual air-gap, dual-stator arrangement. This low-inertia configuration, coupled with a rotor and strong permanent magnets, can be driven by power electronic switching devices with PWM inverters to provide desirable features such as high-power/weight ratio, high-torque/current ratio, and torque/inertia.

Efficiency is high over the entire operating speed range because there is no conducting current loss and back iron core loss in the rotor compared to that of the radial (conventional) field motor to prototype motor; total motor weight is low because the heavy back iron in the rotor is eliminated; commutating inductance is less because of the large magnetic air gap as compared to conventional radial field motors. In the prototype motor, the magnetic air gap is large compared to that of the conventional radial field motor because the prototype motor (dual air gap, double stator with single rotor) does not have rotor back iron and the permanent magnet material has the same permeability as air. This permanent magnet disc rotor can be built for speed operation higher than that of earlier dc motors with windings on the rotor, thus increasing power
density of the motor. Finally, because there are no commutators, slip rings, brushes, or exposed end windings, no maintenance is required and reliability is higher than that of brush-type or conventional radial field motor configurations.

2.3.1 Stator

In the axial field, brushless, permanent magnet motor, the stator has the following characteristics.

1. The stator slots and phase windings are radially oriented.
2. The slot width at any diameter is constant over its full depth.
3. The ratio of slot width to tooth width varies throughout radial direction.
4. The stator return path is circumferential because the primary air-gap flux is in the axial direction.
5. The stator diameter is a function of the active length of the current-carrying conductor requirement, the number of conductors (since this effectively determines the inner diameter), and the end-turn bundle packing technique.
6. The axial stator length can be approximated by the electric-loading capability represented by the slot depth and the thickness of the flux return path of the stator magnetic circuit.

The laminated stator is ideally suited for production from a continuous strip into a tightly coiled spring form as shown in Figures 2.16 and 2.17. Figure 2.18 and Figure 2.19 depict a side and 3-D view of the stator with slots and laminations.

In most radial-configuration motor stator laminations, stampings (using punch and die techniques) that have been produced from fully annealed lamination material, such as low-carbon electric steel, generally require only a stress-relief anneal to return the magnetic material to its original magnetic state.
A coating for the lamination can then be applied to increase interlaminar resistance to reduce eddy current flows. There are materials available whose surface coating can withstand the annealing temperature and still provide high interlaminar resistance to eddy currents. Extensive inquiry, however, found that such a coating was applied only to oriented magnetic materials, and thus raised the question as to whether the lamination material, which has been obtained in a fully annealed condition, would require the final annealing process. If this final operation is necessary, using a continuous strip method to produce the stator core lamination in the axial field configuration requires a high temperature interlaminar coating. After the coil has been formed and machined, it would be very difficult to unwind after annealing to coat the surface. Therefore, a high-temperature coating material is a crucial item to be considered in the axial field stator core lamination.

Figure 2.16  Wound Metal Ribbon Coil before Machining the Slots
Figure 2.17 Top View of Stator

Figure 2.18 Side View of Stator
2.3.2 Rotor

The rotor consists of the rotor structure and permanent magnets embedded in that structure. As shown in the Figures 2.12 and 2.13, the primary magnetic flux passes axially through each gap and rotor permanent magnet until it reaches the stator back iron, which is a flux return path. The flux in the stator back iron turns in the circumferential (tangential) direction and returns to the air gap under the next pole. The flux return paths are therefore required only at the ends of the stators. Because there is no back iron in the rotor, the capability of the motor torque can be increased by adding more permanent magnets and using stronger and longer permanent magnets in the axial direction.

A prototype rotor structure made of titanium material is shown in Figure 2.20 without permanent magnets. The rotor matrix must be a nonmagnetic material such as stainless steel, titanium, or aluminum. Because the rotor matrix must withstand both centrifugal force (when the rotor rotates at a certain speed) and axial magnetic forces (such as attractive force and cogging force), the
Figure 2.20 Prototype Motor Rotor Structure
strength of the rotor matrix is a primary consideration. If the rotor matrix structural strength is not sufficient to withstand the axial attractive force between the rotor permanent magnets and the stator steel as the rotor is rotating, the rotor matrix will be warped. In some extreme cases the rotor surface could abrade the stator surfaces. Additionally, it is important to examine the structural strength of the rotor matrix based on the selection of the rotor permanent magnet material and calculation of the attractive axial force between rotor permanent magnets and stator steel.

If the rotor structure is a nonmagnetic, but electrically conducting material such as stainless steel, titanium, or aluminum, normally there are induced eddy currents except at very low commutation motor speeds. Such eddy currents may reduce commutation inductance, but they are a source of additional power losses. A common method to reduce eddy current power losses is lamination of the rotor matrix and permanent magnet materials. In this study the permanent magnet was laminated; the rotor matrix, however, was not laminated because of the geometry of the matrix and the low conductivity. Table 2.2 shows the four most common permanent magnet materials used in motor design with their properties summarized using information obtained from manufacturer's catalogs. As seen in this table, compared to the other three magnetic materials, the Nd B Fe has the highest energy product. It also has the lowest resistivity and lowest operating temperature. When Nd B Fe magnets are used, eddy current reduction and proper thermal management must be considered at the design stage.
Table 2.2 Permanent Magnet Material Properties

<table>
<thead>
<tr>
<th></th>
<th>Nd B Fe</th>
<th>Sm Co</th>
<th>Alnico</th>
<th>Ceramic</th>
</tr>
</thead>
<tbody>
<tr>
<td>Energy Product</td>
<td>25-48</td>
<td>16-31</td>
<td>5-11</td>
<td>1-4</td>
</tr>
<tr>
<td>(MGOe)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Maximum</td>
<td>150</td>
<td>300</td>
<td>550</td>
<td>300</td>
</tr>
<tr>
<td>temperature</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Density</td>
<td>0.27-0.28</td>
<td>0.30-0.50</td>
<td>0.25-0.30</td>
<td>0.10-0.18</td>
</tr>
<tr>
<td>(pound/in³)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistivity</td>
<td>0.15-0.16</td>
<td>0.5-0.6</td>
<td>0.4-0.5</td>
<td>3.0-5.0</td>
</tr>
<tr>
<td></td>
<td>(mΩ cm)</td>
<td>(mΩ cm)</td>
<td>(mΩ cm)</td>
<td>(mΩ cm)</td>
</tr>
</tbody>
</table>

There are other methods to prevent eddy current power loss in the permanent magnet material. A thin aluminum or copper cover sheet placed on the top of the rotor surface lowers commutating inductance with low eddy current power loss. This thin sheet of conducting material acts as a high-frequency shielding, one of many ways to reduce the eddy current effect on the permanent magnets. The permanent magnets are also partially protected against demagnetization as a result of PWM inverter electromagnetic field shoot-through. A similar effect can be achieved by surrounding the permanent magnets with aluminum or copper sheaths. There are some disadvantages to surrounding the permanent magnet with highly electrical conductive material, such as the damping effect by created eddy current in the sheet.

It is important to note that while the air-gap flux density is limited by the remanent flux density of the permanent magnets, the rotor volume is directly proportional to the permanent magnet length and thus to the remanent flux density of the permanent magnets. The rotor volume in the axial field
configuration is inversely proportional to the energy density of the permanent magnet material used because the rotor axial length is equal to the permanent magnet length. This characteristic of the axial field configuration makes the use of high-energy density rare-earth Sm Co or Nd B Fe type permanent magnets, as opposed to ferrite (ceramic) or alnico types, especially appropriate.

There are a variety of high-energy density permanent magnet materials now available for use in high-performance, high-power density motors. New permanent magnet materials are currently under development. At this time, the best choices are Sm Co and Nd B Fe. Using ferrite and alnico in this configuration will result in an unacceptably large motor. In this study, trade-offs of different rare-earth magnet materials are studied. Study results show that the permanent magnet selection is dependent not only on the output power requirement, but also on thermal management; refer to Table 2.2.

Many magnetic materials are available in the form of cylinders or rectangular and trapezoidal shapes. It can be demonstrated that shaping the magnets to match the current-carrying windings yields a significant advantage in net flux per pole [8,11,12,14].

2.4 DRIVE TECHNIQUE

Figure 2.21 shows a drive system block diagram of the dual air-gap, axial field prototype motor studied in this research. The motor drive system can be divided into three major components: (1) the controller circuit, (2) the power electronic circuit and (3) the prototype motor with a rotor position sensor. The controller circuit is composed of a central processing unit (CPU) - microprocessor, a waveform generator, and PWM circuits. The CPU circuit calculates a motor shaft torque command based on a speed command and
Figure 2.21  Drive System Block Diagram
measured rotor shaft speed. The waveform circuit generates reference current waveforms for the PWM circuits based on the torque command and the rotor position. The PWM circuit calculates the switching instances for the second motor drive system component, which is the power electronic stage.

The power electronic circuitry includes gate drivers and H-bridge that react with the PWM circuitry. The third component is the prototype motor with rotor position sensor.

A high-power density, brushless motor design is feasible with the advent of high-power-capable metal oxide semiconductor (MOS) switching devices, i.e. IGBTs. The controller size can be dramatically reduced by employing components such as programmable array logic (PAL) devices. These devices combine multiple discrete logic chips into a single chip configured by software, thus, significantly reducing board size and parts count and adding flexibility and increased reliability to the controller design.

2.4.1 Controllers

The control subsection is responsible for implementing commands for motor action. In addition to external motor commands, the control subsection relies on inputs from the power electronics and the rotor position sensor for determining what actions to take. Primary commands are implemented by controlling the on and off states of the power electronic devices in the power electronics circuitry. Although motor control is implemented differently for different motor types, the fundamental action taken by the control subsection is to control the motor current. For all motors, the torque produced is a function of the current that flows in the motor windings. The control of motor current using microelectronics enables instantaneous control of torque to the mechanical load.
Digital control, implemented with microcontrollers or digital signal processor (DSP) integrated circuits, enhances the optimization of motor performance. For example, the motor current can be modified in real time to eliminate torque ripple for acoustic signature reduction [17]. Another example is adaptive speed control to eliminate the nonlinearity attributed to the mechanical load characteristics.

2.4.2 Power Electronic Devices

The purpose of the power electronic devices in a motor drive is to convert the input electrical energy into the form required by the motor system. Different motor types require different power electronic subsystems. Common to all high-power density, brushless motor drives is the requirement for switching large currents at high voltages into inductive motor windings. The term switching means turning on and off in a controlled manner.

Many power electronic devices have been developed over the past twenty years to meet a variety of requirements. Using the catalogs published by the manufacturers of power electronics devices, the author has compiled information in the relationship between device capacity and frequency (see Figure 2.22). Over the past decade, the development of MOS power devices has led to a revolutionary improvement in performance and has generated a number of new applications and capabilities that were unachievable with prior devices such as silicon controlled rectifiers (SCR) and thyristors. The capabilities of MOS devices can be summarized as follows:

1. Device current gain is high (>100), typically an order of magnitude or two greater than earlier bipolar devices;
2. Device switching speed is high (>25 kHz) at high current levels; and
3. Device switching and conduction losses are much lower than in bipolar devices.

![Figure 2.22 Power Semiconductor Family Ratings](image)

These capabilities have resulted in dramatic improvements in the power density, efficiency, and reliability of the power electronics subsection of motor drives in electric motors. Moreover, they have opened the door to the implementation of sophisticated control algorithms for improved system performance.

At this time, low power MOSFETs are used in low voltage (<200 volts) applications, and IGBTs are used in applications where higher power capability is required. It is anticipated that MCTs will become available in high-power ratings within the next five years. At that time, the use of IGBTs will decrease
because MCTs will offer similar power-handling capabilities at higher efficiencies.

2.4.3 Rotor Position Sensors

The capability to produce electronic commutation at precise instants is essential to high motor performance. There are three types of devices used for shaft position sensors: (1) resolvers, (2) hall devices, and (3) optical encoders. In each case, output signals of square-wave pulses (digital signal) are required as input to the electronic commutation logic network. In this study, the temperature, accuracy, weight, and volume for the three position devices were compared. Because of the possible magnetic field interference between the rotor permanent magnet and hall effect device, the hall device was eliminated as a position sensor candidate. The optical encoder and resolver, selected as possible candidates, were considered to two prototype motors, and it was found that there are some trade-offs between these two sensor technologies. Because of the higher thermal rating, higher system reliability, and more accurate shaft position information, the resolver was found as the best position-sensing device.

2.5 SUMMARY

This chapter described the axial field prototype motor in detail. Specifically, this chapter discussed the critical motor components and design issues of the prototype axial field permanent magnet motor and the advantages of using the axial field motor configuration instead of using the conventional radial-field motor configuration. Additionally, this chapter addressed the stator manufacturing process, including stator lamination and interlaminar coating; rotor structure and permanent magnet materials required to construct the axial field motor; and drive techniques of the designed axial field dual air-gap motor.
CHAPTER 3

MOTOR PARAMETER CALCULATION

3.1 INTRODUCTION

In this chapter, several important motor design parameters and performance characteristics, including flux distribution, flux linkages, back emf waveform, inductance, and energized torque waveform, are calculated. The purposes of the motor design parameters and characteristics calculation are to (1) model the electromagnetic design, (2) improve the current prototype motor design, (3) integrate with the controller design, and (4) to develop a motor design tool for future motor designs.

This chapter details how to calculate and characterize the motor parameters and characteristics in a brushless, axial field permanent magnet motor. The two methods used in this study are FEA analysis and lumped parameter magnetic circuit analysis. Each method of calculation is presented briefly in the subsection 3.1.1.

A two- and three-dimensional FEA package MSC/EMAS (The MacNeal-Schwendler Corporation/ElectroMagnetic Analysis System) with preprocessor and postprocessor (MSC/XL) program is used in this study.

3.1.1 Method of Analysis

To predict the designed motor parameters and performance characteristics, it is necessary to calculate the magnetic field throughout the
motor structure. There are many kinds of field calculation methods, and each method has a different level of accuracy.

Both 2- and 3-D FEA models are used to analyze the axial field permanent magnet motor considered in this study. Using these two types of FEA models permits the inclusion of intricate physical shapes of the stator and rotor and nonlinear magnetic material properties for an accurate analysis. The main disadvantage of using FEA methods is the excessive engineering effort and computer resources required to set up the FEA model and to analyze the results. To reduce such problems, mesh generation, solution process, and result calculations of the FEA modeling and analysis have been mechanized using the generic FEA modeling capability program discussed in subsection 3.2.2. This program package alleviates the need to create a new mesh for each geometric or rotor position change.

The primary objectives of the FEA method used in this study are to investigate the primary flux paths in the motor, to check the maximum flux density in the structure for saturation, and to evaluate motor parameters such as torque vs. rotor angle curve, back emf waveform, motor torque constant, and inductance value. The FEA-calculated flux paths are utilized in developing lumped parameter magnetic models of the motor that can be used as a part of design tool for future axial field permanent magnet motors.

All the FEA analysis in this chapter uses a quasi-static, nonlinear magnetic field analysis. A quasi-static magnetic field is a dynamic field in which the inertia effect and the eddy current effect can be neglected. For example, the magnetic field produced by changes in rotor position can be assumed to be a series of magneto-static fields, which are quasi-static magnetic fields. This is
based on the assumption that the magnetic fields produced by the permanent magnets and current sources are not affected by the other dynamic effects, such as eddy current, hysteresis, and the rise and fall time of the current. A series of quasi-static, nonlinear magnetic analyses is performed in this chapter to determine the back emf waveform and one-phase-on energized torque waveform as a function of rotor position. Inductance values are also calculated using different cases (staggered and unstaggered stator).

For validation of the FEA model, measured back emf waveform, and inductance values from a prototype motor are compared to model results. Detailed prototype motor measurement test setup, procedures, and results are presented.

The lumped parameter magnetic circuit model is utilized in this study because it relates the motor geometric and material characteristics to the motor output parameters through closed-form algebraic expressions. The impact of geometry, excitation, and material property changes on the overall system can be assessed, and new designs can be rapidly and accurately proposed without going through FEA modeling and the construction of costly prototypes.

Lumped parameter magnetic model development is not as simple as modeling electrical circuits. In electrical circuits, electric current remains principally in the intended conductive paths, and there is very little leakage. In magnetic circuits, however, magnetic flux is not principally confined to magnetic material, and there is a significant amount of flux leakage [25]. Accordingly, flux leakage paths for the magnetic flux have to be included to develop more accurate lumped parameter magnetic circuit models. In this study, FEA models and
resulting flux distribution plots are used to identify leakage flux paths so that accurate lumped parameter magnetic circuit models can be developed.

3.1.2 Force / Torque Calculation

In this subsection, the virtual work method and Maxwell's stress tensor method of force/torque calculation from FEA results are described, and a simple comparison of cogging (detent) force is made to assess the methods.

A. Virtual Work Method (Co-energy Method)

The virtual work method (co-energy method), one of the two techniques used to calculate force using the FEA results, is well known, and force calculation is simple to perform. The general expression for force $F$ calculation is [26]

$$F_x = \frac{\partial W_{co}}{\partial x}$$  \hspace{1cm} (3.1)

where $x$ is any direction, and $W_{co}$ is the work or magnetic co-energy associated with the force. Equation (3.1) is derived from the definition of work as force multiplied by distance, and is called the virtual work expression for force. An approximated virtual work expression is

$$F_x = \frac{W_{co} (x + \Delta x) - W_{co} (x)}{\Delta x}$$  \hspace{1cm} (3.2)

where the object on which the force is desired is displaced a small amount $\Delta x$ in the direction of the unknown force.

The virtual work method can be explained with the aid of Figure 3.1. Two B/H curves are shown for the entire FEA model. Curve 1 is for the first position, for example, rotor position one; curve 2 is for the rotor position two. For each
rotor position, an FEA model must be built and analyzed to obtain the total magnetic co-energy in Equation (3.2). For both FEA models, the excitation is kept the same, giving $H = H_M$, where $H_M$ is the permanent magnet coercive force in this example. The area between the two curves is the energy difference, much like the hysteresis power loss in magnetic materials. The energy difference is used to calculate the mechanical work using Equation (3.2).

![Figure 3.1 B/H Curves at Two Positions with Constant Excitation](image)

In this study, the displacement of the permanent magnet is made by changing the permanent magnet element materials in the FEA model. $W_{co}$ is calculated for both positions of the rotor, and the force is calculated using Equation (3.2). When motion is confined to rotation, as in most motors and generators, torque, rather than force, must be considered. The basic relationship between torque and force is that the torque is given by the product of a tangential force acting at a radius and, thus, has units of force times lever-arm length. All electromagnetic devices, such as motors and generators, are intended to produce
force or torque. Therefore, a very accurate prediction of force/torque is important in the design process.

B. Maxwell's Stress Tensor Method

The Maxwell's stress tensor method also enables the calculation of force in electromagnetic devices using FEA results. This subsection outlines the Maxwell's stress tensor method. In contrast to the virtual work method, which uses a volume integral to determine the stored magnetic co-energy, the Maxwell's stress tensor method can compute local stress at all elements of a bounding surface. Therefore, the sums of the local stresses by using a surface integral provide the net force. Because the derivation of the Maxwell's stress tensor is covered in most advanced magnetic field text books [27], only a brief derivation is quoted below.

Using Lorentz formulation, the magnetic force \( \mathbf{f} \) between a charge \( q \) moving with velocity \( \mathbf{v} \) in magnetic field \( \mathbf{B} \) is

\[
\mathbf{f} = q \mathbf{v} \times \mathbf{B}
\]  

The force density \( \mathbf{F} \) can be calculated from Equation (3.3)

\[
\mathbf{F} = \lim_{\delta V \to 0} \frac{1}{\delta V} \sum \mathbf{f}_i
\]

\[
= \lim_{\delta V \to 0} \frac{\sum q_i v_i \times \mathbf{B}_i}{\delta V}
\]  

\[
\mathbf{F} = \mathbf{J}_f \times \mathbf{B}
\]  

Equation (3.6) represents the magnetic force density \( \mathbf{F} \) on the current density \( \mathbf{J}_f \) interacting with magnetic field \( \mathbf{B} \).
Substitute $\nabla \times \mathbf{H} = \mathbf{J}$ and $\mathbf{B} = \mu \mathbf{H}$ into Equation (3.5),

$$\mathbf{F} = \mu (\nabla \times \mathbf{H}) \times \mathbf{H}$$  \hspace{1cm} (3.6)

By vector identity, Equation (3.6) becomes

$$\mathbf{F} = \mu (\mathbf{H} \cdot \nabla) \mathbf{H} - \frac{\mu}{2} \nabla (\mathbf{H} \cdot \mathbf{H})$$  \hspace{1cm} (3.7)

In concise tensor form, with some manipulation, the $i^{th}$ component of (3.7) is shown to be

$$F_i = (\mu \frac{\partial H_i}{\partial x_j} - \frac{\mu}{2} \frac{\partial}{\partial x_i} (H_k H_k))$$  \hspace{1cm} (3.8)

Using the property of the Kronecker delta

$$\left[ \frac{\partial}{\partial x_i} = \delta_{ij} \left( \frac{\partial}{\partial x_i} \right) \right]$$

and with some manipulation, this expression is written as

$$F_i = \frac{\partial}{\partial x_j} \left( \mu H_j H_i - \frac{\mu}{2} \delta_{ij} H_k H_k \right) - H_i \frac{\partial \mu H_j}{\partial x_j}$$  \hspace{1cm} (3.9)

The last term is

$$H_i(\nabla \cdot \mathbf{B}) = 0$$

Finally Equation (3.9) can be written as

$$F_i = \frac{\partial T_{ij}}{\partial x_j}$$  \hspace{1cm} (3.10)

where the Maxwell's stress tensor $T_{ij}$ is given by

$$T_{ij} = \mu H_j H_i - \frac{\mu}{2} \delta_{ij} H_k H_k$$  \hspace{1cm} (3.11)
where i, j and k are the three components of a vector in the three direction. The complete Maxwell's stress tensor $T$, written out in full, has the following form:

$$
T = \begin{bmatrix}
(B_i^2 - \frac{1}{2}|B|^2) & B_i B_j & B_i B_k \\
B_j B_i & (B_j^2 - \frac{1}{2}|B|^2) & B_j B_k \\
B_k B_i & B_k B_j & (B_k^2 - \frac{1}{2}|B|^2)
\end{bmatrix}
$$

(3.12)

The components of the stress tensor $T_{ij}$ can be calculated in the area of interest if the magnetic field intensity $H$ or magnetic flux density $B$ is known. With symmetry, at most six components are needed, $T_{ij} = T_{ji}$.

For an interface between two materials (such as air and stator iron in the motor), the resulting force, $F_i$ comes from the contributions of each magnetic force acting on the same surface with opposite normal vector ($T_{ij}^{\text{iron}}$ and $T_{ji}^{\text{iron}}$). The magnetic force can be simplified with the fact that $\mu_{\text{air}} << \mu_{\text{iron}}$, and the ratio of $\mu_{\text{air}}/\mu_{\text{iron}}$ is approximately 1/2500. Therefore $T_{ij}^{\text{iron}}$ can be neglected in calculation of magnetic force with an error boundary of 0.02% [26].

In this study, the magnetic force in the stator tooth face and tooth side along the air gap is calculated using the following equations [28].

$$
f_x = B_x B_y + \left(B_x^2 - \frac{1}{2}|B|^2\right)
$$

(3.13)

$$
f_y = B_y B_x + \left(B_y^2 - \frac{1}{2}|B|^2\right)
$$

(3.14)

In the 2-D FEA modeling, the magnetic vector potential, $A$, variation in the $z$-direction is assumed to equal zero; that is, $B_z = 0$. Thus, instead of the surface integration (3-by-3 matrix), which is required in the 3-D FEA, only a contour integration (2-by-2 matrix) is required in the 2-D FEA model.
Although the Maxwell stress tensor involves more calculations than the virtual work method, it enables determination of electromechanical boundary conditions in a concise form. Another advantage of Maxwell's stress method is that it solves problems without calculating the current density while using the field distribution only.

C. Force Calculation Method Comparison

In this comparative study, a sample cogging (detent) force is calculated using the virtual work method and Maxwell's stress tensor method as a function of 12 different rotor positions. The FEA model used in this calculation is the 2-D multisegment FEA model for the cogging torque analysis described in subsection 3.2.4. A detailed FEA model description is presented in the following sections. For each rotor position, a nonlinear, quasi-static field analysis is performed using FEA. The cogging forces are then calculated using the two different methods for different rotor positions. As shown in Figure 3.2, no significant differences are found between the cogging force calculation results of the two different methods.

The virtual work method requires total magnetic co-energy, which can be calculated using magnetic flux density \( \mathbf{B} \) in the \( x \)- and \( y \)-direction. At each element, magnetic flux density \( \mathbf{B} \) is calculated by the FEA solver. The total magnetic co-energy can be obtained using a summation of each element of magnetic field density \( \mathbf{B} \) in the FEA model. Most modern FEA packages provide macro programs to calculate the total magnetic co-energy to find force using Equation (3.2). The Maxwell's stress tensor method requires additional mathematical calculations with an external computer program. An advantage of using Maxwell's stress tensor method is that it requires one FEA model for one rotor position force calculation. (The virtual work method requires two FEA
models for one rotor position force calculation.) However, the developed FEA generic modeling capability program builds and analyzes FEA models with minimum interaction with the design engineer. The number of FEA models is not critical for the series of calculations where accurate results and minimum user-interface are the primary considerations. Even though the virtual work method requires two FEA models for one rotor position, its FEA generic modeling capability minimizes the total engineering time so that it approximates that time required if using the Maxwell's stress tensor method. After quantitative study of these two force calculation methods, the virtual work method was selected in this study because of its accuracy, limited number of calculations, and existing macro programs.

Figure 3.2 Force Calculation Method Comparison
3.2 FINITE ELEMENT ANALYSIS

In this section, motor design parameters and performance characteristics are analyzed by an FEA program that finds the magnetic vector potential \( A \) and magnetic flux density \( B \) throughout the 2- and 3-D region containing nonlinear magnetic materials. The FEA program requires that the region be divided into a number of finite elements having vertices called grids. The location of the grids, material property (permeability), boundary conditions, and excitation (current density or coercive force) of each element are required as input data. In addition to calculating \( A \) and \( B \) throughout the region and plotting the flux pattern, the program computes the flux linkage \( \lambda \), inductance \( L \), magnetic energy, and magnetic co-energy, \( W_M \) and \( W_{co} \) respectively, in the region of interest.

These calculated motor design parameters and performance characteristics are compared with lumped parameter magnetic circuit calculation and experimental measurements to verify the accuracy of the FEA models and to evaluate the proposed motor design.

3.2.1 FEA Model Description

It can be seen from Figures 2.12 and 2.13 that the motor has symmetry that allows it to be modeled by one-pole pair section of the full motor. Mechanically, one-pole pair section of the whole motor represents electrically 360 degrees; thus, one pole is 180 electrical degrees. In many cases the 180 electrical degree model is sufficient to simulate the full motor. Because the FEA package is limited in the total number of possible grids, modeling one-pole pair, or in some cases only a pole section of the motor, permits a finer finite element mesh. The outline of the 3-D FEA mesh and 2-D FEA mesh used for the motor model is illustrated in
Figure 3.3 and Figure 3.4. In the 2-D FEA model, the quad element type was chosen rather than the triangular element because of its accuracy.

The FEA model boundary conditions used throughout this study are the single-point constraints (SPC) and multipoint constraints (MPC) [28]. All the interior grid points in the model are unconstrained, while grid points on the exterior of the mesh are constrained in a manner dependent on the boundary conditions. In the model shown in Figure 3.5, the flux in the stator back iron exterior surface is assumed confined to the iron outer boundary. Flux lines along such a boundary (not crossing it) can be shown to enforce $A = 0$ along the boundary, which is called single point constraints [28]. SPCs are applied for such exterior grid points. Because the model has identical pole pairs, each pole pair boundary has periodic boundary conditions. For the model shown in Figure 3.5, periodic boundary conditions are expressed as

$$A(x_0, y) = A(x_0 + p, y)$$

(3.15)

where $x_0$ is the starting position of the one-pole section of the pole pair, and $p$ is the total distance of one-pole pair section. Periodic $A$ is enforced by exchanging a constrained grid with the controlling grid and adding the controlling grid terms to the finite element matrix equation. This periodic boundary condition is called multipoint constraint. Figure 3.6 depicts the symmetrical magnetic flux distribution in one of the FEA models. The periodic condition is applied to both right ($x=x_0 + p$) and left ($x=x_0$) sides of the model.
Figure 3.3 3-D FEA Model of One-Pole Section
Figure 3.4 2-D FEA Model of One-Pole Pair Section
Figure 3.5 Boundary Conditions Applied in FEA Model
Figure 3.6 Magnetic Flux Distribution in FEA Model
3.2.2 Generic FEA Modeling Capability Program

In this section, the generic FEA modeling capability program is discussed. The primary objective of the generic FEA modeling capability program is to mechanize FEA analysis by using a number of specialized FORTRAN programs (A through E in Figure 3.7), VAX command procedures (1 and 2), MACRO programs, and MSC/XL preprocessor and postprocessor input files. A few simple FORTRAN programs have been developed in conjunction with VAX/VMS command procedures and MACRO programs to (1) automate the building of 2-D and 3-D FEA models, (2) process the series of FEA model calculations with minimum user-interaction, and (3) provide necessary output data based on user requirement. Refer to Figure 3.7 for a generic FEA modeling capability program flow chart.

The generic FEA modeling capability program package works with both 2- and 3-D MSC/EMAS and MSC/XL preprocessor and postprocessor with developed FORTRAN programs to calculate back-emf waveform and torque vs. angle curves for both cogging and energized torque, and to determine motor parameters. This generic modeling capability program enables a designer to quickly assess changes in the axial field motor design such as number of poles, number of rotor poles, number of stator slots, dimensions of the motor, and excitations for each model.

The generic FEA modeling capability program can be used for modeling of the double-stator, single-rotor type axial field motor. To use this capability for the radial field motor configuration, several modifications of the FORTRAN programs A and B are required. Three FORTRAN programs (A through C) have
Figure 3.7 Generic Modeling Capability Flow Chart
been developed to automate mesh generation and analysis of the model using a simple interactive procedure. As can be seen in Figure 3.7, FORTRAN program A asks motor dimensions, while program B calculates FEA model geometry parameters that the MSC/XL preprocessor can process. Based on three different excitation requirements for an FEA model including permanent magnet only, permanent magnet and stator current, and stator current only, a different mesh density for the model is calculated using program C. Based on the user input data in FORTRAN program A, FORTRAN program C calculates rotor permanent magnet positions for each movement and steps with corresponding finite element and grid numbers. The method to simulate rotor permanent magnet movement is discussed in detail in subsection 3.2.7. FORTRAN program C does not create the FEA model, but it prepares all the necessary input data and information for the FEA preprocessor and postprocessor (MSC/XL) to create FEA models in matrix format (data files for FEA solver). The created FEA data files by MSC/XL are then submitted to the FEA solver program using VAX/VMS procedure 2 for a series of processes without user interaction. The FEA models are solved by MSC/EMAS program. The FORTRAN program D then calculates specific output files that are used for different output analyses such as back emf waveform, cogging torque, energized torque, and inductance. Using postprocessor program (MSC/XL) and FORTRAN program E, desired modeling results and plots can be obtained. A detailed discussion of FEA calculation steps are presented in subsections 3.2.7 through 3.2.11 and Chapter 4.

3.2.3 FEA Model Checking

The accuracy of the FEA field calculation results primarily depend on the resolution or density of meshing and the element type used [29]. In this
subsection, the 2-D FEA models with quad elements are checked for convergence to verify the accuracy of the model results. The purpose of conducting this test is to (1) find the optimum total number of elements with minimum CPU computing time and (2) identify the highest and lowest sensitive areas of mesh density in the model for future modeling. As shown in Figure 3.8, the model is divided into four different sections, i.e., stator back iron, slot and tooth, air gap, and permanent magnet areas. Figure 3.9 shows an example of subsection layers: the back iron has three layers in the y-direction, the slot and tooth has five layers in the y-direction, the air gap has two layers in the y-direction, the permanent magnet has three layers in the y-direction and six layers in the x-direction.

To determine the optimum number of layers in each subsection, an FEA run is made for a given number of layers (back iron - two layers, slot and tooth - five layers, air gap - two layers, and permanent magnet - two layers) in all four subsections, and co-energy is calculated with CPU time measurement. The number of layers in a particular subsection is then increased by adding more layers while holding the other subsection layers constant, thus increasing a subsection mesh density in the y-direction. Because six layers in x-direction are sufficient, only y-direction layers have been changed (see Figure 3.10, x-dir). Co-energy and CPU time are again calculated for increased mesh densities. The finite element aspect ratio in magnetic analysis is more relaxed than the structural analysis, and it allows a ratio of up to 10:1 instead of 4:1 in structural modeling. Nevertheless, the aspect ratio has been kept within 5:1 to 7:1 throughout out this study.

In Figure 3.10, the co-energy variation as a function of the number of layers in each subsection is presented. The number of each subsection layer is increased while the other subsection layers are held constant. The figure shows
Figure 3.8 Four Sections of Model for Convergence Test

Figure 3.9 Example Model for a Convergence Test
that the most sensitive areas in the model are the number of layers in permanent magnet and air gap. As a result, the FEA model construction efforts were focused on these two areas to obtain accurate FEA analysis. Based on the convergence test results, the optimum number of elements (layers) for each subsection were determined.

Figure 3.11 shows the co-energy and CPU time vs. total number of elements in the model and illustrates a clear trade-off between the total number of elements in a model and cost of computing time. As shown in Figure 3.11, for the total number of elements above 1400, the resulting accuracy does not
improve, but the CPU time goes up as a square function of the total number of elements.

The total number of elements for optimum results is 1724 for the given geometry and permanent magnet excitation. It can be concluded that any particular FEA model has its own optimum number and location of elements that depend upon geometry and type of excitation. Thus, a convergence test is necessary before processing a series of models. The time spent to check the accuracy of the model in the beginning of the process will produce more accurate final results.

![Figure 3.11 Co-energy and CPU Time vs. Total Elements](image-url)
As an example, two different cogging torque calculation results are compared in Figure 3.12. The solid line result was obtained using a total of 1724 elements in the FEA model, and the dotted line result was obtained using a total of 312 elements in the FEA model. Both models had the same geometry and dimensions. The first model (1724 elements) has a concentration of many more layers in the permanent magnet and air gap area than does the second model (312 elements). The result of comparing these two curves supports the contention that convergence testing is an essential step in FEA modeling and must be examined before conducting a series of calculations using FEA models.

![Figure 3.12 Cogging Torque Result Comparison](image-url)
3.2.4 Multisegment FEA Modeling Technique

Section 3.2.4 explains the 2-D multisegment FEA modeling technique and its functions. The rationale for selecting this modeling technique, instead of 3-D modeling, is also discussed. Finally, the specific six segment 2-D model developed in this study is described.

Motor symmetry typically simplifies the FEA model to a one-pole pair section, or even one-pole section in some cases. Because most FEA software packages are limited in the total number of grids, modeling motor symmetry conditions permits a finer mesh and reduces computational time and engineering costs. The outlines of the 2-D and 3-D FEA motor models used in this study are illustrated in Figure 3.13 and Figure 3.14. The objectives of the 3-D FEA model were (1) to verify the accuracy of the multisegment 2-D model result and (2) to compare the 3-D model developing time and CPU time requirements with 2-D modeling. Figure 3.15 shows the 3-D FEA air-gap flux density distributions at five different, equally spaced, radius locations. This result was used to verify the multisegment (six) model results in Section 4.2.2. However, the 2-D model, validated by the 3-D results, is easier to use and yields sufficiently accurate results for cogging torque, axial force calculation, and energized torque analysis. End-turn effects are not considered in this study because the stator diameter and the number of slots are large, while the end-turn length is relatively small.
Figure 3.13 2-D One-Pole Pair FEA Model
Figure 3.14 3-D One-Pole Pair Section Model
Figure 3.15 Magnetic Flux Distribution for Various Radii
There are three reasons why the 2-D multisegment modeling technique was chosen.

First, the time required for 3-D modeling, computing, and interpreting of the results is significantly greater than that required for 2-D modeling and analysis. For example, one FEA 3-D model with 7088 total hexa elements takes almost 3 hours of VAX CPU time for a single run — an expensive process, particularly because a complete back emf waveform or cogging torque calculation requires analysis of at least 30 different rotor positions with given motor dimensions. The 2-D model with 1724 quad elements requires only 3 minutes for a single run, or 18 minutes to run all six multisegment models. The CPU time charge for the VAX computer is approximately $100/hour, so it is important to consider not only the total number of elements and element type in the model but also the model type, such as a 2-D or 3-D model. In many cases, the 3-D physical geometry of a real system can be modeled using a simplified 2-D model with a few basic assumptions. This modeling provides acceptable results and is the primary reason why the new 2-D multisegment modeling technique is used in this study.

Second, because of the nature of the prototype motor geometry, the slot width of the designed axial motor is constant throughout the radial direction, and both tooth and permanent magnet width are a function of radius. The innermost radius has a slot/tooth ratio close to 1, but the outermost radius has a slot/tooth ratio of approximately 1/2 (refer to Figure 3.16). To properly model this effect, a multisegment 2-D modeling technique was developed by dividing the motor along its radial axis into radial sections. These sections have different radial thicknesses or 2-D model depth as shown in Figure 3.17.
Figure 3.16 Six-Segment Stator Top and Side View
The slots for the six-segment model, as seen from the tangential perspective (Figure 3.16 and Figure 3.17), are of equal width and are parallel for the entire radius of the stator. However, the tooth width, as seen from the tangential perspective, continually varies with changes in radial length. As the changes take place, the combined tooth and adjacent slot area must be divided into an integer number of meshes for the FEA model with each having an identical tangentially incremental width. This requirement for the rotor movement can be simulated using the material property changing technique discussed in subsection 3.2.7. For these reasons, each segment of the six-segment model must be of a different radial thickness.

Third, using the multisegment modeling technique to calculate the cogging torque allows the permanent magnet skewing effect to be analyzed using a multisegment model whereas the 3-D FEA model requires two separate sets (unskewed permanent magnet and skewed permanent magnet). Furthermore, in the 2-D multisegment model, because of the proper positioning of the permanent magnets, the skewing effect can be analyzed. (See subsection 4.3.1 for a detailed discussion on permanent magnet skewing.)

With this modeling technique, the arc shape of the prototype motor is linearized as a flat 2-D model segment. Because the size of the prototype motor radius is relatively large, the approximation is not significant. The staggered and unstaggered stator with different rotor positions are calculated.
Figure 3.17 Six-Segment Model
3.2.5 Permanent Magnet Modeling

The designed permanent magnet axial field motor consists of two stators and a rotor with permanent magnets embedded. To analyze this permanent magnet motor accurately with the FEA model, an accurate permanent magnet model must be developed. To predict back emf waveforms and cogging torque analysis an accurate permanent magnet modeling technique must be used.

There are two permanent magnet modeling techniques currently used. Figure 3.18 provides the starting point of the first permanent magnet modeling technique. This figure shows the normal B/H curve for a typical rare-earth permanent magnet material. It is assumed that the operating point is somewhere along the straight line from \( P_1 \) to \( P_2 \) shown in Figure 3.18.

![Demagnetization Curve of Permanent Magnet Material](image)

Figure 3.18 Demagnetization Curve of Permanent Magnet Material
The straight line can be expressed as \([16,17,27]\)

\[ B = \mu H + B_r \]  
\[(3.16)\]

Thus

\[ H = \frac{B - B_r}{\mu} \]  
\[(3.17)\]

Using Equation (3.17), Ampere's law becomes

\[ \nabla \times \frac{B - B_r}{\mu} = J \]  
\[(3.18)\]

\[ \nabla \times \frac{B}{\mu} = J + \nabla \times \frac{B_r}{\mu} \]

\[(3.19)\]

Thus, the permanent magnet can be modeled by a material of permeability \(\mu\) with a current density \(J_{PM}\)

\[ J_{PM} = \nabla \times \frac{B_r}{\mu} \]  
\[(3.20)\]

A rectangular-shaped permanent magnet as shown in Figure 3.19 is modeled as a material with uniform permeability \(\mu\) and with current density \(J_{PM}\). Because \(B_r\) is assumed uniform throughout the magnet, \(J_{PM}\) exists only on the magnet surface. For two dimensional Cartesian coordinates where

\[ B_r = B_r u_y \]  
\[(3.21)\]

\[ \nabla \times \frac{B_r}{\mu} = u_z \frac{\partial}{\partial x} \left( \frac{B_r}{\mu} \right) = J_{PM} \]  
\[(3.22)\]

where \(u_y\) and \(u_z\) are the unit vector in the \(y\)- and \(z\)-directions respectively.

\(B_r\) is a step function equal to zero outside the permanent magnet and equal to \(B_{rPM}\) inside the permanent magnet. Thus Equation (3.22) is rewritten as
\[ J_{PM} = \frac{\Delta B_Y}{(\mu \Delta x)} = \frac{B_{rPM}}{\mu \Delta x} \quad (3.23) \]

The related current is

\[ I = J_{PM} l_{PM} \Delta x = \frac{B_{rPM} l_{PM}}{\mu} \quad (3.24) \]

where \( l_{PM} \) is permanent magnet length in the \( y \)-direction and \( \Delta x \) is the Kronecker delta. This sheet current is along the two sides of the permanent magnet model as shown in Figure 3.19.

![Figure 3.19 Developed Model of Permanent Magnet](image-url)
In Figure 3.18, if a material represented by a curved line from P1 to P2 was to be used, modeling would be possible if H is expressed as an analytic function of B. Such an equation would enable a solution similar to that for stator lamination material [30].

In the second permanent magnet modeling technique, the permanent magnet area is discretized into many elements and magnetic dipole moment (magnetization) with permeability is applied to every element (see Figure 3.20). Because the permanent magnet material develops magnetic dipole moments in response to external magnetic fields [31], the dipole moment per unit volume is called the magnetization M. For convenience, the magnetic field strength H is defined in terms of B and M using Equation (3.16)

\[ H = \frac{1}{\mu} B - M \]  

(3.25)

where M is the magnetization, B is magnetic flux density, and H is magnetic field intensity. The units of H and M are the same as a sheet current, in MKS units, A/m. When the external field is removed, the coercive force H_c at B = 0, which is numerically equal to the magnetization. Thus, the permanent magnet can be modeled using the H_c value and permeability \( \mu \). Figure 3.20 shows the permanent magnet area with applied Hc value and permeability \( \mu \).

To validate model accuracy of the two permanent magnet modeling techniques discussed above, model results were compared with experimental measurements. In the first method, a uniform sheet current and permanent magnet material relative permeability \( \mu_{PM} \) were applied to the two edges of FEA permanent magnet mesh as shown in Figure 3.19. In the second model, the value of H_c, the coercive force (current density), and the relative permeability \( \mu_{PM} \) were applied to all elements in the model permanent magnet area.
Figure 3.20  Permanent Magnet Model with Coercive Force Applied
3.2.6 Experimental Validation of Permanent Magnet Models

To measure the permanent magnet strength accurately, a rigid test fixture was constructed containing an x-y-z micrometer positioning device with an aluminum carriage and stage as shown in Figure 3.21. An axial gauss meter probe was attached to the aluminum bar carriage. The entire test fixture was made of nonmagnetic material to avoid possible magnetic flux induction by the device. In this test fixture, an air-gap length of 0.127 mm (5/1000 in.) was set between the permanent magnet and gauss meter probe. The magnetic strength was measured for every 1.0 mm in the x- and y-directions over the entire permanent magnet (Figure 3.21). Figure 3.22 depicts approximate shape and dimensions of the permanent magnet used in this test. Note that the dimensions used in the permanent magnet tested are not the same as the dimensions of the permanent magnets used in the prototype motor. Figure 3.23 shows the permanent magnet material properties for the tested permanent magnet.

![Figure 3.21 Magnet Strength Measurement Test Setup](image-url)
Figure 3.22 Shape and Dimensions of Permanent Magnet Measured

![Diagram of a permanent magnet with dimensions and views.](image)

Figure 3.23 Tested Permanent Magnet Material Properties

![Graph showing demagnetization curve and energy product values.](image)
Figure 3.24, a 3-D plot of the permanent magnet strength measurement, shows the approximate shape of the measured permanent magnet. There are four high raised field in the edge of permanent magnets.

Figure 3.25 shows the results from two FEA models with two different modeling techniques and experimentally measured values from the middle of the permanent magnet (dotted line in Figure 3.22). The FEA results show that the sheet currents method and the coercive force (current density) method are identical except at the edge of permanent magnets. There is a close correlation between the experimentally measured value and the coercive force method. The sheet currents method reveals a spike at edges of the permanent magnet. This spike is crucial when permanent magnet position-dependent analysis is performed, such as cogging torque and back emf calculations.

Previous cogging torque calculations performed by other authors were done using the sheet currents method for the permanent magnet model, and the result contained significant errors [25]. In this study, the permanent magnets were modeled using the coercive force method, and the results obtained were very close to the experimentally measured values.

In general, the sheet currents method and the coercive force method show a close correlation except with the cogging torque calculation, where the permanent magnet modeling technique and mesh density are very sensitive to output results.

If the Lorentz force equation is used to calculate torque, only the sheet currents method can be used to model permanent magnets. Many technical papers show reasonable correlation between calculated and measured values by
using sheet currents method. In this study, the coercive force method, an easier and more accurate than the sheet current method, was extensively used to model the permanent magnets. The coercive force method requires more CPU time than does the sheet currents method because the coercive force method applies magnetic dipole moment or coercive force $H_c$ in every element in the permanent magnet area with material property $\mu$. In the sheet currents method, the equivalent current density is applied for only the two side edges of the permanent magnet area. While the additional CPU time may be viewed as a drawback, the objective of the modeling analysis in this study was to obtain accurate cogging torque results, rather than a low-cost analysis.

![3-D Plot of Permanent Magnet Strength](image)

**Figure 3.24** Permanent Magnet Strength Measured Value in 3-D Plot
3.2.7 Air-Gap Flux Distribution

Accurate prediction of the magnetic flux distribution in a motor model is critical information for motor design purposes. Flux distribution, for example, not only gives the degree of magnetic saturation throughout the various areas of the designed motor, but it also gives the ratio of leakage to useful flux in its magnetic paths. This information is used to find optimum stator back-iron thickness as a function of the different residual flux density of permanent magnet materials. The air-gap length has also been approximated to the point where the leakage is minimum. A pictorial display of the flux distribution can be obtained from the FEA preprocessor and postprocessor package by plotting the specified contours of equal magnetic vector potential within each element. This pictorial display of a flux distribution represents a useful tool for detection of errors in the implementation and use of the FEA program in field analysis, such as faulty
assignment of boundary conditions, material properties, excitations, and inaccurate implementation of the rotor rotation simulations.

The air-gap flux distribution, as a function of rotor position in electric motors, provides key information for the calculation of induced back emf waveforms, cogging torque, energized torque waveforms, inductance, and other motor performance characteristics. Figure 3.26 shows a 2-D model of an FEA result for a rotor and stator with permanent magnet excitation only. Table 3.1 lists the material properties used in this model, and Figure 3.27 shows B/H data for the stator steel used in the FEA model. Figure 3.27 shows the flux density waveforms (y-direction) at the middle of the air-gap (radius=17 cm, and air-gap length = 2.0 mm) over the one-pole section of the motor model. The effect of slot openings is obvious in this waveform. The reduction in the y-direction flux density value opposite the slot openings is attributed to the higher y-directional reluctance of the flux path at these locations. The air-gap flux density of different rotor positions is calculated using a series of "snapshot" FEA field solutions. A complete electrical cycle of the rotor permanent magnet rotation can be obtained using snapshot analysis.

Table 3.1 Material Properties and Permanent Magnet Excitation

<table>
<thead>
<tr>
<th>Material</th>
<th>Permeability (μ)</th>
<th>Excitation (coercive force)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permanent magnet (Nd B Fe)</td>
<td>1.05</td>
<td>800,000 (A/m)</td>
</tr>
<tr>
<td>Lamination steel (M19)</td>
<td>3,000</td>
<td>0.0</td>
</tr>
<tr>
<td>Titanium</td>
<td>1.0</td>
<td>0.0</td>
</tr>
<tr>
<td>Air</td>
<td>1.0</td>
<td>0.0</td>
</tr>
</tbody>
</table>
Figure 3.26 2-D FEA Model of One-Pole Section
Figure 3.27 B/H Curve for Stator Lamination Material
Figure 3.28  Air-Gap Flux Density Distribution
3.2.8 Back Emf Calculation

This subsection presents the results of the back emf waveform calculation using an FEA model. The back emf waveform is an important motor characteristic that must be considered for design and analysis of electric motors. This waveform can be used to predict the output torque waveform, and subsequently the output power can be calculated using back emf waveforms. A series of snapshot FEA solutions provide accurate air-gap flux density of different rotor positions. Therefore, the magnetic-flux linkage, as a function of rotor position, can be obtained.

The back emf waveform is computed by determining the change in flux linkage in the stator phase coils versus rotor position. The voltage caused by a changing permanent magnet flux linking a phase coil of turns is found through Faraday’s law as

$$E = \frac{d\lambda}{dt} = N \frac{d\phi}{dt} = N \frac{\partial\Phi}{\partial\theta} \omega$$

(3.26)

where $E$ is induced voltage, $\lambda$ is flux linkage in a phase coil, $N$ is number of turns in one phase coil, $\theta$ is rotor position, and $\omega$ is the rotor angular velocity. This implies that two FEA solutions are required to find $E$, and the solutions must correspond to two different positions of the rotor. The most efficient method to compute the flux throughout the cross section of any particular coil is to use the change in the magnetic vector potential (MVP) from one side of the stator tooth to the other, as derived below.

The definition of magnetic flux [31] is
\[ \phi = \int \mathbf{B} \cdot dS \]  
(3.27)

and by using the definition of vector potential

\[ \mathbf{B} = \nabla \times \mathbf{A} \]  
(3.28)

and Stokes’ identity, the surface integral can be replaced by a closed-line integral around the surface,

\[ \phi = \int \nabla \times \mathbf{A} \cdot dS = \oint \mathbf{A} \cdot dl \]  
(3.29)

For 2-D problems the magnetic flux between two grids is simply

\[ \phi = (A_1 - A_2) d \]  
(3.30)

where,

\( d \) = depth of the 2-D model

\( A_1 \) = MVP evaluated at the left side of the tooth

\( A_2 \) = MVP evaluated at the right side of the tooth.

By the geometrical definition, \( \theta \) (Left) > \( \theta \) (Right), where \( \theta \) is zero at the \( +x \) axis and is measured counterclockwise. This is based on application of the Stokes’ theorem in conjunction with the definition of the flux density (\( \mathbf{B} \)) as the curl of the MVP (\( \mathbf{A} \)).

Throughout this study, the back emf waveforms were obtained from the analysis of 30 different snapshot FEA runs, each at a different rotor position. For each position, the flux cutting through a phase winding is determined by using the vector potential calculation of Equation (3.30). The rotor was incremented a small angle, 0.5 mechanical degrees, for each flux calculation. All 30 FEA models
for 30 different rotor positions are created and executed as a single batch job using the generic FEA modeling capability described earlier in this chapter for the determination of the back emf as a function of the rotor position.

There are two methods to simulate the rotor movement: relocation of the permanent magnet elements by physical displacement in the FEA model, and changing the material property of the permanent magnet elements. In the first method (the most common) after the permanent magnet section of the model has been relocated to the desired position, then the moved grids must be matched with all the other grids, and the boundary conditions must be reassigned. When only a few rotor movements are required, this method can be used with user interaction. This relocation method is not useful when the rotor must move many steps with small displacements.

The second method to simulate rotor movement involves changing the material property of the permanent magnet elements based on rotor movement. This method is incorporated into the FEA generic modeling capability and is extensively used in this study. In this second method, the rotor movement is simulated by changing material properties on the leading edge ($\Delta x_1$) and trailing edge ($\Delta x_2$), as seen in Figure 3.29. This simulation eliminates relocating the whole rotor portion of the model, which would require additional steps, such as changing boundary conditions and equivalencing the grid points on either side of the rotor with the air-gap grid points. Using the material property changing method, if the rotor moves in the x-direction with $\Delta x = 2$ mm, then the material property of the permanent magnet can be extended 2 mm in positive x-direction, i.e., the 2 mm portion of the rotor core material ($\Delta x_1$) is changed to the permanent magnet material, and the permanent magnet material portion in the trailing edge ($\Delta x_2$) is changed to the rotor core material (see Figure 3.29).
The material property changing method requires that the mesh size in the x-direction be uniform to simulate a uniform rotor movement. The user, therefore, must know the step size of the rotor movement before simulation starts. Step size depends on how many data points are required to complete the back emf waveform. Finally, by applying Equation (3.26), the back emf waveform is calculated as a function of theta, x-direction (see Figure 3.29). The FEA calculated back emf waveform is compared with experimental measurements.

3.2.9 Experimental Validation of Back Emf Calculation

To validate the FEA back emf calculation, experimental measurement was performed using test setup shown in Figure 3.30. In this test setup, the prototype axial motor (right) is back driven by a brush dc radial motor (left). All three phase-winding terminals are connected to the multichannel oscilloscope.
Figure 3.30 Back Emf Waveform Measurement Test Setup Picture
(Tektronix TDS 544A Digital Oscilloscope), which is not shown in the figure. Back emf waveforms are measured at different motor speeds.

Figure 3.31 shows the comparison between measured back emf and FEA calculated back emf. The time on x-axis (speed or frequency) is not shown because classified information may be derived from this data. The measurement curve is for phase A on stator 1. For the experimental test, the rotor was back driven at a very slow constant speed at which there is no eddy current effect. Therefore, a quasi-static, nonlinear magnetic field analysis can be used. Because the speed is the same speed used in the simulations, a direct comparison in magnitude can be made. Comparison of the FEA calculation and measurement curves indicates that the measured waveforms have a flatter peak and do not precisely match the sinusoid of the FEA models. However, the measurement curve approximates a sine wave with the frequency and magnitude close to those of the FEA model results, thus validating the FEA models as accurate analysis tools that can provide a method for evaluating motor designs.

As a result of this study it was found that there are a number of concerns that must be addressed in computing the back EMF waveform using FEA models including the:

1. Effect of the FEA model mesh density,
2. Effect of step size in x-direction,
3. Use of the nonlinear relative permeability, and
4. Effect of eddy currents in the stator lamination.

Based on the first three concerns, the FEA models were built using the optimum number of elements based on the convergence test result shown in Figure 3.11, a rotor movement step size of 0.5 mechanical degree, and a nonlinear
B/H curve data to the stator material. Effects of eddy currents in the stator and rotor were not considered in the 2-D FEA models. The result shows a clear correlation between the FEA-calculated and experimental measurements with a 10 percent error in magnitude. Mainly, the error (10 %) was contributed by the end-turn winding effect, and end turns are not incorporated in the 2-D FEA models. Since the motor torque constant can be inferred from the measured back emf waveform as a function of different motor speed in normal conditions in most permanent magnet motors, the designed motor output torque was calculated in the early design stage using the back emf waveform obtained in this chapter. Comparing the calculated torque constant with the prototype measured value showed that a close correlation exist.

![Back Emf Waveforms from FEA and Measurement](image)

**Figure 3.31** Back Emf Waveforms from FEA and Measurement
3.2.10 Inductance Calculation

In this subsection, the phase inductance value calculations using FEA results are presented. Accurate prediction of the phase inductance value at the design stage of the motor is critical because this value is directly applied to PWM circuit design. Many technical papers show unreasonably large errors in phase inductance value calculations when the inductance value is calculated by 2-D FEA models [14,15,16,25]. In this study, the phase inductance values for all different cases were calculated using multisegment 2-D FEA models, and the results approximate experimental measurement values.

The inductance value calculation using FEA can be obtained in two ways. First, using the relationship between the total co-energy within the FEA model and applied constant current excitation,

\[ L = 2 \frac{W_{co}}{I^2} \]  

(3.31)

where \( W_{co} \) is total magnetic co-energy in the FEA model, \( I \) is applied phase current. Second, using Equation (3.26) and (3.30) from subsection 3.2.8, the inductance is

\[ L = \frac{\lambda}{I} = \frac{N \phi}{I} \]  

(3.32)

where \( \lambda \) is flux linkage, \( N \) is effective number of turns in coil, and \( I \) is applied phase current.

As is discussed in subsection 3.1.2, the total magnetic co-energy calculation is easy to obtain; therefore, Equation (3.31) is used by a FORTRAN program within the generic modeling program to calculate inductance of each six-segment FEA model. Figure 3.32 and Figure 3.33 show one of the six-
Figure 3.32 One-Phase-On FEA Flux Plot for Staggered Stator
Figure 3.33 One-Phase-On FEA Flux Plot for Unstaggered Stator
segment 2-D FEA models with flux contour plots of the staggered stator and the unstaggered stator.

The purpose of staggering the stator is mainly to reduce the cogging torque (see Chapter 4 for a detailed discussion and calculation of this subject). In the following subsection, the FEA-calculated phase inductance value results are compared with experimental measurements, and the results are discussed. Since the rotor structure is made out of titanium, whose permeability is approximately 1.0, the same as air, (see Table 3.1), the rotor structure does not affect flux paths at all.

3.2.11 Experimental Validation of Inductance Calculation

Calculated inductance values using FEA models and prototype-measured phase inductance values are summarized in Table 3.2. The experimental measurement results shown in Table 3.2 were obtained using unstaggered and staggered stators. The inductance values for all three stator windings (phase A, B, C) were measured using a digital LCR meter (Hewlett Packard, 4263A) with a frequency range of 10Hz to 100 kHz and an input voltage signal at 1 volt. The error between calculated and measured result is within 10 percent, which is expected because the end-turn winding effect was not incorporated in multisegment 2-D FEA models. To find staggering effect on the inductance value, two FEA models (staggered and unstaggered stator) results were compared. The comparison revealed that the stator staggering does not affect the inductance value as shown in Table 3.2. It is apparent by examining Figures 3.32 and 3.33 that only a small portion of the total flux produced by stator windings goes through the opposite stator and the majority of flux flows as slot leakage flux. Therefore, the major inductance of the phase winding is attributed to
leakage flux paths in the stator slots, rather than flux paths that cross the air gaps and return through the other stator. The stator slot/tooth profile, shown in Figure 2.12, is such that the slots are relatively deep and have a large effective magnetic air gap between the two stators. The effective magnetic air gap can be defined as the distance between stator 1 surface and stator 2 surface. The deep slots provide a parallel flux path in the x-direction for the stator magneto motive force (MMF), which results in an increase in the slot leakage reactance of the motor. FEA prediction of the inductance values and the measured values is within 10 percent. The small error rate may be attributed to two factors: (1) the multisegment 2-D FEA modeling technique was used and (2) the ratio between radial length (depth in 2-D FEA model) and end turn length (x-direction in 2-D FEA model) is relatively large in the designed prototype motor geometry. Thus, the end turn length is relatively shorter compared to the radial winding length.

### Table 3.2 Comparison of Inductance Values

<table>
<thead>
<tr>
<th></th>
<th>Unstaggered (µH)</th>
<th>Staggered (µH)</th>
<th>Difference (µH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>FEA result</td>
<td>184</td>
<td>188</td>
<td>4</td>
</tr>
<tr>
<td>Prototype measured</td>
<td>203</td>
<td>213</td>
<td>10</td>
</tr>
<tr>
<td>Error (%)</td>
<td>9.4</td>
<td>12</td>
<td></td>
</tr>
</tbody>
</table>

### 3.2.12 Energized Torque Calculation

In this subsection, the one-phase-on energized torque vs. rotor angle is calculated using FEA models. The purpose of this calculation is to predict maximum output torque and its waveform using simple 2-D FEA models for
evaluation of preliminary designs and to compare the torque angle curve with the calculated back emf waveform.

Three different methods to calculate the energized torque are presented in this subsection. First, the torque is calculated using the virtual work method (co-energy method), which uses the relationship between the electromagnetic torque and the co-energy function

$$T(i,\theta) = \frac{\partial W_c(i,\theta)}{\partial \theta}$$  \hspace{1cm} (3.33)

The co-energy function $W_c$ is found in the FEA through the following expression [2,3,28]

$$W_c = \iiint \left[ \mathbf{B} \cdot d\mathbf{H} \right] dv = \iiint B_z^2 \frac{2}{\mu} dv$$  \hspace{1cm} (3.34)

where $B = \sqrt{B_x^2 + B_y^2}$ and $\mu = \mu_0 \mu_r$.

Secondly, the energized torque can also be determined from the following equation using the back emf constant $K_E$, which was previously obtained from FEA results.

$$T(i,\theta) = K_T i \sin (A\theta + (k - 1) \frac{2\pi}{3})$$  \hspace{1cm} (3.35)

where $K_T$ is the torque constant. Note that in international system (SI) of units $K_T$ is equal to $K_E$.

And finally, the $d\mathbf{F} = \mathbf{I}d\mathbf{L} \times \mathbf{B}$ or Lorentz force method can be utilized to calculate energized torque. This requires knowledge of the $\mathbf{B}$ field at the various current source locations. In this method, the $\mathbf{B}$ field from stator excitation and $\mathbf{I}d\mathbf{L}$ from the rotor permanent magnets are used.
The virtual work method was incorporated in the generic FEA modeling capability package to obtain the results from a section of the multisegment FEA model. For the un staggered stator configuration the torque angle curve is shown in Figure 3.34. The torque angle curve shown in Figure 3.34 includes stator energized torque and cogging torque. The calculation method used is the virtual work method, which calculates total magnetic co-energy.

![Figure 3.34 Energized Torque Result from FEA Models](image)

It is well known that the shape and frequency of the back emf waveform are almost identical to the torque vs. angle curve in most brushless permanent magnet motors [2, 18]. Based on this fact, the one-phase-on motor torque vs. angle calculation from the FEA models are validated. The torque vs. angle curve (Figure 3.34) approximates the FEA-generated back emf waveform (Figure 3.31).
As a result, the maximum torque value was estimated using these models, rather than using the actual prototype motor torque vs. angle measurements. A torque transducer to measure torque of such magnitude (1500 Nm to 2000 Nm) was not available to the author. Because of the embedded cogging torque and only 12 discrete points (12 rotor positions with 24 FEA models) were calculated, apparent discontinuities appear in Figure 3.34. The results were normalized to prevent disclosure of classified information. It is anticipated that the FEA modeling and comparison of the back emf waveform vs. the torque-angle curve can be applied to more detailed models to validate pre-prototype designs.

3.3 LUMPED MAGNETIC CIRCUIT ANALYSIS

3.3.1 Model Development

This subsection presents the development of the simple, lumped parameter magnetic models used for the designed prototype motor. The purpose of these models was to predict the back emf waveform, phase inductance value, and one-phase-on energized torque vs. rotor angle curve using closed-form algebraic equations.

The lumped model analysis is based on linear magnetic material property and assumes no saturation in the magnetic materials and that reluctance associated with the flux paths is not a function of rotor position. Because the motor phases can be energized differently in various drive configurations, the lumped model was developed so that any or all of the three phases would be energized at one time. Two models were developed, one for unstaggered and another for staggered stator configurations. Figure 3.35 and Figure 3.36 show a side view of the two configurations. Staggering the stator reduces the cogging torque (see Chapter 4 for specific details).
Figure 3.35 Side View of Unstaggered Stator Model

Figure 3.36 Side View of a Staggered Stator Model
To formulate a lumped magnetic circuit model, the magnetic flux paths from an energized phase FEA solution of an unstaggered and staggered stator model were examined as shown in Figure 3.37 and Figure 3.38. The FEA result shows that the primary flux paths exist directly under each tooth pitch, mainly in the axial direction (y-direction in 2-D models) and that leakage occurs in the theta direction (x-direction in 2-D models) attributed to the fringing effect near the tooth tips. Based on this information, two lumped parameter circuit models, unstaggered and staggered stator, were developed having 6 (unstaggered stator model) and 12 (staggered stator model) parallel branches containing lumped parameter elements representing energy sources and drops. Figure 3.39 and Figure 3.40 depict these lumped parameter models, unstaggered stator and staggered stator, respectively. These models represent one-pole pair of the entire motor, which, because of symmetry, is adequate to predict overall motor behavior.

It is necessary to explain the lumped magnetic circuit model simplified from the staggered stator model shown in Figure 3.40 into the Figure 3.41 model. In order to simplify the model in Figure 3.40, the stator MMF sources in the both stators are divided by half, and the remaining circuit elements are unchanged. For example, $M_a$ is the MMF source from phase A current excitation, which is divided in half to support $R_R/2$ and $F_1$. The following subsection 3.3.1A describes all electrical circuit equivalent lumped parameter magnetic circuit elements based on the two lumped magnetic circuits shown in Figure 3.39 for the unstaggered stator and Figure 3.41 for the simplified staggered stator.
Figure 3.37 Unstaggered Stator FEA Flux Plot
Figure 3.38  Staggered Stator FEA Flux Plot
A. Magneto Motive Force Source Calculation

The stator has rectangular teeth (see Figure 2.12 and Figure 2.13) with a cross sectional area that decreases toward the rotational axis of the motor. The teeth are rectangular to minimize saturation and to provide a uniform slot area for the possible rectangular wire used to maximize the packing factor in the phases. Each stator phase winding encircles three teeth and provides an MMF of \( Ni/2 \) where \( N \) is the total number of turns in phase coil and \( i \) is a phase current. The unstaggered stator has the teeth aligned so that one magnet covers three teeth of both stators (Figure 3.35). The staggered stator has the teeth offset so that two teeth are covered by the permanent magnet on one stator and three teeth are covered by the permanent magnet on the other side (see Figure 3.36).

In Figures 3.39 and 3.41, three energy sources exist in each of the circuit branches. Two represent the amp-turns developed in the energized phase winding of each stator, and the other energy source represents the permanent magnets. The MMF magnitude of one phase is denoted by \( M' \)'s. The MMF magnitude is equal to half the number of turns per phase multiplied by the phase current.

\[
\text{MMF}_i = \frac{1}{2} N_i i_i , \quad M_a = \sum_1^3 \text{MMF}_i
\]  

(3.36)

where \( i \) represents the phase number. Because the motor has three phases, each coil of a phase spans three teeth, and each branch of the lumped model "sees" 3 \( Ni \) sources. The MMF source is considered constant in each branch.
Figure 3.39 Unstaggered Stator, 6 Branches, Lumped Magnetic Model
Figure 3.40 Staggered Stator, 12 Branches, Lumped Model
Figure 3.41 Simplified, 12 Branches, Staggered Stator Lumped Model
The rotor permanent magnet MMF is assumed to be sinusoidally distributed in the air gap. Consequently, the magnet source representing the flux-creating capacity of the permanent magnet can be written as a function of rotor position and location of the model branch with relation to zero rotor angle

$$\text{MMF} = \frac{3.0 \, B_r \, l_m}{\pi \, \mu_0 \, \mu_r} \sin (B \theta + \alpha) \quad (3.37)$$

where

- \( l_m \) = length of the magnet,
- \( B_r \) = magnet residual flux density,
- \( B \) = number of magnet pole pairs,
- \( \theta \) = rotor position in mechanical degrees, and
- \( \alpha \) = phase shift equal to tooth pitch in electrical degrees.

The magnitude of the sine term is found by averaging the MMF distribution over the tooth pitch of 60 electrical degrees.

B. \hspace{1cm} \textbf{Reluctance Calculation}

The reluctance, \( R \), calculation is one of the most difficult in a magnetic analysis because of the complexity of the geometric shapes and magnetic field fringing effect. However, careful consideration of the calculation method can provide reasonable results. The general form of reluctance is given in the following equation

$$R = \frac{\text{MMF}}{\phi} \quad (3.38)$$

The reluctance values to be derived for this model were based on the field patterns given in Figure 3.42. The field lines were assumed to consist of straight
line segments and circular arcs, similar in shape to the flux lines in Figure 3.37 and Figure 3.38.

Using the reluctance formula

\[ R = \frac{1}{\mu_0 \mu_r} \int \frac{I}{dA} \]  \hspace{1cm} (3.39)

where:

\( dA = \) differential cross section of the flux tube,
\( I = \) length of the flux tube.

By using Equation (3.39), the unstaggered and staggered stator reluctances for the paths shown in Figure 3.42 [32] can be derived.

Unstaggered:

\[ R_g = \frac{l_g}{\mu_0 t_w T} \]  \hspace{1cm} (3.40)

\[ R_{ga} = \frac{\pi}{\mu_0} \ln \left(1 + \frac{\pi s_w}{2 l_g}\right) T \]  \hspace{1cm} (3.41)

\[ R_T = \frac{R_g R_{ga}}{2 R_g + R_{ga}} \]  \hspace{1cm} (3.42)

Staggered:

\[ R_1 = \frac{\pi}{2 \mu_0} \ln \left(\frac{l_g + 0.25 \pi s_w}{l_g + 0.5 \pi s_w - \frac{t_w s_w}{2}}\right) T \]  \hspace{1cm} (3.43)

\[ R_2 = \frac{l_g}{\mu_0} \left(\frac{t_w s_w}{2}\right) T \]  \hspace{1cm} (3.44)

\[ R_1 = R_3 \]  \hspace{1cm} (3.45)

\[ R_r = \frac{R_1 R_2}{2R_2 + R_1} \]  \hspace{1cm} (3.46)
Figure 3.42 Reluctance Models
where:

\[ OD = \text{outer diameter}, \]
\[ ID = \text{inner diameter}, \]
\[ T = (OD - ID) / 2, \]
\[ l_g = \text{gap between stators}, \]
\[ t_w = \text{tooth width}, \]
\[ s_w = \text{slot width}, \]
\[ R_t = \text{unstaggered stator total reluctance}, \]
\[ R_R = \text{staggered stator total reluctance}. \]

The complete derivation of the reluctances can be found in reference [32].

After determining the magnetic circuit parameters shown in Figure 3.39 and Figure 3.41, the next step was to determine flux linkage, back emf waveform, inductance value, and energized torque.

C. Lumped Model Branch Flux Calculation

Typical values for \( F_s \) and \( M_s \) in the lumped parameter magnetic model shown in Figure 3.39 and 3.40 are given below

\[ F_a = F_o \sin (30^\circ - \alpha) \]  \hspace{1cm} (3.47)
\[ F_b = F_o \sin (90^\circ - \alpha) \]  \hspace{1cm} (3.48)
\[ F_c = F_o \sin (150^\circ - \alpha) \]  \hspace{1cm} (3.49)
\[ F_1 = F_o \sin (15^\circ - \alpha) \]  \hspace{1cm} (3.50)
\[ F_2 = F_o \sin (45^\circ - \alpha) \]  \hspace{1cm} (3.51)
\[ F_3 = F_o \sin (75^\circ - \alpha) \]  \hspace{1cm} (3.52)
where

\[ F_0 = 3.0 \frac{B_r}{\mu_0} \frac{l_m}{\pi} \frac{\mu_r}{\mu_0} \]  \hspace{1cm} (3.53)

for unstaggered stator and

\[ F_0 = 3.105 \frac{B_r}{\mu_0} \frac{l_m}{\pi} \frac{\mu_r}{\mu_0} \]  \hspace{1cm} (3.54)

for staggered stator,

where

\[ B_r = \text{residual flux density of permanent magnet}, \]
\[ l_m = \text{permanent magnet length}. \]

\[ M_a = \frac{1}{2} (N_1 i_1 - N_2 i_2 - N_3 i_3) \]  \hspace{1cm} (3.55)

\[ M_b = \frac{1}{2} (N_1 i_1 + N_2 i_2 - N_3 i_3) \]  \hspace{1cm} (3.56)

\[ M_c = \frac{1}{2} (N_1 i_1 + N_2 i_2 + N_3 i_3) \]  \hspace{1cm} (3.57)

Flux is calculated in each of the branches by using Kirchoff's voltage law, treating magnetic energy sources as voltage sources, reluctances as resistance, and magnetic flux as current. Generalized unstaggered stator branch flux, \( \phi_n \), can be written in the following form

\[ \phi_n = \frac{\sum N_i i_i + F_0 \sin \left( \frac{\mu_i}{6} (2n - 1) - \alpha \right)}{R_t} \]  \hspace{1cm} (3.58)

where

\[ i = \text{phase number}, \]
\[ n = \text{branch of the model representing the tooth pitch being analyzed}, \]
\[ R_t = \text{total reluctance of each branch}. \]  The flux linkage can be written as

\[ \lambda_i = N_i \sum \phi_n \]  \hspace{1cm} (3.59)
where \( i \) is the phase number, and the flux cutting coil \( i \) is summed. From the flux linkage calculations, the self inductance, the mutual inductances, and back emf for each phase can be found. The calculated inductance value can be written in terms of the flux linkages. To compare lumped parameter results with the FEA and experimental results, only the one-phase-on one stator must be considered. This simplifies the flux calculations and results for the unstaggered stator in the following expressions

\[
L_i = \frac{\partial \lambda_i}{\partial i_i} = \frac{6 N_i^2}{R_t} \tag{3.60}
\]

\[
M_{im} = \frac{\partial \lambda_i}{\partial i_m} = \frac{2 N_i N_m}{R_t} \tag{3.61}
\]

\[
E_i = \frac{\partial \lambda_i}{\partial \theta} \tag{3.62}
\]

\[
E_i = K_{Ei} \sin (B \theta + (i - 1) \frac{2\pi}{3}) \tag{3.63}
\]

\[
K_{Ei} = \frac{4 AB N_i}{R_t} \tag{3.64}
\]

\[
A = \frac{3 B r l_m}{\pi \mu_0 \mu_r} \tag{3.65}
\]

where

\( i \) = phase number,

\( \lambda_i \) = flux linkage,

\( L_i \) = self inductance,

\( M_{im} \) = mutual inductance,

\( R_t \) = total reluctance of each branch,

\( E_i \) = phase back emf,

\( K_{Ei} \) = back emf constant,

\( B \) = number of magnet pole pairs,
l_m = length of the magnet, and
B_r = magnet residual flux density.

In a similar way, the one-phase-on energized torque can be found by applying the following equation

\[ T_i = K_{Ti} i_i \sin(\theta B + (i - 1) \frac{2\pi}{3}) \quad (3.66) \]

where \( K_{Ti} \) is motor torque constant, and it is equal to \( K_{Ei} \) in SI unit.

The lumped model results of back emf waveforms, inductance, and energized torque are compared with the experimental results in the following subsections.

### 3.3.2 Back Emf Calculation

The back emf magnitude and waveforms are critical quantities for design and control of a permanent magnet motor. The back emf was calculated by using the lumped parameter model and Equation (3.63). Figure 3.43 shows the results for the models of the staggered case. The models predict a back emf of the form

\[ E_i = K_{Ei} \sin(\theta B + (i - 1) \frac{2\pi}{3}) \theta \quad (3.67) \]

where

\[ E_i = \text{phase back emf}, \]
\[ K_{Ei} = \text{back emf constant}, \]
\[ i = \text{phase number}, \]
\[ B = \text{number of pole pairs}, \text{ and} \]
\[ \theta = \text{mechanical angle}. \]

As a result of using the assumptions made in the lumped magnetic model, the back emf wave shape is close to a pure sinusoidal curve as compared to the
back emf waveform of the measured value. Equation (3.68) was transferred to
EXCEL spread sheet program, which showed an approximate 17-percent error.
The lumped magnetic model with nonsinusoidal (more like trapezoidal, see
Figure 3.25) permanent magnet field distribution and permanent magnet pole to
pole leakage must be modeled to reduce this large error. This result is acceptable
in the preliminary design stage because it predicts a relatively close frequency
and magnitude of back emf function of motor speed. More detailed calculations
can be performed using FEA, and the results show close correlation.

![Graph showing back emf waveform comparison](image)

**Figure 3.43** Back Emf Waveform Comparison
3.3.3 Inductance Calculation

To obtain an accurate inductance value for the motor, both magnetic flux across the air gap and flux resulting from slot leakage must be modeled. The previously derived lumped model did not consider slot leakage, and the model results produced unreasonably large error in the inductance value. This subsection presents the development of a new model that includes slot leakage and was based on the multisegment 2-D FEA result with no magnets. Figure 3.44 shows one of the six-segment 2-D FEA magnetic flux plots. The permanent magnets were removed in this model because inductance values are a function of the stator coils and motor geometry.

From the FEA flux plot in Figure 3.44, a lumped magnetic model was developed as shown in Figure 3.45. The total flux through the energized coil can be found using the axial (air-gap) and slot leakage flux:

\[
\begin{align*}
R_L &= \frac{s_w}{\mu_0 l_1 T} \\
\Phi_R &= \frac{N \cdot i}{2 R_R}, \\
\Phi_L &= \frac{N \cdot i}{R_L}
\end{align*}
\]  

(3.68)  

(3.69)

where

- \( R_L \) = slot leakage reluctance,
- \( s_w \) = slot width,
- \( T \) = thickness of tooth
- \( l_1 \) = slot depth,
- \( \Phi_R \) = air-gap flux,
- \( R_R \) = air-gap reluctance, which has been defined in Equation (3.46),
- \( N \) = number of coil turns,
- \( \Phi_L \) = slot leakage flux established in loop, and
- \( i \) = applied phase current.
Figure 3.44 One-Phase-On FEA Model with Flux Plot

Figure 3.45 Lumped Parameter Model to Predict Inductance
The total flux cutting the coil is then [33]

\[ \phi_{\text{total}} = 3 \phi_R + 2 \phi_L \]  \hspace{1cm} (3.70)

Inductance was also calculated by examining the magnetic energy stored by the inductance in a current-carrying coil, resulting in the equation

\[ L_{\text{self}} = \left( \frac{4 \sum_{j=1}^{\infty} \phi_j^2 R_j}{i^2} \right) \]  \hspace{1cm} (3.71)

where \( j \) specifies the various fluxes going through the coil. This quantity was then reduced to the basic lumped model elements by applying Equation (3.69)

\[ L_{\text{self}} = \frac{4 N^2}{R_L} + \frac{3 N^2}{R_R} \]  \hspace{1cm} (3.72)

The values of inductance predicted by the lumped parameter magnetic model using Equation (3.72) showed close correlation with experimentally measured and FEA-calculated values, Table 3.3.

<table>
<thead>
<tr>
<th>Method used</th>
<th>Inductance (mH)</th>
<th>Error (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Prototype measurement</td>
<td>203</td>
<td>0.0</td>
</tr>
<tr>
<td>FEA model</td>
<td>184</td>
<td>9.4</td>
</tr>
<tr>
<td>Lumped model</td>
<td>177</td>
<td>12.8</td>
</tr>
</tbody>
</table>

Table 3.3 Inductance Values Summary
3.3.4 Energized Torque Analysis

The one-phase-on energized torque was calculated using the Equation (3.66) (as was the back emf waveform). The lumped parameter magnetic model result is presented in Figure 3.46. As explained in subsection 3.2.11, there are no experimental measurements available. The purpose of this calculation is to compare the energized waveform with the FEA result. The FEA result is higher than the lumped magnetic model result because of the embedded cogging torque in the FEA calculation. The peak torque value was normalized to protect the classification of this project. It can be seen from Figure 3.46 that the lumped magnetic model results closely approximate the 2-D FEA result. This correlation (8 to 10 percent error) is one of the many advantages of using a lumped magnetic model in the initial design stage to obtain the approximate output torque function of the rotor angle within a limited time frame.

Figure 3.46 Comparison of One-Phase-On Energized Torque Results
3.4 SUMMARY

This chapter described the calculation of several important design parameters and performance characteristics of the prototype dual air-gap, axial field, brushless, permanent magnet motor. The two methods of analysis used in this research – the finite element analysis (FEA) and the lumped magnetic circuit analysis – are explained, and validation for their use is presented.

The specific design parameters and performance characteristics addressed in this chapter are flux distribution, flux linkages, back emf waveform, inductance, and energized torque waveform.

Two well-known force/torque calculation methods using FEA results, virtual work method and Maxwell’s stress tensor method, were described and compared. A simple cogging torque calculation comparison was conducted using these two methods. Because of its accuracy, lower number of calculations involved, and its compatibility with the generic FEA modeling capability program package, the virtual work method was chosen.

Several critical FEA modeling techniques were examined, including the FEA model convergence test and permanent magnet modeling techniques. Developed FEA model results were compared and validated by experimental measurements. It was found that (1) the FEA model convergence test is an essential modeling step when FEA models included air gaps and permanent magnets and (2) the permanent magnet modeling technique is critical when the FEA model includes permanent magnet excitation. Previously, many authors had realized that, after having incurred unreasonably large errors in back emf waveform calculation and cogging torque analysis, the FEA model mesh density and permanent magnet modeling techniques were critical FEA modeling
considerations. In this study, the FEA models were developed based on the convergence test results and the permanent magnet model using the dipole moment (coercive method) application. Designed motor parameters, back emf waveform, phase inductance, and energized torque calculations were performed using 2-D quasi-static nonlinear FEA models.

Lumped magnetic circuit models were developed using FEA results. The back emf waveform, inductance value, and one-phase-on energized torque waveform were calculated using the developed lumped magnetic circuit model. The motor parameter calculations obtained from the FEA models and lumped magnetic models were compared with experimentally measured values from the prototype motors. Comparison of the calculations showed a close correlation.

A unique generic FEA modeling capability program was developed by the author during this study. The primary objective of this program was to mechanize the FEA analysis, from FEA modeling to result presentation, using FORTRAN programs, VAX command procedures, macro programs, and pre-/postprocessor input files. By specifying minimum information, including the number of rotor permanent magnet poles, the number of stator slots, the inner and outer radius, and the excitations, a 2-D or 3-D (or both) FEA motor model was developed with very little user intervention. This generic FEA modeling capability program was used to process extensive numbers of FEA models that were built and analyzed for back emf calculation, cogging torque analysis, and energized torque calculation. Currently, this capability program works only for dual air-gap, axial field motor configurations. Additional development for other motor configurations, such as the slotless axial field motor configuration, is under consideration.
The motor back emf waveforms were calculated from the analysis of many different snapshot FEA models, each at a different rotor location. The rotor movement was simulated by changing material properties on the leading and trailing edges of the permanent magnet. This simulation method supported use of the generic FEA modeling capability program. The motor torque constant was calculated using a result of the back emf calculation. The results of comparing the FEA-calculated back emf waveforms with the experimental prototype measurements validated using an FEA model as an accurate analysis tool, which can provide a method for evaluating motor designs.

Motor phase inductance values were calculated using multisegment 2-D FEA models, and the results were compared with experimental measured values. The calculated phase inductance values were also used to calculate the designed motor electrical time constant with approximated phase resistance value. In other studies, the inductance calculation using FEA models showed unreasonably large errors, but in this study the error rate was less than 10 percent. The lower error rate was mainly attributed to three factors: (1) the FEA model was developed using multisegment 2-D modeling technique rather than a single 2-D model, (2) the FEA model had a high-mesh density in the air gap, slots, and teeth area, and (3) the end turn length was relatively shorter than a typical radial winding length in the designed axial field motor.

One-phase-on energized torque was calculated using the FEA models. The maximum torque value was estimated using these FEA models, rather than using the actual prototype motor torque measurements. The preliminary design was validated by comparing the FEA-calculated value with the design requirement.
Lumped parameter magnetic models yielded acceptable results for the back emf, energized torque waveform, and inductance values, but they need refinement. The model results were validated by comparing them to FEA calculated and experimentally measured values.

In summary, axial field, permanent magnet motor parameters and performance characteristics were analyzed using the FEA and the lumped magnetic circuit model methods to obtain quantitative, designed motor parameters and characteristics, a requirement for design optimization and parameter sensitivity studies. In this study, it was found that when using the 2-D FEA method in conjunction with the generic FEA modeling capability program, a variety of useful design and performance analysis data can be obtained with minimum user interaction. FEA models have been developed in this research for a brushless, dual air-gap, axial-field permanent magnet motor. The effects of mesh discretization for the FEA model have been presented. The relationship between the total co-energy and the element-meshing density provides a useful means to gauge the required mesh size and computer time in an FEA analysis. In addition, the lumped magnetic circuit model can be used to estimate the back emf waveform, phase inductance value, and energized torque waveform.

The developed FEA models and lumped magnetic circuit models provide axial field permanent magnet motor design tools to analyze and improve existing designs and thus eliminates building prototypes to test various design parameters and performance characteristics. FEA modeling is useful for design and synthesis and is essential when significant geometric changes are made. A more refined lumped magnetic circuit model that includes flux leakage, particularly between the permanent magnets, is needed to improve accuracy of back emf waveform and energized torque waveform calculations.
CHAPTER 4

COGGING TORQUE, AXIAL FORCE ANALYSIS, AND THE EFFECT OF ROTOR MISALIGNMENT

4.1 INTRODUCTION

This chapter analyzes cogging torque, axial force variation, and rotor misalignment effect of a large-horsepower, brushless, dual air-gap, axial field permanent magnet motor. As part of the overall design process, cogging torque, axial force, and rotor misalignment effects are considered as major design and fabrication issues and are studied using 2- and 3-D FEA models. The nonlinear behavior of the stator magnetic material, the importance of relative permeability of the permanent magnet material and the complex air-gap geometry dictate using a nonlinear FEA method to obtain an accurate field solution.

The cogging torque, axial force variation, and rotor misalignment study requires a series of FEA models and analyses because all these calculations are a function of rotor position; thus, the generic FEA modeling capability program package was used extensively in this study, which reduced the engineering effort of building and analyzing many FEA models and simulations.

A multisegment FEA modeling technique is used to perform the cogging torque analysis, the axial force variation, and the rotor misalignment effect study. The primary purpose of this modeling technique is to simulate a
real 3-D physical system using 2-D FEA models and thus simplify the complicated cogging torque calculation. The cogging force is produced by the rotor permanent magnets attempting to align themselves with the stator teeth to maximize the magnetic flux crossing from the rotor permanent magnet to stator steel, i.e., the reluctance between the stator and rotor is minimized. In most permanent magnet motor configurations, every rotor permanent magnet is at the same relative position with respect to the stator teeth. Therefore, total cogging force/torque is equal to the algebraic sum of that cogging force/torque produced by each pole or permanent magnet.

Combinations of staggered and untaggered stators with skewed and unskewed permanent magnets are studied to discover a method to reduce cogging torque. In particular, the effects of stator staggering and rotor misalignment are closely examined.

The trade-off between cogging torque reduction and axial force variation is presented in this chapter. Before this research, cogging torque analysis was difficult to perform accurately, even using the FEA method [25]. As a result of this study, however, by using a generic modeling capability program package with an optimum number of elements in each section of the model (based on the convergence test result) and an improved permanent magnet modeling technique (coercive force or current density method), cogging torque can be calculated accurately.

The FEA results are compared to experimental measurements obtained from the prototype motor.
4.2 COGGING TORQUE ANALYSIS

This section examines the physical phenomena of cogging torque. Specifically, this section elaborates on two of six identified methods to reduce cogging torque; it details cogging torque calculation using FEA multisegment 2-D models; and it evaluates the effect of stator staggering on the output torque. A SIMULAB (an analysis and data manipulation software tool based on the MATLAB program) model is developed to calculate cogging torque from the FEA-calculated force data and to simulate the permanent magnet skewing effect. Finally, the cogging torque analysis results are summarized and trade-offs are discussed.

In a permanent magnet motor, cogging torque results from the interaction of the rotor permanent magnets with the magnetic steel teeth on the stator. For a slotted stator configuration, the air-gap permeance, or reluctance, is nonuniform for three reasons: (1) the shapes of stator teeth and slots, (2) the space between the rotor permanent magnet poles, and (3) the saturation of the stator lamination material. These nonuniform reluctances or magnetic flux paths cause air-gap flux density to vary with rotor position and result in cogging torque, which generates noise/vibration when the rotor rotates.

Reduction of cogging torque is a major consideration in the design of permanent magnet motors. Cogging torque usually adds an undesirable harmonic component to the torque-angle curve, resulting in torque ripple. Torque ripple produces vibration and noise, both of which may be amplified in variable speed drives when the torque frequency coincides with a mechanical resonant frequency of the stator or rotor.
As discussed in Chapters 1 and 2, the high-energy product, permanent magnet technology is a major contributing factor in the development of the brushless, permanent magnet motor as an alternative choice in many drive systems. A disadvantage of the high-energy density, permanent magnets that are now available is the high cogging torque. If high performance and quiet operation are required, cogging torque reduction is a significant design consideration, especially for underwater application where high-energy density, permanent magnets are used to develop high-power density motors.

Figure 4.1 depicts the unstaggered cogging force/torque as a function of rotor permanent magnet position. When the permanent magnets are aligned with the stator teeth, the cogging force is zero; that is, the x-directional force is zero and only the y-direction force (attractive force) exists. The positions shown in Figure 4.1 (a), (c), and (e), where the cogging force is zero, are called equilibrium positions. When the permanent magnets are between teeth, any small disturbance causes the magnets to restore themselves to the nearest aligned position; thus, unaligned cogging force positions are unstable. Because of symmetry, the cogging force is maximized when the permanent magnets are positioned halfway between stator teeth (points (b) and (d) in Figure 4.1). In Figure 4.2, the staggered stator cogging force/torque as a function of rotor permanent magnet position is shown. In this figure, the upper stator has been staggered a half-tooth pitch relative to the lower stator. In the staggered case, the physical distance between the stable and unstable positions are one-half that of the unstaggered case, thus the cogging force/torque frequency is doubled. Cogging force/torque magnitude depends on the tooth shape and the ratio between the tooth width and slot width. When the two cases (unstaggered and staggered) have the same tooth shape
Figure 4.1 Unstaggered Cogging Force as a Function of Rotor Position
Figure 4.2  Staggered Cogging Force as a Function of Rotor Position
and the same tooth-width-to-slot-width ratio, the magnitude of the staggered case will be less than that of the unstaggered case because there are less abrupt flux changes in the staggered case. The attractive y-directional force is a function of the strength of the rotor permanent magnet material and the air-gap length on both sides of the rotor. The attractive (axial) force can be balanced by precisely placing the rotor between the two stators; thus, the net attractive force on the rotor is zero because of the equal and opposite forces on the two stators. If the rotor is not placed exactly in the middle of the two stators, the attractive force is not balanced. (This unbalanced attractive force and rotor misalignment effect are discussed in Section 4.3.)

At a constant speed, the frequency of the cogging torque is a function of the total number of teeth covered by one rotor permanent magnet pole. For example, three teeth covered by one rotor permanent magnet pole will be one-half of the six teeth covered by the same magnet pole. The magnitude of the cogging force is directly proportional to the air-gap flux or strength of the rotor permanent magnet. The cogging torque is not only a function of the total number of teeth covered by magnet, but it is also directly proportional to the ratio of slot-to-tooth-width.

4.2.1 Cogging Torque Reduction Method

In this subsection, cogging torque reduction methods are presented and applied to the prototype axial field motor.

Previous studies of conventional radial field motors have shown that cogging torque can be reduced by a variety of techniques, but the most common methods are offsetting or staggering the stator and skewing the rotor permanent magnets. In radial field motors, the staggering of the stator
lamination stack with the winding is not difficult to apply. Although the rotor permanent magnet skewing method is not widely used, it has been studied by many authors to reduce cogging torque in conventional radial field motors [34,35].

Cogging torque in the designed axial field motor configuration can be reduced by (1) skewing of the stator slots or rotor permanent magnets, (2) increasing the number of slots, (3) using semi-closed stator slot openings, (4) using fractional slot/tooth pitch, (5) staggering stators, and (6) reshaping the permanent magnets. Of these six methods, two that have no effect on motor output torque and can be easily manufactured were selected: rotor permanent magnet skewing and stator staggering.

In the first method, skewing of the rotor permanent magnets is accomplished by using specially shaped magnet pieces that have a lopsided appearance as shown in Figure 4.3. Figure 4.4 shows a piece-by-piece skewing of the permanent magnet using six-segment permanent magnet pieces. The effective cross-sectional area is the same for the unskewed and skewed permanent magnets as shown in Figure 4.4. The skewing angle of the permanent magnet was selected based on the results of previous experience at KEČ. The optimum skewing angle has not been determined to date by ongoing quantitative studies. Figure 4.5 shows unskewed and skewed permanent magnets superimposed on the stator model, and Figure 4.6 depicts unskewed and skewed permanent magnets with a linearized stator model.

The second method for reducing cogging torque in the double stator axial field configuration is to stagger two stators; that is, one stator is rotated slightly so that its teeth are not aligned with those on the opposite stator, as
illustrated in Figure 4.7. Staggering the stator decreases the magnitude of the cogging torque because the air-gap flux changes are less abrupt. A similar phenomenon results from decreasing the slot/tooth pitch and increasing the number of slots and teeth. The frequency is doubled because the total number of slots and teeth exposed to the permanent magnet are doubled (see Figure 4.2).

Figure 4.3 Unskewed and Skewed Permanent Magnets
Figure 4.4 Permanent Magnet Skewing Using a Six-Segment Model
Figure 4.5  Skewed and Unskewed Permanent Magnets with Stator

Figure 4.6  Linearized Skewed and Unskewed Permanent Magnets
Figure 4.7 Staggered vs. Unstaggered Stator
4.2.2 Cogging Torque Calculation Using Finite Element Analysis

This section presents the cogging torque calculation using FEA multisegment 2-D models for both the staggered and unstaggered stators. The skewing of the rotor permanent magnet is simulated using a developed SIMULAB model and all the calculated FEA cogging force data (staggered and unstaggered stators).

The cogging torque is analyzed using six-segment 2-D FEA models. The six-segment models were chosen because comparisons with 3-D results and experiments show that it is possible to accurately predict the cogging torque with skewed permanent magnets using 2-D FEA models (thus reducing computer and engineering time). The cogging torque for each segment is calculated using the virtual work method (as discussed in Chapter 3):

\[
T_i(x) = \frac{W_{coi}(x + \Delta x) - W_{coi}(x)}{\Delta x} \cdot L_{ri} \cdot N_p \cdot l_{armi}
\]

(4.1)

where \(T_i\) is \(i^{th}\)-segment cogging torque, \(W_{coi}\) is \(i^{th}\)-segment total magnetic co-energy, \(x\) is rotor displacement, \(L_{ri}\) is \(i^{th}\)-segment radial thickness or model depth of the 2-D model, \(N_p\) is the total number of poles in the whole motor, and \(l_{armi}\) is \(i^{th}\)-segment lever arm length of each model.

The 2-D FEA models are used to calculate cogging torque by rotating the rotor permanent magnet through 30 steps (0.5 mechanical degrees per step) or 30 different rotor positions within one tooth pitch.

Figure 4.8 depicts magnetic flux contour plots from the staggered and unstaggered stators at a particular radial value (R=17 cm). Six multisegment
2-D FEA models for the unstaggered stator are used to generate the cogging force versus rotor position and electrical angle plot as shown in Figure 4.9 (a) and (b). In this figure, the six force curves represent six segment FEA cogging force calculation results. In all six curves, the analysis started at the first point (zero rotor position in Figure 4.9.a) where the permanent magnet is aligned with the stator teeth. As a result, no phase shift is seen between the various cogging curves in Figure 4.9.a. However, the shift is considered in the analysis described below. Even though the electrical angle for the tooth pitch of all six segments is the same, the physical length of the tooth pitch for each segment is different.

The six curves (C₁, C₂, C₃, C₄, C₅, and C₆) shown in Figure 4.9.a and b are the unstaggered stator cogging force calculated by six different segment FEA models. A second set of six cogging force curves (not shown) are also calculated for the staggered stator case using another six different segment FEA models. Each 2-D segment consists of 30 different 2-D FEA models within one tooth pitch where each model represents a different rotor position. The cogging force C₁ starts from the most inner radius at approximately R=14 cm, and C₆ at the most outer radius, about R=20 cm. Each cogging force result is obtained using 30 different FEA models to simulate the 30 different rotor positions. The number of FEA models for each segment is dependent on the rotor displacement for each movement (simulation step size). Two different step sizes (N=number of steps) showing displacement in the x-direction are examined. Figure 4.10 shows the comparison result, which indicates no significant difference. Because of the sufficient step size, the total number of FEA models in each segment is the same (30 FEA models) for all six segments.
Figure 4.8 Flux Plots of Unstaggered and Staggered Stator
Figure 4.9 Unstaggered Cogging Force vs. Rotor Position for Six Segments
Figure 4.10  Comparison of Two Different Number of Step Results
Both method of the cogging torque reduction (stator staggering and rotor permanent magnet skewing) requires mathematical manipulation of the calculated cogging force data from FEA models. Figure 4.11 shows the logic of the developed program. Each row represents one of the segments of the 2-D FEA model. An individual row comprises four major mathematical functions.

The first function generates a mechanical permanent magnet position (rotor position). To model permanent magnet skewing, a radius-dependent skewing factor is added to the mechanical position of each segment. Setting all skewing factors to zero represents an unskewed permanent magnet. In other words, a zero-skewing vector models an unskewed permanent magnet; a relative nonzero skewing vector simulates the skewed permanent magnet. The skewing vectors are calculated based on the permanent magnet skewing angle. Each segment skewing angle equals the total skewing angle multiplied by the ratio between the segment radial thickness and the total stator radial thickness.

The rotor electrical position functions, shown in the second column of Figure 4.11, convert mechanical permanent magnet positions to electrical positions (see Figure 4.9.b). The electrical position is required because the calculated cogging torque of all six segments with 30 different rotor positions is converted to electrical positions and stored as a look-up table. The electrical position is then used as input to a cogging force look-up table. The look-up tables for each segment are generated by a six-segment FEA model as shown in Figure 4.9. The cogging force for each segment is then multiplied by its respective lever arm radius because torque is force times the lever arm.
length. Each lever arm equals the radius from the origin of the stator to the mid-point of each segment in the radial direction. Finally, the cogging torque contributions from each segment are summed to calculate the total cogging torque as a function of electrical position.

Figure 4.11 SIMULAB Model for Cogging Torque Calculation
The total number of 2-D FEA models can be summarized as follows. There are six 2-D FEA model segments representing one-pole section of the prototype motor. Each segment consists of three tooth pitches and each tooth pitch consists of 30 different 2-D FEA models, which represent 30 different rotor position. There is a six-segment FEA model for the staggered stator and a six-segment FEA model for the unstaggered stator. The total number of 2-D models for the unstaggered stator was 180 FEA models, and the total for the staggered stator was 180.

The results of the cogging force FEA calculation for unstaggered and staggered comparisons, based on an unskewed permanent magnet are shown in Figure 4.12, with three experimentally measured points. The cogging torque is clearly reduced by staggering the stators. Experimentally measured peak cogging torque results on the staggered prototype motor verifies the FEA-calculated value. (Peak cogging torque was not measured on an unstaggered stator because an unstaggered stator was not available.)

The cogging torque analysis using FEA models in conjunction with the SIMULAB program is summarized in Figure 4.13. In this figure, all possible combinations of the considered cogging torque reduction methods are applied and shown as a function of rotor position in electrical degrees. Experimentally measured peak cogging torque of the prototype motor, with staggered stators and skewed permanent magnets, is also shown in Figure 4.13. This measured peak value approximates the peak value predicted by the FEA model. In Figure 4.13, two different scales are used, one with a broken-line horizontal grid-line and the other with a solid grid line. Two scales are used because the maximum cogging torque (55Nm) and the minimum cogging torque (0.5Nm) differ by two orders of magnitude, which cannot be
Figure 4.12 Cogging Torque FEA Results
portrayed by a single scale in one figure. The unskewed permanent magnet with unstaggered stators (1) and skewed permanent magnet with unstaggered stators (3) are represented on the broken-line horizontal axes that have a vertical range of 55 Nm to -55 Nm. The unskewed permanent magnet with staggered stators (2) and skewed permanent magnet with staggered stators (4) are plotted on the solid-line horizontal axes that have a vertical range of 2.2 Nm to -2.2 Nm. The prototype motor measured peak cogging torque values are also plotted on the solid-line horizontal axes with a vertical range of -2.2 Nm to 2.2 Nm. Both sets of axes share the same origin and x-axis (electrical degree).

Table 4.1 provides cogging torque reduction methods and all possible combinations with calculated results. The skewed permanent magnet with staggered stator was the most efficient combination because it has the lowest cogging torque and was, therefore, applied to the first prototype motor (unit 1).

In this study, cogging torque calculations were also attempted using a lumped parameter magnetic circuit model, based on the following assumptions.

1. Relative permeability of the magnetic materials is infinite.
2. Saturation of the magnetic materials is not considered.
3. Relative permeability of the magnet is the same as air.
4. There are simplified flux paths.

Significant errors (40 - 60%), attributed to the use of the above assumptions, were found when this model was used. These assumptions were not made in FEA calculations.
Figure 4.13 Graphical Summary of Cogging Torque Analysis
Table 4.1 Numerical Summary of Cogging Torque

<table>
<thead>
<tr>
<th>Cogging torque (Nm)</th>
<th>Stators</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Unstaggered</td>
</tr>
<tr>
<td>Permanent magnets</td>
<td></td>
</tr>
<tr>
<td>Unskewed</td>
<td>55 (1)</td>
</tr>
<tr>
<td>Skewed</td>
<td>13 (1/4)</td>
</tr>
</tbody>
</table>

* The values in parentheses are the approximated ratios.  

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4.2.3 Staggered Stator Effect on Output Torque

This subsection studies the staggered stator effect on output torque. The purpose of staggering the stator is to reduce cogging torque. As shown, the cogging torque for the staggered stator is reduced to 4 percent of that for an unstaggered stator. To determine the impact on the output torque by staggering the stators, the torque constant was calculated using the back emf constant. The back emf constant can be calculated by measuring many back emf waveforms with different motor speeds. The torque constant can be inferred from the back emf constant in normal conditions in most permanent magnet motors based on the following assumptions: (1) stator lamination material and the permanent magnet operate in the linear region in B/H curve, (2) eddy current effects are negligibly small, and (3) the permanent magnet is directly responsible for the torque production [25].

A wide range of back emf waveforms are measured with motor speed (100 - 1300 rpm) to calculate the back emf constant. From this measured back emf waveform as a function of motor speed, the motor torque constant was calculated. The torque constant is equal to the back emf constant in the MKS system.

The test stand with a dc motor as a back drive motor (see Figure 3.29) and the designed axial field prototype motor was used for the back emf waveform measurement. The back emf waveforms were measured using a Tektronix TDS544A four-channel digitizing oscilloscope. Because the bed of the test stand was not secured to the ground, the test was conducted between 100 rpm and 1132 rpm. One stator can be adjusted (depending upon the staggering angle), and is locked in place.
First, the staggering angle between two stators was checked by measuring two back emf waveforms of the same phase terminals from two stators; for example, the back emf waveform 1 from stator 1, phase A and waveform 2 from stator 2, phase A. By comparing the phase shift of these two waveforms, the accuracy of the stator staggering angle could be measured. Figure 4.14 shows two back emf waveforms for the staggered stator case. The phase shift of two waveforms is also shown in this figure; the staggering angle is 30 electrical degrees (half tooth pitch). When the stator returned to the unstaggered stator, the same procedure was used to ensure the two stators were aligned; that is, there was no phase shift between the two waveforms (see Figure 4.15). Because the purpose of figures 4.14 and 4.15 is to measure phase shift for staggered and unstaggered stators, magnitude scales are not shown.

The back emf waveforms, functions of shaft speed, were measured and plotted. The frequency and peak-to-peak voltage were also recorded in each plot. These data were transferred to an EXCEL spread sheet program to calculate back emf constant and thus torque constant. The output torque and output power were predicted from these calculations. Figure 4.16 shows the back emf peak-to-peak voltage vs. rotor speed for the unstaggered and staggered stator. From this plot two motor torque constants, unstaggered and staggered stator, were calculated and are shown to be nearly identical. The results indicate that the staggered stator does not affect the output torque, but does reduce the cogging torque.

The eddy current effect on stator lamination was also studied using back emf waveform measurements with different rotor speeds.
Figure 4.14  Staggered Stator Back Emf Waveforms

Figure 4.15  Unstaggered Stator Back Emf Waveforms
Figure 4.16 Comparison of Back Emf Constant
For rotor speeds up to 1000 rpm, the back emf vs. speed curve is linear. At the tested speed of 1132 rpm, the back emf constant slope changes (see Figure 4.16). As discussed in the previous section, the test setup was not secured to ground. As a result, motor speed was limited to 1132 rpm because of safety considerations. The back emf slope change at 1132 rpm may be a measurement error or the eddy current effect starting at that speed. Further testing is required to confirm the result.

4.3 AXIAL FORCE / ROTOR MISALIGNMENT EFFECT

Section 4.3 presents analysis results on the prototype motor axial force variation on the stator and rotor misalignment effect. The purpose of the axial force calculation on the stator is to (1) find the staggering effect on the axial force variation, (2) identify the stator back iron thickness for structural stiffness, and (3) select bonding material between the stator back iron and the cooling plate. The rotor misalignment effect study provides important information on axial field motor design with regard to bearing and shimming material selection. The rotor misalignment also causes an adverse effect on the cogging torque reduction, which results in an increase in noise and vibration. FEA 2-D models with two force calculation methods (Maxwell's stress tensor and virtual work method) are used in this study. For the axial force variation on the stator the Maxwell's stress tensor method was used because integration is easier using the Maxwell's stress tensor method, and for the rotor misalignment effect study, the virtual work method (co-energy method) was used. The axial force variation result was verified by vibration tests on the staggered and unstaggered stator versions and is presented in more detail in subsections 4.3.1 and 4.3.2.
4.3.1 Background

In the course of this research, it was found that the primary disadvantage of the axial field motor technology was the strong axial attractive force between the stator and the rotor permanent magnet material. The axial field motor configurations that have been previously studied and built have a single-stator, a single-rotor configuration. This configuration can produce large attractive forces between the stator core magnetic material and rotor permanent magnets, particularly when the high-energy density permanent magnets are used. This attractive force on the rotor can be reduced by incorporating a double stator with a single rotor, which counterbalances the attractive forces and also provides additional thrust on the rotor. If the rotor is placed exactly in the center of the two stators, then the net attractive force on the rotor surface can theoretically be reduced to zero because of the force balance. As a result, a double-stator with a single-rotor configuration axial field motor was designed and prototyped to obtain high-power density, high-efficiency, and quiet operation.

During the prototyping period, a rotor misalignment was detected. This rotor misalignment was the result of limitations in the current manufacturing technology, which does not provide "perfect" alignment because the bearing and shimming materials are not precise and strong enough for the designed high-power and high-energy density axial field motor. Thus, bearings and shims cannot keep the rotor centrally positioned between the stators. When the rotor is perfectly centered, each stator produces the same attractive force in opposing directions and thus cancels the effect of the forces. However, if the rotor is off center, the difference in attractive forces caused by this rotor misalignment increases as a square
function of the rotor misalignment. Therefore, in this research it was necessary to obtain an estimate of the attractive forces between the rotor and stators as a function of the rotor misalignment. With this information a bearing and shimming material can be selected to withstand the possible attractive forces created by rotor misalignments. The FEA results of this research outline the range of forces that can be expected in a typical double-stator motor compared with those found in the same motor with one of the stators removed. This result can be used to select the rotor bearings and determine acceptable manufacturing tolerances.

4.3.2 Axial Force Variation on Stator

This subsection presents the FEA model calculations for the axial force variations on the staggered and unstaggered stator using Maxwell's stress tensor method, the consequences of the staggered stator, and the results of a vibration test that was conducted to verify the trends of the axial force variation.

To find the staggering effect on the axial force variation, the force calculations on the staggered stator and unstaggered stator using FEA models as a function of the rotor position are analyzed. Figure 4.17 shows unstaggered and staggered stator models; the bold lines on the stator surfaces are the contour in which the force density is calculated using Maxwell's stress tensor method (Equations 3.13 and 3.14).

A significant result was obtained from the FEA calculation of the axial force. The FEA model predicted that the axial force variations would increase if the stator were staggered. The FEA results, shown in Figure 4.18, are plots
Figure 4.17 Two FEA Models for Axial Force Calculation
Figure 4.18 Axial Force Variation
of axial force variation with the staggered and unstaggered stators as a function of the rotor position. The motor stators are staggered in order to reduce the cogging torque. This result shows that trade-offs exist between cogging torque reduction and axial force variation and that these factors must be considered in the double-stator, axial field motor design. Two courses of action could be considered.

The first course of action would be a two-step process: (1) build the dual air-gap, axial field motor with unstaggered stators and skewed rotor permanent magnets, which provides lower axial force variation, and then (2) compensate for the cogging torque by advanced control of the stator currents [17].

The second course of action would be to keep the stator staggered and increase the stator back iron thickness, which reduces the axial force effect. The cogging torque would be minimized and the power density would be reduced by the increased stator back iron thickness.

This axial force variation also impacts the attachment between the stator back iron and cooling plate. If the bonding is not secure, then the attachment will be broken. The axial force variation result may be used as a guide line for the bonding material and selection of attachment method.

The results of axial force analysis and the cogging torque are provided in Table 4.2. This table has four major cells of numbers, and each cell is diagonally divided into two triangular sub-cells. The upper sub-cell describes cogging torque ratio (based on unstaggered and unskewed), while the lower sub-cell describes the axial force ratio. To interpret the table, consider the
upper left-hand major cell as a baseline, or model for a motor with unskewed magnets and unstaggered stators. The remaining three major cells contain the factors by which cogging torque or axial force is affected, by either magnet skewing, stator stagger, or a combination of both features. A number less than one means a reduction, while a number greater than one denotes an increase.

Table 4.2 Cogging Torque and Axial Force Trade-off

<table>
<thead>
<tr>
<th>Cogging torque</th>
<th>Stators</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Unstaggered</td>
</tr>
<tr>
<td>Axial force</td>
<td></td>
</tr>
<tr>
<td>Permanent magnets</td>
<td>Unskewed</td>
</tr>
<tr>
<td></td>
<td>Skewed</td>
</tr>
</tbody>
</table>

Although staggering the stators significantly decreases the cogging torque, it increases the axial force variation by a factor of nine. In order to verify this trend, the axial vibration measurement is made using an accelerometer attached in the axial direction, while the prototype motor is back-driven by a dc motor. The motor vibration was measured for both staggered and unstaggered stator cases, the results of which are shown in Figure 4.19. The experimentally measured axial vibration for the staggered case is significantly higher than that for the unstaggered case. These experimental results are consistent with the results from the FEA models. Because cogging torque is more easily controlled than axial force, the optimal design for the prototype unit 2 motor included unstaggered stators.
Figure 4.19  Vibration Test Comparison
4.3.3 Rotor Misalignment Effect

This subsection presents the rotor misalignment effect on a dual air-gap axial field motor. In addition, a possible way to detect this misalignment using back emf measurement is proposed. The predicted rotor misalignment on the prototype motor, using back emf measurements, was compared with the FEA-calculated results to approximate the rotor misalignment.

During prototype motor construction, it became evident that the dual air-gap configuration may cause several manufacturing difficulties. For example, the axial alignment of the rotor, i.e., an inability to ensure equal air gaps on either side of the rotor, was a concern. This condition is referred to as “rotor misalignment” and is illustrated in Figures 4.20 and 4.21. Rotor misalignment results from inaccurate dimensional tolerances in the bearing system and shimming that support the rotor between the two stators. When the rotor is misaligned, it creates many operational problems, e.g., the rotor may rub the stator surface by an excessive axial attractive force between the rotor permanent magnets and the stator steel as the rotor is rotating. In addition, the rotor matrix can experience warping, which becomes more severe as the misalignment increases. Because the rotor permanent magnets are closer to one stator than the opposite stator, the back emf waveform magnitude differs as a function of motor shaft speed in the same phase of the two stators thus altering the output torque and possibly causing the controller circuit to malfunction. Finally, the mechanical bearings are subjected to unbalanced forces, resulting in reduced bearing lifetime.
Figure 4.20 Rotor Aligned vs. Misaligned, Stator Staggered vs. Unstaggered

Figure 4.21 Side View of Aligned Rotor vs. Misaligned Rotor
The same FEA models for the cogging torque analysis and calculation method were used to determine the rotor misalignment force and effect. The total magnetic co-energy for each FEA model was calculated as a function of each rotor permanent magnet position. This calculation was done in the same manner as were the cogging torque calculations. Figure 4.22 shows two FEA models (A and B) for the unstaggered stator case to calculate axial force for a rotor position. The rotor with the permanent magnet portion of the model (model A) shifted slightly toward one stator (movement in the axial direction in the prototype motor and in the y-direction in Figure 4.22). The axial force was then determined using the virtual work method. Because the rotor is misplaced in the y-direction, Equation (3.2) must be rewritten as

\[ F_y = \frac{W_{co}(y + \Delta y) - W_{co}(y)}{\Delta y} \]  \hspace{1cm} (4.2)

where \( F_y \) is the attractive force in the y-direction, \( W_{co} \) is the total magnetic co-energy in the model, and \( y \) is the direction in which the rotor is misaligned.

The first step in analyzing the rotor misalignment effect was to study the total force between the rotor permanent magnet and the single stator, calculated by using 2-D FEA models as a function of air gap between the rotor and the stator. The result of this calculation is shown in Figure 4.23. As shown in this figure, the force is almost the square function of the air gap between the stator and the rotor permanent magnet.

The second step in analyzing the rotor misalignment effect was to study the staggering of the stator vs. rotor misalignment effect as a function of rotor position. A series of 2-D FEA models were analyzed to calculate the co-energy variation in the staggered stator case. In these models, the rotor
Figure 4.22  FEA Models for Rotor Misalignment Effect Calculation
Figure 4.23  Axial Force Function of Displacement of Rotor
permanent magnet was misaligned by 0.25mm, which means one air gap is 0.25mm longer than the other air gap. Figure 4.24 shows the co-energy and frequency variation of the results. It is clear from Figure 4.24 that the misalignment of the rotor adversely influences the cogging torque reduction effort. These results are explained as follows. If the rotor is misaligned toward Stator 1, the air gap between rotor and Stator 1 is smaller than the other air gap. Because of this air-gap difference, the magnetic flux increases by the square function of the air-gap length. The rotor experiences a strong cogging and axial force from Stator 1, and this force dominates the cogging force. Therefore, the cogging torque reduced by the staggered stator is eliminated by rotor misalignment. Rotor misalignment causes cogging torque increases and axial force variation increases. Therefore, the noise and vibration increase.

The final step in analyzing the rotor misalignment effect was to study the method of detecting rotor misalignment. This step did not require making measurement holes in the prototype motor housing, which is totally enclosed and sealed (see Figure 4.25). A simple back emf comparison method that enabled measurement of the magnitude and frequency of the back emf waveform from the stator winding, which is a function of motor shaft speed and air-gap flux, was considered. Therefore, if the back emf waveforms obtained from the same phase from the two stators are the same at all speeds,
Figure 4.24 Aligned vs. Misaligned (0.25mm) Rotor
Figure 4.25 Assembled Prototype Motor
then the rotor is in perfect alignment; otherwise, the rotor is misaligned. The back emf waveform magnitude is also a function of stator geometry, phase winding length, and the end turn winding scheme. The winding inductance and resistance of all three phases from two stators are measured and compared in Tables 4.3 and 4.4, respectively. The inductance values for all three stator windings (phase A, B, C) are measured using a digital LCR meter (Hewlett Packard, 4263A) with accuracy +/- 0.05 % of reading. The differences are within 2 percent. Thus, the two-stator windings are nearly identical.

Table 4.3  Winding Inductance Measured Values

<table>
<thead>
<tr>
<th></th>
<th>Stator 1 (µH)</th>
<th>Stator 2 (µH)</th>
<th>Difference (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phase A</td>
<td>149.4</td>
<td>152.0</td>
<td>1.71</td>
</tr>
<tr>
<td>Phase B</td>
<td>146.0</td>
<td>147.8</td>
<td>1.22</td>
</tr>
<tr>
<td>Phase C</td>
<td>145.2</td>
<td>147.0</td>
<td>1.2</td>
</tr>
</tbody>
</table>

Table 4.4  Winding Resistance Measured Values

<table>
<thead>
<tr>
<th></th>
<th>Stator 1 (Ω)</th>
<th>Stator 2 (Ω)</th>
<th>Difference (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phase A</td>
<td>2.10</td>
<td>2.13</td>
<td>1.4</td>
</tr>
<tr>
<td>Phase B</td>
<td>2.00</td>
<td>2.08</td>
<td>3.8</td>
</tr>
<tr>
<td>Phase C</td>
<td>2.17</td>
<td>2.20</td>
<td>1.34</td>
</tr>
</tbody>
</table>
The back emf waveforms were measured using the prototype motor to compare the rotor misalignment effect with FEA calculations. Back emf waveform measurements were conducted by using the same test setup as shown in Figure 3.29. Figure 4.26 shows the prototype motor measured back-emf curves from a slow speed, and Figure 4.27 shows the FEA-calculated back emf with the same speed applied. A slow speed was chosen in the experimental measurement to eliminate the high-speed eddy current effect on the stator winding, thus affecting the back emf waveform. Based on the experimental measurement (Figure 4.25), this author projected the prototype rotor misalignment to be 0.25mm and used that estimate in the FEA models (Figure 4.26). The 0.25mm misalignment means one air gap is 0.25mm longer than the other air gap. It can be seen from Figure 4.26 the prototype motor measured two back emf waveforms that the magnitude is approximately 10 percent different in the same phase for the two stators; it is assumed that the rotor is not perfectly aligned in the prototype motor. The calculated FEA waveforms magnitude differences between these two curves are about 15 percent. The difference between the experimental result and the FEA-calculated results is attributed to the estimation in the FEA model. The phase shift between the two waveforms is a result of the staggered stators.

From this comparison between the prototype-measured and the FEA-calculated, it can be concluded that (1) the rotor misalignment exists in the prototype motor (unit 1), (2) the rotor misalignment can be detected by measuring the back emf waveforms and (3) this rotor misalignment can be estimated using FEA calculations.

The axial force variation and rotor misalignment effect results are significant factors in dual air-gap, axial field motor design because they clearly
Figure 4.26 Prototype Measured Back Emf Waveforms

Figure 4.27 FEA-Calculated Back Emf with Rotor Misalignment
show the trade-offs between cogging torque reduction, rotor alignment technique, and noise/vibration reduction. The axial force variation of the staggered stators is proportional to the square of the rotor misalignment -- an expected result because the attractive magnetic force on the rotor increases rapidly as the air gap to one stator decreases. The effectiveness of stator staggering in reducing cogging torque is also diminished when the rotor is misaligned because the air-gap flux density is the squared function of the distance between the rotor permanent magnet and the stator teeth surface. Since cogging torque and axial force are both a function of the air-gap flux density and stator geometry, cogging torque and axial force increase by rotor misalignment. Finally, rotor misalignment can be detected using back emf magnitude comparisons, which can be verified by FEA models.

4.4 SUMMARY

This chapter analyzed the cogging torque, the axial force variation, and the rotor misalignment effect of a large-horsepower, brushless, dual air-gap, axial field permanent magnet motor. As part of the overall design process, cogging torque, axial force, and rotor misalignment effects were considered as major design and fabrication issues and were studied using 2-D quasi-static nonlinear FEA models. Because the cogging torque, the axial force variation, and the rotor misalignment studies required a series of FEA models and analyses, which are a function of rotor position, the generic FEA modeling capability program package was used extensively in these calculations. The generic modeling capability requires minimum interface with the computer
and automatically provides results and plots of all calculations, based on input data.

Two cogging torque reduction methods, staggering stators and skewing permanent magnets, were studied. Staggered and unstaggered stators with skewed and unskewed permanent magnet combinations were analyzed for methods to reduce cogging torque. The effect of stator staggering and rotor misalignment effects were also examined. A trade-off between cogging torque reduction and axial force variation was determined in this chapter. Previously, cogging torque analysis was the most difficult analysis to perform accurately, even using the FEA method. In this chapter, a generic modeling capability program package with an optimum number of elements in each section of the model, based on the convergence test result and an improved permanent magnet modeling technique, were used to perform an accurate cogging torque calculation. Using this technique enabled the first thorough and accurate study of cogging torque. The FEA results were validated by comparing them to experimental measurements obtained from the prototype motor.

Multisegment 2-D FEA models were developed and used to calculate cogging torque, axial force variations, and rotor misalignment effects. The 2-D model (multisegment) was a simplified model; however, comparisons with 3-D results showed that the 2-D model could accurately predict the cogging torque, axial force, and back emf waveform. The model results were validated by comparing those results with experimentally measured values. The multisegment 2-D FEA method was developed to reduce the computational time and engineering cost incurred with using 3-D FEA modeling.
Permanent magnet skewing effect on the cogging torque was analyzed using cogging torque analysis in conjunction with a developed SIMULAB program. By skewing the rotor permanent magnets and staggering the stators, the cogging torque was ideally reduced to 1 percent of that for a motor with no skewing and staggering. The most significant factor in reducing cogging torque was stator staggering. However, staggering the stators also increased the axial force variations. Staggered stator effect on output torque was also examined. It was found that staggering stator has no impact on output torque. In a motor having a light rotor structure, such stresses by axial force variations cause mechanical vibration, noise, and fatigue.

In this study, it was found that misalignment of the rotor had a detrimental effect on both axial force and cogging torque. Misalignment of the rotor increased axial force variations by almost nine times and counteracted the improvements achieved by stator staggering to reduce cogging torque. Increased axial force variations can be controlled by increasing stator back iron thickness. For the high-power density motor design, this increased stator back iron reduces the power density of the motor. Therefore, increasing stator back iron is not acceptable. The FEA model was shown to be accurate in predicting the cogging torque and rotor misalignment effect by comparing torque and back emf to experimentally determined values for the staggered stator prototype motor.

Based on the findings of this chapter, the new prototype motor (unit 2) was redesigned by (1) skewing the permanent magnets to reduce cogging torque, (2) unstagerring the stator to minimize the axial force variation, (3) increasing the air gap to minimize noise/vibration, and (4) selecting a new bearing and shims based on the axial force calculation result.
CHAPTER 5

MOTOR EDDY CURRENT POWER LOSS CALCULATION

5.1 INTRODUCTION

This chapter discusses the eddy current power loss comparison between un laminated and laminated rotor permanent magnet material. The prototype axial field motor rotor contains Nd B Fe permanent magnets embedded in the rotor structure. The rotor permanent magnets are laminated to reduce the eddy current power losses. A simplified un laminated and laminated rotor permanent magnet material with an applied magnetic field from the stator excitation current is shown in Figure 5.1.

![Diagram of un laminated and laminated permanent magnet blocks.](image)

Figure 5.1 Un laminated and Laminated Permanent Magnet Block
A trade-off study is conducted between eddy current power loss reduction and total cost. The purpose for the eddy current power loss comparison between laminated and unlaminated rotor permanent magnet is three-fold. First, the cost of the lamination manufacturing process is extremely high for rare-earth permanent magnet materials. Specifically, the cost of a laminated permanent magnet is almost ten times the cost of an unlaminated permanent magnet block of the same size. Second, the interlaminar insulation and bonding material create a small air gap between the laminations. Because of this air gap, the permanent magnet packing factor is reduced; that is, for a given volume, a reduced amount of permanent magnet material is present. Third, laminated permanent magnets cause a nonuniform surface magnetic field distribution due to the air gaps between laminations.

In the design of high-power density, high-efficiency, permanent magnet motors, the maximum operating temperature constraints caused by the stator winding insulation limits, rotor permanent magnet demagnetization, and bearing overheating have been critical design issues. To increase the efficiency and power density of the motor, the material usage in the motor must be maximized, and all possible power losses (such as current-carrying winding losses, eddy current and hysteresis losses, and mechanical losses) [36,37,38] must be minimized. This study focused only on the eddy current power losses in the rotor permanent magnet.

Trying to reduce eddy current losses requires knowing the location and magnitude of the eddy currents and the resultant heat produced in the motor parts. Knowing how much heat is produced is important because
overheating can cause stator winding insulation breakdown and rotor permanent magnet demagnetization. Additionally, the failure of brushless permanent magnet motor performance is directly affected by stator winding breakdown or the rotor permanent magnet demagnetization. It is imperative, then, that all possible heat sources be minimized in the motor design stage. Conduction current power loss ($I^2R$) in current-carrying conductors and eddy current loss in the stator core are the major heat sources in the stator.

Eddy current and hysteresis losses in the stator core materials have been studied extensively because they limit the motor performance and winding temperature, and therefore, limit the life of the motor and winding insulation. Lamination of the stator core material is the most common method of reducing eddy current losses and has been successfully used for many years to reduce eddy current power losses. Eddy current losses in the current-carrying conductor can also be reduced by using multistranded wires (see subsection 2.2.3).

There is, however, no published research on the subject of eddy current power loss in the rotor permanent magnet. Because high-power density and high-efficiency motor design requires the highest energy density permanent magnet material available, which is Nd B Fe, there is no merit in studying eddy current power loss for different rotor permanent magnet materials, such as alnico, ferrite, and Sm Co. (These materials exhibit far less sensitivity to demagnetize in the high temperature than does Nd B Fe.) Therefore, it is important to determine how to reduce eddy current power losses, and to what extent, to establish a trade-off between power loss reductions and cost.
The purpose of this chapter is to compare the total eddy current power loss ratio between unlaminated permanent magnets and laminated permanent magnets using an electromagnetic lumped model and FEA methods; the results are compared with the experimentally measured eddy current power loss ratio.

The underlying principles of eddy current are discussed before analyzing the laminated permanent magnet. Eddy currents are produced by either a time-varying magnetic field, applied to a stationary conductive material, or by a uniform steady magnetic field (permanent magnet field) applied to a moving electrical conductive material. The eddy current phenomenon, where a time-varying magnetic field produces an emf voltage that establishes a current flow in a conductive material, was first discovered by Faraday in 1831.

Figure 5.2 is a graphical representation of the applied time-varying magnetic field and induced eddy current in the conductive material in a macroscopic view.

Figures 5.3, a through d, present a microscopic view of the eddy current phenomenon and its effect when a conductive material is laminated. Figure 5.3a shows an applied time-varying magnetic field (bold arrows) in a block of electrically conductive material and the eddy current flow loops around the applied field (circular line around bold arrows). Figure 5.3b shows the resulting eddy current flow in the conductive material block. There is some eddy current cancellation between neighboring loops when the eddy current flows are opposite each other. Therefore, the major eddy current flow loop resides at the outer portion of the conductive material block. Figures 5.3c and 5.3d show the lamination effect on the eddy current flow. As illustrated in
Figure 5.2 Macroscopic View of Eddy Currents in a Conductive Material

These two figures, the resulting eddy current must flow through a longer path in Figure 5.3d (loop 1 + loop 2) than the path in Figure 5.3b (loop 1); thus, eddy current power loss is reduced by increasing the eddy current flow path or resistance. Mathematically, the emf can be assumed to be the same for the un laminated and laminated (Figures 5.3b and 5.3d) because the emf is a function of a cross-sectional area with an applied magnetic field. Based on the assumption that the air gap between the lamination is negligibly small, the generated emf voltage is the same for the un laminated and laminated. The relationship between emf voltage and eddy current is $\text{emf} = \frac{d\phi}{dt} = I_{\text{eddy}} R_{\text{eddy}}$ where emf is the generated emf voltage in the conductive material, $\phi$ is the magnetic flux enclosed by conducting surface area, $I_{\text{eddy}}$ is eddy current flow in the conductive material, and $R_{\text{eddy}}$ is the eddy current path resistance. Thus, for a given time-varying magnetic field, less eddy current flow occurs.
Figure 5.3 Microscopic View of Eddy Current Flow in a Conductive Material
when the resistance is high. This results in eddy current power losses \( P = I_{\text{eddy}}^2 R_{\text{eddy}} \) that are low.

There are several methods to reduce eddy current power losses in the conductive material. The first and most common method is lamination of the conductive material. The second method is mixing electrically nonconductive additives in the conductive material -- a complicated and expensive process. A third method to reduce the eddy current power losses is using anisotropic conductive material.

The method used in this study to reduce the eddy current power losses in the permanent magnet is lamination of the rotor permanent magnet. Orthogonally laminating the rotor permanent magnets in the direction of the imposed time-varying magnetic field causes the induced eddy current loop to experience higher resistance to eddy current flow, as shown in Figure 5.3 (c and d). The degree of eddy current power loss reduction attributed to lamination of rotor permanent magnets is a direct function of lamination thickness. There is a trade-off, however, between lamination thickness and the permanent magnet material packing factor, and it must be carefully studied to find the optimum lamination thickness.

Eddy current power loss was calculated using two models: the lumped parameter electromagnetic circuit model and the FEA model. The loss ratios calculated from these two models were compared with the experimentally measured loss ratio. There were two reasons for using a ratio comparison instead of direct quantitative comparisons: (1) to find a trade-off between eddy current power loss reduction and the consequences of permanent magnet lamination and (2) to use a simple calculation method and model that could
be conducted within a limited time frame. Because the ratio comparison was used, it was not necessary to use the same shape and dimensions as the experimentally measured permanent magnet.

5.2 EDDY CURRENT EFFECTS IN MOTORS

This section discusses eddy current effects in a permanent magnet motor.

There are four adverse consequences of eddy currents in permanent magnet motors. First, there is heat generation in the motor, which causes irreversible demagnetization of the rotor permanent magnet, stator winding insulation breakdown, and bearing failure by overheating. To prevent this overheating problem, a cooling system must be provided. The cooling system, however, adds unwanted weight and complicates the manufacturing process.

The second consequence of eddy currents in permanent magnet motors is the creation of magnetic reaction fields. There are two major eddy current flows: (1) eddy current flow in the stator lamination produced by movement of the rotor permanent magnet and (2) eddy current flow in the rotor matrix and permanent magnet created by the stator PWM excitation current waveform. Those two eddy current directions are always in opposition to each other; thus, there are increased magnetic reaction fields. These reaction fields decrease the effective fields applied from the rotor permanent magnet and stator current excitation field.

The third consequence of eddy currents in permanent magnet motors is the stray forces resulting from the interaction of the applied magnetic fields
from the stator excitation and rotor permanent magnet to the eddy current magnetic field. One of these stray forces can be defined as an eddy current damping force between the stator lamination and the rotor permanent magnet. Finally, the fourth consequence is that motor parameters such as resistance and inductance change with eddy currents [39] and affect total motor power losses and the functioning of PWM circuits. An increase in eddy current density results in an increased ac resistance of the current-carrying conductor and a reduced inductance value. Since the motor electrical time constant is a direct function of these two parameters (resistance and inductance), the variation of these two parameters is not desirable.

In permanent magnet brushless motor design it is important to predict and estimate the eddy current effect so that motor parameter, efficiency, and performance can be determined. In this study, eddy current power losses are calculated for a stalled rotor subjected to the stator excitation ac PWM currents with different magnitudes and frequencies. For the stalled (speed = 0), permanent magnet, brushless motor, the losses are primarily attributed to the $P^2R$ heating in the current-carrying windings. At no-load maximum speeds, the losses are attributed to the $P^2R$, eddy currents, hysteresis, windage, and friction. Between stalled rotor and maximum speed, the losses are a combination of all $P^2R$ loss, eddy current and hysteresis loss, windage loss, and friction loss. The stator core material and current-carrying conductor experience induced eddy currents from the moving rotor permanent magnets and they experience imposed ac currents from the inverter-fed PWM excitations. Even though a permanent magnet motor rotates at a speed synchronous with the fundamental PWM phase current waveform
frequency, eddy currents are produced by higher harmonics in the current waveform from the PWM circuit.

The stalled rotor state was examined because the measurements of a moving or dynamic state for eddy current losses is extremely difficult to perform. Eddy current power losses for laminated and un laminated permanent magnets are calculated and compared in Sections 5.3 through 5.6.

5.3 LUMPED MAGNETIC CIRCUIT ANALYSIS

This section compares the results of the eddy current power loss calculation of laminated and un laminated permanent magnets using lumped circuit analysis. Lumped electromagnetic circuit equations are developed to calculate eddy current power losses in the rotor permanent magnet. The lumped circuit analysis is used because the derivation and solution are relatively simple compared to those of the Maxwell's field equation. The derived equations are based on the combination of Faraday's law and Ohm's law. The permanent magnet shape and dimensions used in this calculation are shown in Figure 5.4. A simplified rectangular conductive block is used to derive lumped circuit equations and to compare un laminated and laminated permanent magnet eddy current power losses. Figure 5.5 provides permanent magnet material properties that were used in this calculation. Using this data and information, the skin-depth function of applied frequency was plotted and is shown in Figure 5.6.

Skin depth plot is used in the lumped electromagnetic circuit model. The skin depth or penetration depth is given by

\[ \delta = \frac{1}{\sqrt{\pi \cdot f \cdot \mu \cdot \sigma}} \]  \hspace{1cm} (5.1)
where permeability and conductivity of the material are $\mu = \mu_0 \cdot \mu_r$, and $\sigma$.

Figure 5.4 Simplified Permanent Magnet Block
**NOMINAL MAGNETIC PROPERTIES**

<table>
<thead>
<tr>
<th>Property</th>
<th>Value 1</th>
<th>Value 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>$B_r$ (Residual Induction)</td>
<td>11500 Gauss</td>
<td>1150 mTesla</td>
</tr>
<tr>
<td>$H_C$ (Coercive Force)</td>
<td>10900 Oersteds</td>
<td>860 kA/m</td>
</tr>
<tr>
<td>$H_{pi}$ (Intrinsic Coercive Force)</td>
<td>14000 Oersteds</td>
<td>1100 kA/m</td>
</tr>
<tr>
<td>$BH_{max}$ (Maximum Energy Product)</td>
<td>$31 \times 10^4$ G0e</td>
<td>245 kJ/m$^3$</td>
</tr>
<tr>
<td>Operating Point for Maximum Energy Product</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$B_O$</td>
<td>5600 Gauss</td>
<td>560 mTesla</td>
</tr>
<tr>
<td>$H_O$</td>
<td>5600 Oersteds</td>
<td>440 kA/m</td>
</tr>
<tr>
<td>Permeance Coefficient at $B_O/H_O$</td>
<td>1.0</td>
<td></td>
</tr>
<tr>
<td>Reversible Permeability (Recall Permeability, $\mu_{rev}$)</td>
<td>$1.08 - 1.10$</td>
<td></td>
</tr>
<tr>
<td>Reversible Temperature Coefficient of Induction</td>
<td>$-0.10 ~ ^\circ C$ (25$^\circ$ to 100$^\circ$ C)</td>
<td></td>
</tr>
<tr>
<td>For Magnets Operating near $BH_{max}$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Reversible Temperature Coefficient of Coercive Force</td>
<td>$-0.60 ~ ^\circ C$ (25$^\circ$ to 100$^\circ$ C)</td>
<td></td>
</tr>
<tr>
<td>$H_K$</td>
<td>12000 Oersteds</td>
<td>$\geq 950$ kA/m</td>
</tr>
</tbody>
</table>

**TYPICAL PHYSICAL PROPERTIES**

<table>
<thead>
<tr>
<th>Property</th>
<th>Value 1</th>
<th>Value 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Constituents</td>
<td></td>
<td>TM-RE-B</td>
</tr>
<tr>
<td>Density</td>
<td>.270 lbs/in$^3$</td>
<td>7.4 g/cm$^3$</td>
</tr>
<tr>
<td>Curie Temperature</td>
<td>590 °F</td>
<td>312 °C</td>
</tr>
<tr>
<td>Recommended Maximum Operating Temperature</td>
<td>302 °F</td>
<td>150 °C</td>
</tr>
<tr>
<td>Coefficient of Linear Expansion</td>
<td>$5.0 \times 10^{-5}$ °C</td>
<td>to Orientation</td>
</tr>
<tr>
<td>(25$^\circ$ to 200$^\circ$ C)</td>
<td>$-3.0 \times 10^{-4}$ °C</td>
<td>11 to Orientation</td>
</tr>
<tr>
<td>Electrical Resistivity</td>
<td>160 µ Ohm-cm</td>
<td></td>
</tr>
</tbody>
</table>

**MECHANICAL PROPERTIES**

<table>
<thead>
<tr>
<th>Property</th>
<th>Value 1</th>
<th>Value 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Compressive Strength</td>
<td>$1.1 \times 10^3$ lbs/in$^2$</td>
<td>$7.6 \times 10^4$ N/m$^2$</td>
</tr>
<tr>
<td>Tensile Strength</td>
<td>$1.2 \times 10^4$ lbs/in$^2$</td>
<td>$8.3 \times 10^4$ N/m$^2$</td>
</tr>
<tr>
<td>Flexural Strength</td>
<td>$3.6 \times 10^7$ lbs/in$^2$</td>
<td>$2.4 \times 10^7$ N/m$^2$</td>
</tr>
</tbody>
</table>

*Most permanent magnet materials are a class of materials that lack ductility and are inherently brittle. Such materials should not be designed as structural components in a circuit. Measurement of such properties as hardness and tensile strength are not feasible on commercial materials with these inherent characteristics. Therefore, specifications of these properties are not acceptable. These materials may be machined using techniques that are normally applied to difficult-to-machine alloys.

Preliminary data as of 6/16/85

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Figure 5.5 Permanent Magnet Material Property Data Sheet
Figure 5.6 Skin-Depth Function of Frequency for Permanent Magnet

Eddy current in the conductive material is directly related to the amount of applied magnetic flux and the emf induced in that material. Therefore, it is important to find the relation between time variations of the magnetic flux enclosed by the conductive material and the induced emf voltage within that material. Refer to Figure 5.7. It can be seen that when the magnetic flux is decreasing with time and the induced emf is positive, a
counterclockwise eddy current is produced in the conductive material. Eddy current polarity produces a magnetic field that increase the magnetic flux. When the magnetic flux is increasing with time, the induced emf is negative, thereby producing a clockwise current. The polarity of the current generates a magnetic field, which decreases the magnetic flux. This is consistent with Lenz's law [32].

![Diagram showing induced emf voltage, induced eddy current, and applied magnetic field]

Figure 5.7 Applied Magnetic Field, Induced Emf, and Induced Current

Based on the above discussions, the derivation of the lumped electromagnetic circuit equation can be started using the following well-known theory. A time-varying magnetic field within a stationary closed path induces an emf voltage and associated eddy current (see Figure 5.8). The induced emf voltage is given by

$$E = -\iint \frac{\partial \mathbf{B}}{\partial t} \cdot d\mathbf{S}$$  (5.2)
By Stoke's theorem

$$\nabla \times \mathbf{E} = \frac{\partial \mathbf{B}}{\partial t}$$

(5.3)

which is in a differential form.

Now, by replacing the closed path with a thick conductive plate, and calculating the closed surface area of the conduction plate, the total flux enclosed by the cross-sectional area can be calculated as

$$\Phi = \iint \mathbf{B} \cdot d\mathbf{S}$$

(5.4)

This new model with a thick conductive plate is depicted in Figure 5.9. The plate is now cut by a time-varying magnetic field given by

$$\mathbf{B} = B_0 \cos(\omega t) \hat{z}$$

(5.5)

Figure 5.8  Applied Magnetic Field with Conducting Bar
The magnetic flux enclosed by the conductive plate and directed in the z-direction is given by

\[ \Phi = a \cdot 2b \cdot B_0 \cdot \cos(\omega t) \]  
(5.6)

The induced emf around a rectangular thick conductive plate in the x-y plane, as shown in Figure 5.9, can be calculated using Equation (5.2)

\[ \text{EMF} = -\frac{d\Phi}{dt} = a \cdot 2b \cdot \omega \cdot B_0 \cdot \sin(\omega t) \]  
(5.7)

Figure 5.10 depicts a simplified model for the laminated and un laminated permanent magnet. In this figure, only four laminations are shown; however, the prototype motor permanent magnet has 40 laminations in each permanent magnet. All calculations in this section are based on the assumption that the air gaps between the laminations are negligibly small.
Figure 5.10 Current Flow in Unlaminated and Laminated Permanent Magnet
compared to the thickness of the each lamination. Based on this assumption, the total induced emf voltage is the same for the un laminated and laminated permanent magnet because the cross-sectional areas are assumed to be the same. The induced emf voltage in both the laminated and un laminated permanent magnet blocks can be calculated for an applied $B_x$ from Equations (5.5) and (5.7). The total flux flow through the plate is then

$$\Phi = L \cdot D \cdot B_0 \cdot \cos(\omega t)$$  \hspace{1cm} (5.8)

where $L$ is the y-directional length, $D$ is the x-directional length and the induced emf voltage can then be calculated as

$$\text{EMF} = \frac{d\Phi}{dt} = L \cdot D \cdot \omega \cdot B_0 \cdot \sin(\omega t)$$  \hspace{1cm} (5.9)

The magnitude of eddy current resulting from the induced emf is dependent on the eddy current flow path resistance or flow length. Figure 5.10 shows two models for the laminated and un laminated permanent magnets with one skin depth and mean length of the current flow path. By examining this figure, it is noticed that the total path of the current flow in the laminated permanent magnet is longer than in the un laminated permanent magnet ($\text{loop}1 < \text{loop}1 + \text{loop}2 + \text{loop}3 + \text{loop}4$). This results in a larger resistance and a smaller power loss in the laminated permanent magnet. The resistance calculation is based on one skin-depth penetration of the applied field. One skin depth carries 63 percent of the eddy current [31]. Additional skin depths would contribute negligibly to eddy current power loss, so it is sufficient to consider one skin depth to approximate eddy current calculation. The general form of resistance is given by

$$R_{dc} = \frac{L}{\sigma \cdot S}$$  \hspace{1cm} (5.10)
where $R_{dc}$ is dc resistance, $l$ is the mean current flow length, $\sigma$ is conductivity of the conductive material, and $S$ is the cross-sectional area of the current flow. When the time-varying magnetic field is applied to the conductive material, the ac resistance can be calculated by

$$R_{ac} = \frac{l}{\sigma \cdot S_{\text{effective}}}$$ (5.11)

where $R_{ac}$ is the ac resistance, $\sigma$ is conductivity of the conductive material, $S_{\text{effective}}$ is the effective cross-sectional area (approximately one-skin depth area), and $l$ is the mean length of the current flow path.

The mean length of the current flow path ($l$), can be calculated by

$$l \equiv (2 \cdot L + 2 \cdot D - 4 \cdot \delta)$$ (5.12)

for the unlaminated permanent magnet and

$$l \equiv (2 \cdot L + 4 \cdot b - 4 \cdot \delta)$$ (5.13)

where $2b$ is the lamination thickness ($t$) and $b = D/2N$, $N$ is the total number of laminations for each section of the laminated permanent magnet.

Table 5.1 provides three basic motor design materials and their properties. By using Equation (5.1) the skin depth vs. frequencies for each material was calculated and plotted in Figure 5.11. It can be seen that the permanent magnet material skin depth is greater than that of steel and copper. From this figure, the lamination thickness requirement for all three materials can be determined. As applied frequency increases, the thickness of the lamination must decrease to minimize the eddy current power losses. The logarithmic scale was used in Figure 5.11 because of the skin-depth differences among these three materials, which is almost two orders of
magnitude. This graph was used to compare the skin depth between three different motor design materials.

![Graph of skin depth vs. frequency for different materials]

Figure 5.11 Skin Depth vs. Frequency for Different Materials

Table 5.1 Material Properties for Skin-Depth Plot

<table>
<thead>
<tr>
<th>Material</th>
<th>Copper</th>
<th>Steel (M19)</th>
<th>Nd B Fe (Crumax 315)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistivity (Ω-cm)</td>
<td>172E-4</td>
<td>135E-4</td>
<td>160E-6</td>
</tr>
<tr>
<td>Permeability (μ_r)</td>
<td>1.0</td>
<td>2500</td>
<td>1.09</td>
</tr>
<tr>
<td>Density (kg/m³)</td>
<td>8400</td>
<td>7350</td>
<td>7400</td>
</tr>
</tbody>
</table>
Using the calculated skin depth of the permanent magnet material, the effective cross-sectional area of the permanent magnet can be calculated

\[ S_{\text{effective}} = \delta \cdot h \]  

(5.14)

for the laminated and un laminated permanent magnet where \( \delta \) is the skin depth of the permanent magnet and \( h \) is the thickness (z-direction) of the lamination stack. The un laminated permanent magnet ac resistance can be calculated as

\[ R_{ac} \equiv \frac{(2 \cdot L + 2 \cdot D - 4 \cdot \delta)}{\sigma \cdot \delta \cdot h} \]  

(5.15)

and laminated ac resistance can also be calculated as

\[ R_{ac} \equiv \frac{N \cdot (2 \cdot L + 2 \cdot D/N - 4 \cdot \delta)}{\sigma \cdot \delta \cdot h} \]  

(5.16)

In Figure 5.12, the ac resistance vs. frequency is plotted using normalized values. (The maximum values of the ac resistance and skin depth at zero frequency (dc) are divided by themselves). As shown in this figure, the ac resistance value increases as a function of the inverse exponential as frequency increases. Knowing the induced emf voltage (Equation 5.9) and the ac resistance (Equation 5.15 and 5.16), the eddy current power loss for the laminated and un laminated permanent magnets can be calculated from

\[ P_{\text{eddy}} = \frac{\text{EMF}^2}{R} = I^2 R \]  

(5.17)

\[ P_{\text{eddy}} = \frac{(LD)^2}{2L + 2D - 4\delta} \]  

(5.18)

for un laminated and
\[ P_{\text{eddy}} = N \frac{(LD)^2}{2L + 2\frac{D}{N} - 4\delta} = \frac{(LD)^2}{2NL + 2D - 4\delta} \]  

(5.19)

for laminated permanent magnet \((N = \text{number of lamination})\).

![Graph showing AC resistance vs. frequency and skin depth vs. frequency.](image)

**Figure 5.12** AC Resistance vs. Frequency for the PM Material

Table 5.2 provides a calculated eddy current power loss comparison of unlaminated and laminated permanent magnets. The results of the eddy current power loss using the lumped model shows an almost 15-fold improvement for the laminated permanent magnet compared to the unlaminated permanent magnet.
Table 5.2 Eddy Current Power Loss Comparison (@10KHz)

<table>
<thead>
<tr>
<th></th>
<th>Unlaminated PM</th>
<th>Laminated PM</th>
<th>Ratio</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistance (Ω)</td>
<td>4.00E-4</td>
<td>6.14E-3</td>
<td>15</td>
</tr>
<tr>
<td>Power loss (Watt)</td>
<td>(B_z)^2 (5.685E+5)</td>
<td>(B_z)^2 (3.704E+4)</td>
<td>15</td>
</tr>
</tbody>
</table>

Limitations of this derived equation are threefold: (1) if lamination is thinner than the one-skin depth, then it is difficult to calculate eddy current power loss accurately, (2) detailed calculation requires a new model, and (3) it calculates one-skin depth eddy current power loss only. For the ratio comparison of the eddy current power loss, this derived equation is sufficient.

5.4 FINITE ELEMENT ANALYSIS

This section presents the eddy current power loss calculations for the laminated and unlaminated rotor permanent magnet using the FEA method. Simple 3-D FEA models were analyzed to meet this objective. In the FEA model, the permanent magnet and the surrounding air were modeled using 3-D hexa and tria elements for laminated and unlaminated cases. The two modeling techniques used to model the laminated permanent magnet air gap between laminations were (1) the 3-D and 2-D element combination method and (2) the anisotropic material property method. Because of its accuracy, the 3-D and 2-D element combination method was adopted for use in this study.

Eddy current power losses in the stator lamination materials have already been calculated using FEA method by several authors [40] and [41].

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The 3-D FEA model convergence test, model simplification, and skin-depth effect have been considered in FEA models. Calculated FEA results of the ratio between laminated and un laminated permanent magnets were compared with experimental measured values. FEA models were built based on a simple rectangular rotor permanent magnet with an ac excitation current loop around it (see Figure 5.13).

5.4.1 FEA Model Description

For an accurate analysis of the eddy current power losses, a 3-D FEA model was required to study the nature of the eddy current phenomenon, end turn effect, and magnetic fringing. In the FEA models, it was not necessary to model the stators because the objective of this study was to compare the eddy current power losses ratio in the permanent magnet material for the laminated and un laminated cases.

The FEA model consists of the permanent magnet block with current loop around the permanent magnet. (See Figures 5.13 and 5.14 for the layout of the permanent magnet with the current loop around it.) The permanent magnet and current excitation loop are surrounded by air elements. To model the fringing and end-turn effects in the model, the air element was extended from the permanent magnet and current excitation loop. Due to existing symmetry, only one-quarter of the entire model was analyzed. (The FEA model simplification and reduction process using the symmetry condition is presented in subsection 5.5.2.) The approximated permanent magnet shape and dimensions that were used in FEA models are shown in Figure 5.15. Figure 5.16 shows FEA models for the surrounding air and permanent magnet elements. The permanent magnet elements are separated
in this figure for clarity. This model consists of 4124 total elements with 5252 grid points. The material properties and excitations of the permanent magnet and current loop used in the FEA model are summarized in Table 5.3.

![Diagram of permanent magnet and current loop](image)

**Figure 5.13** Top View of the Permanent Magnet with the Current Loop
Figure 5.14 3-D View of Permanent Magnets and Current Loops
Figure 5.15 Shape and Dimensions of Permanent Magnet Used
Figure 5.16  FEA Model (4124 Total Element)
Table 5.3 Material Properties and Excitation

<table>
<thead>
<tr>
<th>Material Property</th>
<th>Relative permeability</th>
<th>Resistivity (Ohm-cm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permanent magnet</td>
<td>1.05</td>
<td>160E-6</td>
</tr>
<tr>
<td>(Nd B Fe)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Air</td>
<td>1.0</td>
<td>0.0</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Excitation</th>
<th>Peak current (Amp)</th>
<th>Frequency range (KHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current loop</td>
<td>25 - 125</td>
<td>0.5 - 70</td>
</tr>
</tbody>
</table>

5.4.2 FEA Model Checking and Model Simplification

This subsection describes how the FEA models are checked for convergence to verify the accuracy of the model results. The purpose of this check is to validate that the total number of elements in the FEA model is sufficient to obtain an accurate eddy current power loss calculation. In addition, the procedure for FEA model simplification using the designed motor symmetry condition is also presented.

A. FEA Model Checking

The accuracy of the any FEA calculations are primarily dependent on the resolution or density of meshing in the model; and specifically, the accuracy of FEA eddy current calculations mainly depends on the mesh density in the one skin-depth area. In this study, four 3-D models with four
different numbers of elements (with four elements in the one skin-depth area) are created: an initial coarse mesh of 2042 elements, a refined mesh of 4124 elements, a further refined mesh of 6530 elements, and an extremely refined mesh of 7728 elements. All four 3-D models are built in the same manner by creating a 2-D (x- and y-direction) element using the quad and tria element type and a 3-D model created by extruding the base-plane 2-D element mesh in the z-direction and assigning appropriate material properties, boundary conditions, and excitations to the current-carrying coil loop, permanent magnet, and air regions. Figure 5.17 shows the 2-D (quad and tria element) and 3-D element types (hexa and penta element) used in this study.

![2-D FEA model elements](image)

Quad element

![3-D FEA model elements](image)

Hexa element

![Tria element](image)

Penta element

Figure 5.17 Two- and Three-Dimensional FEA Element Used in the Model
As shown in Table 5.4, all four FEA calculation results of the four different models are quite similar, indicating adequate convergence for all four models. However, 2042 3-D elements is a large number of elements for a given geometry and dimensions of the FEA model. Extensive eddy current calculations using different PWM frequencies and range of applied current amplitudes were conducted using the total number of 4124 elements. Based on the FEA results presented in Table 5.4, it has been decided that the 4124-element model is accurate enough to obtain total eddy current power loss in the permanent magnet material.

<table>
<thead>
<tr>
<th>Total number of elements</th>
<th>Hexa element</th>
<th>Penta element</th>
<th>Magnetic co-energy (joules)</th>
<th>Power loss (watt)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2042</td>
<td>1702</td>
<td>340</td>
<td>3.56214</td>
<td>13.20006</td>
</tr>
<tr>
<td>4124</td>
<td>3436</td>
<td>688</td>
<td>3.56444</td>
<td>13.21435</td>
</tr>
<tr>
<td>6530</td>
<td>5442</td>
<td>1088</td>
<td>3.56480</td>
<td>13.22660</td>
</tr>
<tr>
<td>7728</td>
<td>6456</td>
<td>1272</td>
<td>3.56823</td>
<td>13.22905</td>
</tr>
</tbody>
</table>

B. FEA Model Simplification

The purpose of the FEA model simplification is to determine the optimum model size (minimum number of elements) using the motor symmetry conditions. Because FEA calculations using 3-D elements are a time-consuming process, it is preferable to use the minimum number of elements.
This subsection presents the model simplification steps using symmetry conditions and the FEA calculation results obtained from each model. The FEA model simplification steps are shown in Figure 5.18 (a through d). All four steps (a through d) are modeled and analyzed using 3-D FEA models, and the results are summarized in Table 5.5. The stators are not modeled since the purpose of this study is to compare eddy current power losses ratio of the laminated and un laminated permanent magnets. Therefore, the FEA model consists of the permanent magnet block with the current loop around the permanent magnet. The permanent magnet and current excitation loop are surrounded by air elements.

The height (in the z-direction) of the permanent magnet is relatively large, and the magnetic field generated from the stator excitation is assumed to be confined to one-half or less of the permanent magnet material; therefore, the top current excitation loop is removed. To verify this assumption, two FEA models were built and eddy current power losses were calculated using a two-current loop configuration (top and bottom, Figure 5.18a) and a single-current loop configuration (Figure 5.18b). The results (see Table 5.5) show that for the single loop with the full permanent magnet model, the eddy current power loss is approximately one-half the value of the double loop with full permanent magnet model. This result is attributed to application of half excitation. Furthermore, the top current excitation loop has a slight effect on the bottom excitation loop (see Figures 3.26 and 3.27).

In Figure 5.18c, a new FEA model with only one-half of the permanent magnet in z-direction is analyzed, and the result (see Table 5.5) shows only a small difference from that of the single loop/full permanent magnet model. From this result, it can be concluded that the permanent magnet may also be
Figure 5.18 Model Simplification Steps
divided in half without affecting the accuracy of the eddy current power loss. Further FEA model simplification is achieved by dividing the previous model (Figure 5.18c) by one-half (see Figure 5.18d) resulting in the model having approximately only one-half of the power loss of the previous FEA model. This last FEA model (Figure 5.18d) is extensively used in this section to compare the eddy current power losses in the laminated and un laminated permanent magnet material. This FEA simplification step is an essential process in reducing the complexity of the model, which ultimately reduces engineering time and computational resources.

<table>
<thead>
<tr>
<th>FEA models</th>
<th>Power loss (watt)</th>
<th>Normalized power loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>Two current loops with full permanent magnet model (Figure 5.18a)</td>
<td>18.22905</td>
<td>1.000</td>
</tr>
<tr>
<td>Single current loop with full permanent magnet model (Figure 5.18b)</td>
<td>8.54942</td>
<td>0.469*</td>
</tr>
<tr>
<td>Single current loop with one-half permanent magnet model (Figure 5.18c)</td>
<td>8.51296</td>
<td>0.467*</td>
</tr>
<tr>
<td>One-half current loop with one-quarter permanent magnet model (Figure 5.18d)</td>
<td>4.46611</td>
<td>0.245*</td>
</tr>
</tbody>
</table>

* Accounts for small flux flow from stator 1 to stator 2.
5.4.3 FEA 3-D and 2-D Element Combination Technique

The laminated permanent magnet consists of the permanent magnet laminations with interlaminar insulation material between the laminations. This interlaminar material is a nonmagnetic, nonelectrical conductive material, which is similar in magnetic field to the air-gap between laminations. To model the laminated permanent magnet material, the interlaminar material between laminations must be modeled. Modeling of the small air gap is difficult in general, but important because the accuracy of the FEA results varies with the density of the mesh in the air gap.

Two different modeling techniques, which incorporate small air gaps in the FEA model, are examined in this study to provide accurate results with a minimum of CPU time and modeling effort: (1) the 3-D and 2-D element combination technique and (2) the anisotropic material property (permeability and conductivity) technique.

In the first technique, the air gap between the permanent magnet laminations is modeled using 2-D elements, which requires thickness in the z-direction. The thickness represents the air-gap size between the 3-D permanent magnet lamination elements. The accuracy of this method has been proven in many technical papers [40,42]. Figure 5.19, an FEA model using the 3-D and 2-D element combination technique, shows the 2-D air-gap elements with 3-D air elements around it. In this figure, 3-D permanent magnet lamination elements were extruded for clarity. The advantage of the 3-D and 2-D element combination technique is that the air-gap size can be changed without remodeling. Figure 5.20 shows the unlaminated permanent magnet model. In this model, the only concern is the total number of
elements in the model and the modeling in the skin-depth area. The skin-depth modeling consideration is discussed in subsection 5.4.4. The permanent magnet elements are extruded to show the number of elements in the skin-depth area. The current excitation loop around the permanent magnet elements is omitted in this figure. The 3-D and 2-D element combination technique is used in this study because of its ease of use and accuracy. In addition, this technique can be automatically applied with the generic FEA modeling program package.

The second technique uses anisotropic material properties to simulate the lamination effect. In this technique, the equivalent permeability and conductivity in all three directions (x,y,z) are calculated using derived equations; that is, the effective anisotropic material properties are calculated. Table 5.6 shows the calculated anisotropic material properties for the 3-D FEA models used in this study. Derived equations to calculate these anisotropic material properties can be found in the MSC/EMAS application manual or user manual.

Table 5.6 Calculated Anisotropic Material Properties

<table>
<thead>
<tr>
<th>Material</th>
<th>Permeability (henrys/meter)</th>
<th>Conductivity (siemens/meter)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air</td>
<td>1.0</td>
<td>0.0</td>
</tr>
<tr>
<td></td>
<td>x-direction 1.09</td>
<td>x-direction 6.25E+3</td>
</tr>
<tr>
<td>Permanent magnet</td>
<td>y-direction 1.09</td>
<td>y-direction 6.25E+3</td>
</tr>
<tr>
<td></td>
<td>z-direction 0.003</td>
<td>z-direction 6.25E+1</td>
</tr>
</tbody>
</table>

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Figure 5.19 FEA 3-D and 2-D Combination Model
Figure 5.20 Permanent Magnet Elements Extruded in an FEA Model
5.4.4 Skin-Depth Consideration in Eddy Current Calculations

In general, when solving an eddy current problem, the penetration fields in the conductive material are near the surface portions of the electrically conductive material. For this reason, it is sufficient to model the surface regions where the eddy currents are expected to flow. However, it is necessary to determine or estimate the required skin depth that must be modeled. For a conductive material, eddy currents fall off exponentially with depth, reaching 63 percent of its surface density in a distance equal to one skin depth (Figures 5.21 and 5.22) [31]. This skin-depth phenomenon can be accurately modeled by two to four elements per skin depth. The part of the 3-D FEA model with current loop excitations and calculated eddy current flows are shown in Figure 5.21. In this model, three elements are used in one skin-depth area. The magnitude of the eddy current is represented by the length of the arrows in Figure 5.21. Reasonably accurate models can be obtained by estimating the skin depth and providing the appropriate fineness of modeling near the surface area. Detailed modeling is not done for skin depths beyond a depth of three skin layers because the field decreases to less than 0.5 percent of its surface value and continues decreasing [41].

Figure 5.22 shows the results of the 3-D FEA eddy current calculation. It can be seen that the eddy current exists only inside the permanent magnet material, where 63 percent of the eddy currents exist in one skin depth. This figure also shows the applied magnetic flux density produced by the current excitation loop around the permanent magnet block. The applied magnetic field decays rapidly at the edge of the permanent magnet block; the field penetration decays exponentially at the surface of the conductive material.
(Scales have been omitted because values are normalized and two plots are represented in the figure. The significance of this illustration is the established trend.)

Figure 5.21  Applied Current Loop and Induced Eddy Current
Figure 5.22 Magnetic Flux and Eddy Current Density Distribution
In this subsection, one skin depth was modeled using three element layers in one skin-depth area to verify the FEA model accuracy and the skin effect. Calculations in the subsequent subsections use four element layers in one skin-depth area to obtain even more accurate results.

5.4.5 FEA Results

In this subsection, eddy current power loss comparison between laminated and un laminated rotor permanent magnets is presented. The purposes of the FEA eddy current power loss comparison are (1) to find trade-offs between permanent magnet lamination production cost and eddy current power loss reduction and (2) to estimate the eddy current power losses in the permanent magnet of the rotor as a result of high frequency, high harmonic content inverted fed PWM current excitation.

Accuracy of the results from the FEA models used in this study was enhanced because (1) convergence testing was conducted, (2) the model was simplified by using symmetry condition, (3) the 3-D and 2-D combination technique was used to accurately model inter-laminar air gaps, and (4) skin-depth mesh density (4 element layers in one skin-depth) are considered.

The laminated permanent magnet FEA model (Figure 5.19) and the un laminated permanent magnet FEA model (Figure 5.20) are analyzed with different ac current excitations (25 amp-peak to 125 amp-peak) and different frequencies (1 kHz to 10 kHz). The purpose of these analyses is to find the eddy current loss sensitivity between the applied ac current and the applied frequency.
Figure 5.23 provides the eddy current power loss comparison between un laminated and laminated permanent magnets, using different applied frequencies (1 kHz to 10 kHz), with an ac current excitation (125 amp-peak). The results show that eddy current power loss can be reduced by lamination of the rotor permanent magnet. In addition, the eddy current power losses in un laminated permanent magnets rapidly increase as the frequency increases, compared to the gradual increase in laminated permanent magnets (lamination thickness 3 mm).

![Graph showing eddy current power loss vs. applied frequency](image)

Figure 5.23 Eddy Current Loss vs. Applied Frequency (@125 Amp-turn)

Figure 5.24 presents the eddy current power loss comparison between un laminated and laminated permanent magnets using different ranges of ac current excitations (25 amp-peak to 125 amp-peak) with a constant frequency (10 kHz). FEA model results indicate that applied ac current or applied magnetic field has more influence (slope changes more rapidly) on eddy
current power loss than does applied frequency. The results were obtained by using a lamination thickness (3 mm) of the permanent magnet. Further study is required to confirm this result using different lamination thicknesses.

![Graph showing eddy current power loss vs. applied current excitation](image)

**Figure 5.24 Eddy Current Loss vs. Applied Current Excitation (@ 10KHz)**

From the comparison of Figures 5.23 and 5.24, the sensitivity of the eddy current power losses by applied frequencies and applied current excitations can be obtained. The applied current excitations have a greater effect on the eddy current power loss than does the applied frequency. Eddy current power loss for an unlaminated permanent magnet is almost a square function of the applied magnetic field and frequency. The laminated
permanent magnet has less sensitivity to the applied magnetic field and frequency. This result supports the necessity for laminating permanent magnets.

Table 5.6 shows comparisons of eddy current power loss results obtained using the 3-D FEA method and an experimental test conducted using a single stator with a permanent magnet. The FEA results comparison was made between laminated and un laminated permanent magnets with an applied ac current excitation of 125-amp peak and frequency of 10 kHz. The experimental data was obtained applying two 217-amp peak current with a fundamental frequency of 10-kHz triangular current waveforms (the worst-case PWM waveform) in the stator phase A and phase B. (Detailed information on the experimental test is contained in Section 5.5.) As shown in Table 5.6, the FEA calculation eddy current power loss reduction is seven, the experimental test result showed a five-fold reduction (4.7).

<table>
<thead>
<tr>
<th>Method used</th>
<th>Unlaminated magnet</th>
<th>Laminated magnet</th>
<th>Ratio</th>
</tr>
</thead>
<tbody>
<tr>
<td>FEA calculation</td>
<td>22.691 (watt)</td>
<td>3.242 (watt)</td>
<td>7</td>
</tr>
<tr>
<td>Test measured</td>
<td>113.74 (watt)</td>
<td>24.2  (watt)</td>
<td>4.7</td>
</tr>
</tbody>
</table>

5.5 EXPERIMENTAL TESTS

Experimental tests comparing the eddy current power loss between laminated and un laminated rotor permanent magnets were conducted by
KEC using an axial field motor stator with inverter circuit. The eddy current power loss comparison test was performed on a modified motor consisting of one stator with a single permanent magnet. The other stator was removed to allow access to the motor during the test.

A test designed to quantify high-PWM frequency eddy current power losses under abnormal operating conditions, such as a stalled rotor condition, was conducted. Triangular PWM current waveforms were applied to the stator windings (phase A and phase B) at a fundamental frequency of 10 kHz (Figure 5.25). This experimentally applied triangular current waveform was analyzed using Fourier series analysis, the results of which are shown in Table 5.8. Temperature rise in the rotor permanent magnet was measured by thermal couples and recorded by a data acquisition system. The eddy current power losses in the permanent magnet were calculated using a KEC-developed spread sheet program with data collected during the test.

Figure 5.25 Current Waveform Applied to the Experimental Setup
Table 5.8 Results of Fourier Series Analysis

<table>
<thead>
<tr>
<th>Harmonic number</th>
<th>Current value (amp)</th>
</tr>
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<tbody>
<tr>
<td>0</td>
<td>125.0</td>
</tr>
<tr>
<td>1</td>
<td>71.6</td>
</tr>
<tr>
<td>3</td>
<td>8.0</td>
</tr>
<tr>
<td>5</td>
<td>2.9</td>
</tr>
<tr>
<td>7</td>
<td>1.5</td>
</tr>
</tbody>
</table>

5.5.1 Experimental Test Setup

All tests were designed to evaluate and compare the eddy current power losses generated in the two different permanent magnets (laminated and unlaminated). Rather than using the whole rotor with permanent magnets, only one permanent magnet was attached to the single stator. Figure 5.26 shows a single stator with a permanent magnet on the top. The air gap between the stator and the rotor permanent magnet was kept constant by plastic insulation G10 material and the rotor permanent magnet was secured to the stator to prevent possible rotation. Figure 5.27 shows a 2-D view of the test setup. In this figure, the current excitations are also shown.

The results of experimental measurements on the single stator are applicable to the whole motor because the eddy current power losses in the rotor permanent magnet generated by two stators are approximately two times higher than those generated by single stator. This was proven by FEA eddy current power loss comparison shown in Table 5.5.
Figure 5.26  Single Stator with a Permanent Magnet

Figure 5.27  2-D View of Test Setup
During the experimental tests, the triangular current waveforms were applied continuously to phases A and B of the stator; phase C was kept unloaded. This condition simulated the case of a stalled rotor.

5.5.2 Experimental Test Results

The test results demonstrate that high-frequency stator PWM current harmonics induce excessive eddy current power losses in the rotor permanent magnets. Additional rotor heat is created by hysteresis loss inside the permanent magnet when the rotor rotates between the slotted stators. This hysteresis loss is a result of permanent magnet motion relative to the stator slots and teeth. The flux density in the magnet area facing the slot is reduced as a result of the low magnetic permeability of the slot. The flux density in the area facing the tooth is increased because of the high permeability of the tooth. As the permanent magnets move in relation to the stator, their flux density changes. This changing flux generates the hysteresis loss in the magnets, which is not considered in this study.

The tests indicate that the eddy current loss in the rotor permanent magnet can be reduced four to five times by laminating the permanent magnet; however, the lamination process increases permanent magnet production cost nearly ten times. The results of the tests are summarized in the Table 5.9. This table shows trade-offs between laminated and unlaminated rotor permanent magnets. (A direct quantitative number comparison is not feasible because of the different methods using different permanent magnet shapes and dimensions.) The purpose of this study was to verify permanent magnet lamination reduces eddy current power loss.
Further study is recommended to determine lamination thickness optimization.

Table 5.9  Comparison Summary of Eddy Current Power Loss (@ f=10kHz)

<table>
<thead>
<tr>
<th>Method used</th>
<th>Unlaminated magnet</th>
<th>Laminated magnet</th>
<th>Ratio</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lumped model</td>
<td></td>
<td></td>
<td>15</td>
</tr>
<tr>
<td>FEA calculation</td>
<td>8.154 (watt)</td>
<td>1.216 (watt)</td>
<td>7</td>
</tr>
<tr>
<td>Test measured</td>
<td>113.74 (watt)</td>
<td>24.2 (watt)</td>
<td>4.7</td>
</tr>
</tbody>
</table>

Further reduction of the losses could be accomplished by using thinner permanent magnet laminations. The developed FEA model can be used to find optimum lamination thickness. The optimum thickness of the lamination depends on the manufacturability of the thinner lamination and costs. Laminating the permanent magnet reduces eddy current power losses in the permanent magnet, thereby eliminating heat (demagnetization) as a problem in the rotor. (Because each stator has an active cooling plate with liquid flow behind the back iron, the potential for a heated stator to heat the rotor and become a heating problem is eliminated.)

5.6 SUMMARY

In this chapter a comparison of eddy current power losses in laminated and unlaminated rotor permanent magnets was made. Permanent magnets were laminated to reduce eddy current power losses by the stator current excitation PWM waveform. Two methods were used to calculate eddy
current power losses in the rotor permanent magnet: (1) the lumped parameter magnetic circuit method and (2) the FEA method. The objective of this comparison was to find a trade-off between eddy current power loss reduction and the production cost of laminated permanent magnets.

Demagnetization of the permanent magnet caused by generated heat is one of many design considerations when a high-energy, rare-earth permanent magnet such as Nd B Fe is used. Of special interest is the eddy current loss in the rotor permanent magnet from the high-frequency PWM stator current harmonics. One method to decrease induced eddy current in the rotor permanent magnets is to laminate the permanent magnet block.

A simple lumped magnetic circuit equation was derived using Faraday's law and Ohm's law. The result showed almost 15-fold improvement when the permanent magnet was laminated.

An eddy current power loss comparison was conducted using 3-D FEA models. FEA models were built based on the convergence test and model simplification process. The interlaminar air gaps were modeled using the 3-D and 2-D element combination technique. FEA models were analyzed using different current excitations and applied frequencies. The results showed seven to ten times improvement when the permanent magnet was laminated.

Experimental tests were also conducted to compare eddy current power losses between laminated and unlaminated permanent magnets. The results revealed a four to five times improvement when the permanent magnet was laminated.
An analysis of lamination thickness versus eddy current power loss reduction is in progress to identify the optimum thickness of the lamination for a given operating PWM frequency. Further investigation on the derived lumped magnetic circuit equations and FEA models to minimize eddy current power losses is required.
CHAPTER 6

DESIGN METHODOLOGY

6.1 INTRODUCTION

This chapter addresses a design methodology using the integration of FEA and lumped parameter magnetic circuit equations, interactively and iteratively, and an alternative design to the prototype slotless, axial field, brushless, permanent magnet motor.

The research results documented in this chapter address the design methodology for next-generation, axial field, brushless, permanent magnet motor development for underwater applications. The information in this chapter is the result of independent assessment and trade-off studies based on this research, as well as literature surveys, and experimental tests conducted during the prototyping period.

6.2 DESIGN METHODOLOGY

This section focuses on a possible way to integrate the FEA and lumped magnetic circuit models to maximize their capacities for developing design methodology for a new brushless, permanent magnet, axial field motor configuration. FEA and the lumped magnetic circuit method are briefly discussed separately, and an integrated design method is also presented. The new methodology presented here is under development. It has been tested in

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several cases using a radial field configuration. Only qualitative results are discussed.

6.2.1 Finite Element Analysis

Several FEA software packages are available to solve 2-D, axi-symmetric, and 3-D electromagnetic problems. Though some can be run on a personal computer, these computation-intensive codes, especially 3-D problems, are best suited for more powerful computers. A comparison of several FEA packages showed that, when properly used, they provided reasonably accurate results when applied to 2-D and axi-symmetric dc magnetic circuits. The difference between FEA programs became apparent when comparing their ease of use, 3-D modeling, and capability to handle complex mathematical problems, such as eddy currents and transient responses of multi-phase motors.

Prior to FEA modeling, it is important that the magnetic circuit designer have a detailed understanding of the problem to be solved. Approximate predictions using lumped magnetic circuit analysis should be performed in preparation for any computer FEA modeling. FEA results are capable of yielding inaccurate, yet believable, results due to simple, but unnoticed errors on the part of the user. Even the most sophisticated FEA programs can not detect all potential human errors e.g., the use of diameters instead of radii in an axi-symmetric geometry and mixing of units.

It should also be noted that 3-D FEA processing time is approximately ten times more complex than that of 2-D or axi-symmetric analysis. It is possible to approximate a 3-D solution by carefully interpreting and combining several 2-D models. Such a multisegment modeling technique was used in this study to
calculate back emf, cogging torque, and inductance. When 3-D FEA is required, the magnetic circuit designer must have the following resources available: (1) a powerful computer, (2) color monitor with color printer, and (3) extensive time to interpret the results.

If the FEA model contains a significant variation in the geometry in the axial or circumferential direction, or the current distribution has a significant variation in the axial or circumferential direction, then the 3-D analysis must be performed. The term significant is dependent on the sensitivity of the field calculations to the current geometry. As previously indicated, there are a number of ways to account for a 3-D effect, such as fringing and end turns.

The objective of the design stage is to produce the most cost-effective working design. This implies iterative changing of the geometry and reviewing the results. Almost without exception, a 2-D analysis is faster than the corresponding 3-D analysis. However, 3-D analysis provides more detailed information.

6.2.2 Lumped Parameter Magnetic Analysis

The lumped parameter magnetic circuit equations are directly related to the motor geometry, material properties, and current-carrying winding characteristics. Using lumped magnetic equations, the motor output parameters can be calculated through closed-form algebraic expressions. Therefore, the changes in design parameters and variables can be rapidly assessed without using FEA modeling and the construction of costly prototypes. Well-developed lumped parameter magnetic model equations provide the preliminary motor design parameters and variables within a short-turn around, engineering time. After the preliminary design parameters have been examined and verified with
design specifications and objectives, further optimization procedures or refinements can be performed using FEA analysis. One significant advantage of lumped parameter magnetic circuit modeling is that it can be utilized for design trade-off studies such as: (1) the total number of poles versus efficiency, (2) the permanent magnet residual flux density versus stator back iron thickness, (3) the slot to tooth width ratio versus efficiency, and (4) the air-gap length versus output power of the motor.

There are several difficulties associated with using the lumped parameter magnetic circuit models in motor design. In order to derive an accurate lumped magnetic circuit model, the designer must know, or at least have a clear understanding of, the primary magnetic flux paths in the motor to be modeled. Designers not only need to know the flux paths, but also must be able to find the areas where the magnetic fringing effect occurs. In the region where the flux changes in two dimensions, more reluctance paths are required to model these fringing effects.

When accurate magnetic flux plots, parameter calculations, and performance estimates are required, using the FEA method is also necessary.

6.2.3 Integrated New Design Method

Subsection 6.2.3 presents information on the integration of the FEA method and a spread-sheet design program that uses lumped parameter magnetic equations as a future motor design tool. Until recently, the spread-sheet design program and FEA method were used separately. The spread-sheet design program has been used as a point design tool and FEA as a verification or modification of the design. As the power of computers has increased, the integration of the FEA with lumped parameter magnetic equations has become
feasible. The development of a full integration program must be carefully planned because many designers do not understand the intricacies of the FEA method or discretization. The FEA part of the design procedure should be totally transparent to the user so that the only requirement is to fill in a design sheet based on lumped magnetic calculation results.

Previously, electric motors were designed based on analytically derived equations or lumped parameter magnetic model equations, which have been modified over the years by experimentation, empirical factors derived from previous machine performance, and experience. The most common design tool currently being used by motor design engineers is a spread-sheet computer program with magnetic design equations from the lumped parameter magnetic circuit model. Existing lumped magnetic circuit equations can be easily transferred to spread-sheet programs such as EXCEL, LOTUS 123, and MathCAD. More accurate and detailed magnetic field analysis using lumped magnetic circuit model requires additional calculations and models. However, as the need arises to optimize designs, more sophisticated design tools such as FEA, finite difference method (FDM), and boundary element method (BEM) are required. Whatever calculation method is used, it should be capable of accurately modeling some or all of (1) complex geometry, (2) induced eddy currents, (3) magnetic saturation, and (4) nonlinear, nonhomogeneous regions.

This study is pursuing the integration of the FEA with a developed generic FEA modeling capability program and the lumped parameter magnetic circuit equations transferred to a spread-sheet program to design and analyze the next-generation prototype axial field permanent magnet motor. An integrated design tool flow chart is shown in Figure 6.1.
Figure 6.1 Design Process Flow Chart
A design can be started by given design objectives, specifications, and constraints, and when they have become well-defined, motor configuration can be selected based on design requirements, trade-off studies, and previous motor design history. After selection of the motor configuration, motor design material survey must be conducted. Materials surveyed may include (1) stator lamination material, (2) rotor permanent magnet material, (3) rotor matrix material, (4) current-carrying wires, and (5) motor housing material.

The new integrated design methodology shown in Figure 6.1 is described in the following paragraphs. The first step in the motor design using the suggested methodology is to determine the motor design parameters and variables using the lumped magnetic circuit equations in conjunction with the spread-sheet program.

The inputs to the lumped magnetic model are:

1. Output power requirement and rated rotor speed,
2. Peak back emf voltage and maximum bus voltage,
3. Number of poles and phases,
4. Slots/phase,
5. Copper packing factor,
6. Operating temperature,
7. Outside/inside stator radius,
8. Axial length,
9. Air gap length,
10. Magnet length, and
11. Magnet remanence and permeability.

The outputs from the lumped magnet model are:
(1) Back iron thickness,
(2) Number of slots,
(3) Slot width and depth,
(4) Air gap flux density,
(5) Peak back emf voltage,
(6) Core loss,
(7) Wire gage,
(8) Total resistance in the slot,
(9) Output torque,
(10) Number of slots/pole/phase,
(11) Fundamental electrical frequency, and
(12) Permeance coefficient.

When all the design parameters have been calculated using the lumped parameter magnetic model equations, the next step is to transfer all the calculated input data to the FEA generic modeling program. (This input data can be transferred automatically dependent upon computer capability.) Using all the design parameters as input data, FEA models can be built and analyzed using the generic modeling capability program with minimum user-interaction with the computer.

The input data required for the generic FEA modeling program includes:
(1) Outer diameter and inner diameter of the stator and rotor,
(2) Total number of slots and slot width,
(3) Number of phases,
(4) Material properties,
(5) Stator back iron thickness,
(6) Air-gap length,
(7) Rotor thickness (permanent magnet length),
(8) Excitations, and
(9) Boundary conditions.

The results of FEA analysis, such as back emf waveform, one-phase-on
 torque waveform, cogging torque, and inductance value are compared with
design specifications and objectives. This is done at design evaluation (stage I).
At this stage, the motor efficiency is calculated and examined by the engineer
and compared with the design goal. If any discrepancies in design parameters or
variables are found, the design loop is returned to the lumped parameter
magnetic calculation stage, where the lumped parameter magnetic equations
must be reexamined and values recalculated. The recalculated values are again
input to the FEA modeling and analysis process. After completion of stage I, the
design optimization process takes place. In the optimization process, refinement
and reshaping of the geometry, such as slot/tooth width ratio, air-gap length,
stator back-iron thickness, and rotor thickness (permanent magnet thickness)
must be performed. Using the new parameters and geometry, new FEA models
can be built and analyzed by the FEA generic modeling capability program.

In the second design evaluation (stage II), all the recalculated design
parameters and characteristics are compared with design specifications and
design objectives. At this stage, manufacturing problems and design materials
must be examined.

Prototyping can commence after the second design evaluation stage. The
third design evaluation (stage III) is performed after the first prototype motor is
developed and the final design evaluation is performed using all the measured
values and parameters from prototype motor. If the prototype motor does not
meet specifications and objectives, the design must be returned to the optimization stage.

6.3 ALTERNATIVE DESIGN

This section describes the slotless, axial field, brushless, permanent magnet motor an alternative to the slotted, axial field, brushless, permanent magnet motor configuration. Figures 6.2 and 6.3 depict a simplified slotted stator and slotless stator, respectively. The same rotor shown in Figure 6.4 can be used in both slotted and slotless motor configurations. The purpose of these figures is to present a graphical comparison of the two types of motors. It is clear from these figures that the primary differences between the two configurations are slots, teeth, and stator thickness.

6.3.1 Slotless Axial Field Permanent Magnet Motor

As seen in Figure 6.3, the slotless motor stator has no slots and teeth. The slotless motor configuration is possible because of two technological developments: (1) high-energy density permanent magnet materials and (2) multistranded current-carrying wires. There are two reasons why these two developments are necessary for the slotless motor configuration. First, the effective magnetic air gap is large, compared to that of slotted motors. Therefore, high-energy density permanent magnet material is required to compensate for this increased air-gap length, so that the air-gap flux density is the same as it is for the slotted stator. Second, in the slotted stator, all winding coils are placed inside slots and surrounded by stator teeth. In the slotless stator all current-carrying winding coils are exposed throughout the stator surface; therefore, there are more eddy current losses in the wires. The increased eddy current loss can be reduced by using multistranded wires in the stator.
Figure 6.2 Slotted Stator

Figure 6.3 Slotless Stator

Figure 6.4 Axial Field Motor Rotor
One major advantage in the slotless motor configuration is the permanent elimination of cogging torque. In this study, several cogging torque reduction methods were applied to reduce cogging noise and vibration. The best solution for reducing cogging torque is elimination of the cogging force source, which is slots and teeth. The source of the cogging force is a variation of the permeance or reluctance between the stator teeth as "seen" by the rotor permanent magnet. If the stator teeth are removed, the permanence between the stator and rotor does not vary as the rotor rotates. Therefore, removal of the stator slots and teeth results in smooth and quiet operation for all speed ranges [43].

Another advantage of the slotless motor configuration is its ability to distribute current-carrying wires throughout the stator surface. Because there are no slots and teeth on the stator (the slotless motor), the current-carrying windings must then be fixed to the stator back iron instead of slots in the slotted stator. With a fixed axial length of the motor, the total area for the phase current-carrying wire is increased. Because of the increased total area for the windings, more wire can be placed, which reduces the total phase resistance of the winding. Because of the higher reluctance in the slotless motor, the phase inductance value is smaller than that of the slotted stator. The inverter circuit can respond for fast acceleration and deceleration of the rotor by easily controlling the current waveform. Since there are no slots and teeth, the winding can be distributed more evenly throughout the stator, which results in a more sinusoidal back emf waveform and an improved torque waveform [44,45].

The major disadvantage of slotless motors is the large amount of excitation energy required to produce the same air-gap flux because of a longer air gap compared to the slotted motors. Thus, more permanent magnet material
or a stronger permanent magnet is required in the rotor or higher phase current, or both are required in stator windings. Another disadvantage is that the exposed copper winding is subject to eddy current losses, which are induced by the inverter PWM field and rotation of the permanent magnet in the rotor. The result is more drag, heat, and lower efficiency. This loss can be significantly reduced by using multistranded wires and thinner stator core lamination with higher permeable material.

6.3.2 Comparison of Slotless and Slotted Stator

The comparison of slotted stators and slotless stators shown in Table 6.1 was made using the following assumptions [43,44,45,46,47]:

1. Same materials (permanent magnet and stator lamination),
2. Same motor dimensions (axial length, diameter, and air-gap length),
3. Same motor speed, and
4. Same number of winding turns with the same copper wire gage.

In the slotted stator motor, the cogging torque is high when high-energy, permanent magnet materials such as Sm Co and Nd B Fe are used. This high cogging torque is a major design consideration for high-performance and quiet operation applications. Because there are no slots and teeth in the slotless stator motor, there is no cogging torque. As a result, the air-gap reluctance between the rotor permanent magnet and stator lamination material is constant; thus, there is no cogging force.

There are two reasons why the slotted-stator configuration has a higher back emf voltage (open circuit voltage) than that of a slotless-stator configuration. First, the magnetic effective air-gap length is shorter than on the slotless motor, and more magnetic flux is enclosed by stator winding coils, which
Table 6.1 Qualitative Comparison of Slotted and Slotless Motor

<table>
<thead>
<tr>
<th>Motor design considerations</th>
<th>Slotted stator motor</th>
<th>Slotless stator motor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cogging torque</td>
<td>high</td>
<td>eliminated</td>
</tr>
<tr>
<td>Back emf voltage</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Copper loss (I²R)</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Eddy current/ hysteresis loss (stator teeth)</td>
<td>high</td>
<td>eliminated</td>
</tr>
<tr>
<td>Eddy current loss (wires)</td>
<td>low</td>
<td>high</td>
</tr>
<tr>
<td>Hysteresis loss (PM)</td>
<td>high</td>
<td>eliminated</td>
</tr>
<tr>
<td>PM operating point</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Phase inductance</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Winding placement</td>
<td>easy</td>
<td>difficult</td>
</tr>
<tr>
<td>Torque</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Production cost (stator)</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Magnetic flux leakage</td>
<td>low</td>
<td>high</td>
</tr>
</tbody>
</table>
produces higher back emf voltages. Secondly, the rotor permanent magnet flux is concentrated through the stator teeth. This higher flux concentration in each phase winding produces higher back emf voltages. In the slotless-stator configuration, the magnetic flux leakage is higher as compared to the slotted stator; thus, lower back emf voltages are produced in the same set of stator winding coils. In an application where a limited buss voltage from the energy source (e.g., battery) is used, a lower back emf is beneficial to control the output torque. In most brushless, permanent magnet motors, it is desirable to have higher back emf voltages because they provide a high torque constant, which results in high output torque.

The stator phase copper loss ($I^2R$) can be low in a slotless stator with a given motor axial length because there is more cross-sectional area to place copper wires, the results of which could be lower phase resistance and reduced power loss. In a slotted stator, the amount of copper winding used is limited by the slot cross-sectional area.

In a slotted stator, the eddy current and hysteresis losses are higher than that of slotless stator with the losses concentrated in the stator teeth. Because the teeth are made of a magnetic material, they are affected by the rotor permanent magnet field and stator excitation PWM field; therefore, eddy current and hysteresis losses are primarily located in the stator teeth. In the slotless stator, because there are no teeth, there are no eddy current and hysteresis losses.

In the slotted stator, the eddy current loss in the stator copper windings is less than that of the slotless configuration. Since the slotless stator windings are exposed to the rotor permanent magnet flux and the stator excitation field flux, the eddy current loss is higher than that of the slotted stator. This higher
winding eddy current can be significantly reduced by using multistranded wires instead of solid wires.

The hysteresis loss produced in the rotor permanent magnet material is the result of changing air-gap permeance (air-gap reluctance) when the rotor permanent magnet is exposed to the stator slots and teeth. This time-varying permeance creates small hysteresis loops inside the permanent magnet material, which produces hysteresis losses. Because the slotless stator does not have slots and teeth, there are no hysteresis losses in the rotor permanent magnet material. There is, however, eddy current power loss in the permanent magnet by stator winding current excitation.

In the slotted stator, the effective magnetic air gap is short compared to that of the slotless stator motor. This air-gap length with the rotor permanent magnet length determines the magnetic operating point in the demagnetization B/H curve. In the slotted permanent magnet motor, the operating point is higher because of the smaller air-gap length on the demagnetization B/H curve compared to the air-gap length of the slotless motor. Therefore, the operating point is located on the linear part of the B/H curve, allowing use of thinner (in axial direction) permanent magnets for the same output power. In addition, there is less risk of permanent magnet demagnetization. Since there are no stator teeth in the slotless stator, the rotor axial length (permanent magnet length) can be increased, but remains limited by the total motor axial length. This increased permanent magnet length provides higher rotor MMF in the air gap so that the magnitude of stator excitation can be reduced for the same power output.

A slotted stator can improve magnetic flux focusing because the stator teeth have a high permeability $\mu$. Due to the flux focusing, the magnetic pole to
stator phase winding leakage flux is less than what occurs in the slotless stator. Therefore, not only is the back emf voltage higher in the slotted stator motor, but the phase winding inductance is also higher and thus permits inverter commutation despite higher reactance. When a motor has a rapid acceleration and deceleration as a design requirement, then the slotless stator is an optimum candidate. A lower phase inductance value provides better motor stability and easier controller design.

In a slotted stator motor, the stator phase current windings are physically balanced by stator slots and teeth. A disadvantage of the slotted stator, however, is that the slots provide a discrete means of winding distribution compared to the continuous distribution in the slotless stator. Optimum winding distribution can best be achieved in the slotless motor configuration.

Production cost is less for the slotless stator compared to the cost of producing a slotted stator. In the slotless motor, slot machining is not required; a simple roll of steel ribbon can be used for stator lamination. The slotted stator requires a slotting process and an annealing process after the slotting process to restore the original magnetic properties in the lamination material. Since the cost of the stator in motor production is a significant factor, use of the slotless stator can reduce total production cost.

Magnetic flux leakage is high in the slotless motor because of the distributed stator winding and longer air gap between the stator back irons.

In summary, qualitative analysis between the slotted and slotless motor shows that the slotless motor design is better for quiet operation design for use in the future axial field, high-power density, high-efficiency, and quiet operation motor. This slotless, axial field, brushless, permanent magnet motor is a viable
configuration for the following reasons. First, development of rare-earth permanent magnet material energy density is rapidly increasing toward its theoretical limitation. Concurrently, the price of this permanent magnet material is decreasing. Secondly, multistranded wire technology can support the slotless motor with significantly reduced ac conduction loss and reduced eddy current and hysteresis power losses in the stator winding. The slotless motor configuration is a strong candidate for variable-speed, underwater applications where low noise/vibration operation is a major design objective. Equivalent power density, compared to that of the slotted stator motor, may be obtained by using the slotless motor configuration. However, new winding techniques must be developed to place the slotless stator windings. Further quantitative trade-off studies between the slotted stator and the slotless stator must be conducted before commencing the next-generation axial field motor design using the slotless stator configuration.

6.4 SUMMARY

In this chapter, design methodology using the integration of FEA and lumped magnetic circuit equations, interactively and iteratively, and an alternative design were presented for the brushless, axial field, permanent magnet dual air-gap motor configuration. Advantages and disadvantages of the FEA method and lumped magnetic circuit analysis were discussed.

Finally, an alternative slotless, axial field motor design was discussed. Based on the tests conducted in this study, cogging torque was a significant noise/vibration source. A major advantage of the slotless motor design is quiet motor operation throughout a wide speed range because there is no cogging torque. An additional benefit of high-speed operation is the elimination of eddy
current and hysteresis losses within stator teeth magnetic material. The biggest disadvantage of the slotless motor design is the slightly reduced output power, a minor disadvantage that does not outweigh the significant advantage of quiet motor operation. Therefore, the next generation of axial field motors will be slotless, axial field motors.
CHAPTER 7

CONCLUSIONS AND RECOMMENDATIONS

7.1 Overview and Methodology

The objective of this study was to develop a high-power density, high-efficiency, and low-noise/vibration motor for underwater propulsion applications. Because the design goals for such a motor are more complex and rigorous than those of previously designed motors, more advanced and refined analysis tools, techniques, and design methods had to be used and, in some cases, developed.

This research entailed a comprehensive analysis of an axial field, brushless, permanent magnet motor. Critical motor components and design issues of an axial field motor were addressed, and a prototype motor was discussed in detail. Motor parameters and performance characteristics were analyzed using the FEA method and the lumped parameter magnetic circuit method. Additionally, a unique generic FEA modeling capability program was developed to mechanize the time-consuming FEA process and reduce user interaction with the computer. Calculated motor design parameters and performance characteristics were compared with the actual prototype motor measurements.

The scope of this research is the development and verification of detailed FEA models and lumped parameter magnetic circuit models of an axial field, brushless, permanent magnet motor to make design improvements.
Accordingly, motor design parameters, cogging torque and axial force variation, and eddy current power losses in the rotor permanent magnet are investigated. This research is the first thorough quantification of dual air-gap, axial field, permanent magnet motor parameters and performance characteristics using FEA and lumped parameter magnetic circuit analysis.

7.2 Specific Findings

A significant portion of this dissertation deals with the calculation of several important motor design parameters and performance characteristics including flux distribution, flux linkages, back emf waveform, inductance, and energized torque waveform. The purpose of this part of the research was to model the electromagnetic design, improve the existing prototype motor design, integrate with the controller design, and develop a design tool for future motor designs. It was found that when using the 2-D FEA method in conjunction with the generic FEA modeling capability program, useful design and performance analysis can be obtained with minimum user interaction. Comparing FEA-calculated back emf waveform and stator phase inductance values with measured prototype experimental data showed a close correlation between the calculations and measurements (a less than 10 percent error). This close correlation is attributed to the fact that (1) FEA models were built based on the optimum total number of elements by convergence test results and the coercive force permanent magnet modeling technique and that (2) FEA models were analyzed using the multisegment modeling technique.

Additionally, the lumped magnetic circuit models were developed using FEA results. Developed lumped parameter magnetic models were used to estimate the back emf waveform, phase inductance value, and energized torque
waveform. The calculation results showed a 17-percent error on the back emf waveform compared to the prototype motor measured waveform and about a 12-percent error on phase inductance calculation. These error rates demonstrate an extremely close correlation between the lumped parameter magnetic model and the prototype measured values.

Another significant part of this research was analyzing cogging torque, axial force, and the effects of rotor misalignment, all of which were calculated using 2-D quasi-static, nonlinear, multisegment FEA models. The cogging torque, the axial force variation, and the rotor misalignment studies required a series of FEA models and analyses, which were a function of rotor position. The study confirmed that the generic FEA modeling capability package was an essential modeling tool. This modeling program minimized user interface with the computer and provided results and plots of all calculations automatically based on input data.

Two cogging torque reduction methods (staggering stators and skewing permanent magnets) were studied. Staggered and unstaggered stators with skewed and unskewed permanent magnet combinations were analyzed. By skewing the rotor permanent magnets and staggering the stators, the cogging torque was reduced to 1 percent of that for a motor with no skewing and staggering, representing a significant reduction. The permanent magnet skewing effect on the cogging torque was analyzed using a developed SIMULAB program. The most significant factor in reducing cogging torque was stator staggering. However, staggering the stators also increased the axial force variations by almost nine times. The staggered stator effect on output torque was also examined. It was found that staggering the stator has no impact on output torque. In a motor having a light rotor structure, such stresses by axial force
variations cause mechanical vibration, noise, and fatigue. The effect of stator staggering and rotor misalignment effects were also examined. A trade-off between cogging torque reduction and axial force variation was determined. A generic modeling capability program package, with an optimum number of elements and an improved permanent magnet modeling technique (current density method), were used to perform an accurate cogging torque calculation.

As a result of this research, it was found that misalignment of the rotor increased axial force variations and counteracted cogging torque reduction achieved by stator staggering. However, it was also found that increased axial force variations can be controlled by increasing stator back iron thickness. However, increased stator back iron reduces the power density of the motor, which is not acceptable for the high-power density motor design.

Based on the findings of this study, the new (unit 2) prototype motor was redesigned as follows: (1) the rotor permanent magnets were skewed to reduce cogging torque, (2) the stator was unstaggered to minimize the axial force variation, (3) the air gap was increased, and (4) a new bearing was selected based on the axial force calculation result.

Comparing eddy current power losses in laminated and unlaminated rotor permanent magnets was another major focus in this research. Two methods were used to calculate eddy current power losses in the rotor permanent magnet: (1) lumped magnetic circuit model and (2) FEA method. The objective of this comparison study was to find a trade-off between eddy current power loss reduction and the production cost of laminated permanent magnets.

A simple lumped parameter magnetic circuit equation was derived using Faraday's law and Ohm's law. The result showed a 15-fold eddy current power
loss reduction when the permanent magnet was laminated. FEA 3-D models were also built based on a convergence test and model simplification process. The interlaminar air gap between permanent magnet laminations was modeled using the 3-D and 2-D element combination technique. The FEA models were used to analyze different current excitations and applied frequencies. The results showed seven to ten times eddy current power loss reduction when the permanent magnet was laminated. Experimental tests were conducted to find eddy current power loss differences between the laminated and unlaminated permanent magnets. The test result showed a five-fold improvement in eddy current power loss attributed to permanent magnet lamination.

Analysis of the laminated permanent magnet and unlaminated permanent magnet using the lumped parameter electromagnetic model and the FEA model compares favorably with experimental measurements. It was determined that for a specific frequency and lamination thickness, the power loss can be reduced almost five to ten times that of the unlaminated permanent magnet case. The eddy current power loss results that were obtained could possibly be used to predict motor efficiency and cooling requirement. It is important to consider the results of this study when the permanent magnet length is considerably long because there are large eddy current losses that can degrade the magnetic field and possibly demagnetize the permanent magnets. Investigation of lamination thickness versus eddy current power loss to identify the optimum thickness of the lamination for a given operating PWM frequency is in progress. Further investigation can be conducted by using the derived analytical expressions and FEA models for induced eddy current power loss minimization.

Part of this dissertation concentrates on the design considerations for an axial field, brushless, permanent magnet motor: (1) motor configurations and
their characteristics, (2) electromagnetic design concerns, (3) material selection and availability, (4) thermal management, and (5) noise and vibration control.

Available motor design materials and selection criteria were studied to maximize the usage of the material and minimize power losses. An extensive study of material selection criteria for each motor component was conducted. Available stator magnetic core materials and their characteristics were compared and discussed. For the high-power density motor design, vanadium permendur magnetic material was found to be the best choice and was used in the prototype motor. A wide range of current-carrying stator winding conductors were examined. In this study, the solid wire and multistranded wire types were compared experimentally, and the result showed that the multistranded wire is the best choice for the stator designed motor with deep slots. In the deep-slot stator design, the dominant stator winding self-inductance is the slot leakage inductance; thus, the multistranded wire was used in prototype motor to reduce ac resistance conduction losses.

Elimination of waste heat generated inside the motor by conduction losses in current-carrying windings; eddy current/hysteresis losses in the stator magnetic material; eddy current in the rotor permanent magnet and rotor structure; switching losses in the power inverter/converter; and controller circuit loss has been a major motor design concern, especially in the high-power density motors. Different thermal management methods were presented. The most effective way to remove wasted heat in the prototype motor configuration was to add a heat sink with fluid flowing tubes on the back of each stator.

Axial field permanent magnet motor design tools and methodology were presented. Advantages and disadvantages of the FEA method and lumped
magnetic circuit analysis were discussed. The integration of FEA and lumped parameter magnetic models were presented, as well as a multi-iterative new design flow chart.

An alternative slotless, axial field motor design was discussed. Based on the tests conducted in this study, cogging torque was a significant noise/vibration source. A major advantage of the slotless motor design is quiet motor operation throughout a wide speed range due to cogging torque elimination. An additional benefit of high-speed operation is elimination of eddy current and hysteresis losses within stator teeth magnetic material. Therefore, the next generation of axial field motors will be slotless axial field motors.

7.3 Conclusions

For the first time (based on extensive library literature search), in this research, a series of snapshot finite element field solutions using multisegment modeling technique throughout a complete electrical cycle of the rotor permanent magnet rotation and the stator current excitation have been used to account for the rotor movement with different excitations. Several critical FEA modeling techniques were examined, such as the FEA model convergence test and permanent magnet modeling techniques (current density method). Developed FEA model results were compared and validated by experimental measurements. It was found that the FEA model convergence test is an essential modeling step, where FEA models included air gaps and permanent magnets. It was also found that the permanent magnet modeling technique is critical when the FEA model includes permanent magnet excitation.

To improve the FEA method, a generic modeling capability program package using FORTRAN programs was developed. This program permitted the
building of 2-D and 3-D models in conjunction with VAX/VMS command procedures and MACRO programs. The generic modeling capability was developed to mechanize mesh generation and analysis, thus providing the designer with a valuable design and analysis tool.

The main contribution of this dissertation is the development and verification of detailed FEA models and lumped parameter magnetic models that can calculate back electromotive force waveforms, inductance, cogging torque, energized torque waveforms, and eddy current power losses for the axial-field, dual-air gap, permanent magnet motor. The application of these models allows accurate prediction of the motor parameters and the dynamic behavior of such motors without the construction of costly prototypes.

7.4 Recommendations for Future Study

The author considers the results of this research compelling evidence that the advanced analysis tools, techniques, and design methods required to develop the high-power density, high-efficiency, and low-noise/vibration motor are attainable. The results of this investigation are substantive and reveal the opportunity to successfully explore questions beyond the scope of this research project.

Specifically, this author recognizes merit in and recommends this additional research be conducted:

- Extend the application of the generic FEA modeling capability program to determine the energized torque ripple using all three stator phase excitations.
• Modify the generic FEA modeling capability program to use radial field motor configurations.
• Optimize rotor permanent-magnet skewing angle studies using the FEA model and SIMULAB program.
• Conduct a structural FEA modeling using FEA magnetic field calculation results to predict noise/vibration.
• Develop an analytical expression to predict cogging torque using Maxwell's field equations or method of images.
• Study eddy current effect on high-speed motor operation.
• Extend the FEA model to include the rotor rotation dynamic eddy current effect.
• Conduct optimum permanent magnet lamination thickness using developed models.
• Modify the lumped magnetic circuit model to include rotor permanent magnet pole leakage and nonsinusoidal field assumption.
• Complete an integration study of FEA modeling and lumped magnetic circuit modeling as a design package.
• Conduct a quantitative trade-off study between the slotted stator and the slotless stator, axial field motor configuration for the high-power density motor application.

In short, this research has served science and technology in two ways. Detailed axial field, brushless, permanent magnet motor FEA models and lumped parameter magnetic models that can calculate back emf waveforms, inductance, cogging torque, energized torque waveforms, and eddy current power losses have been developed and verified; and, of equal import, it has opened the door for even more scientific advancement.
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