DESIGNING, PROTOTYPING AND INVESTIGATION OF ADVANCED FIVE-PHASE AXIAL FLUX SRM FOR ELECTRIFIED VEHICLE APPLICATION

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DESIGNING, PROTOTYPING AND INVESTIGATION
OF ADVANCED FIVE-PHASE AXIAL FLUX SRM
FOR ELECTRIFIED VEHICLE APPLICATION

By

Anas Labak

A Dissertation
Submitted to the Faculty of Graduate Studies
Through the Department of Electrical and Computer Engineering
in Partial Fulfillment of the Requirements for the
Degree of Doctor of Philosophy
at the University of Windsor

Windsor, Ontario, Canada

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ABSTRACT

The recent interest by the automotive industry in finding an electric motor that has high power density, rugged construction with high temperature adaptability, and fault tolerance ability has revived the research on switched reluctance motors (SRMs). Due to the aforementioned characteristics, SRM is considered an attractive alternative to traditional electric motors in electric vehicle (EV) applications. However, there are challenges such as torque ripple and acoustic noise that need to be addressed. The aim of this dissertation is to provide solutions through fundamental design improvements in order to develop a viable SRM-based propulsion system.

In order to analyze and quantify the major well-known issues of SRM, firstly, a case study based on finite element analysis (FEA) is performed on two types of SRM designs, a conventional SRM and a new design of in-wheel outer-rotor SRM. Despite the improvement in torque ripple provided by the new design, a comparative performance analysis suggests that further design modification is necessary in order to mitigate the acoustic noise and vibration issue in the machine while maintaining the achieved improvements. Consequently, an axial-flux configuration of SRM is proposed and its design and analysis is presented. Detailed procedure of deriving the output power equation as a function of motor dimensions and parameters are provided. A novel modified phase winding design method is thoroughly explained, and the inductance determination by different methods is verified experimentally. A 3-D FEA unveils excessive end core and radial flux fringing effects, subsequently, an exclusive pole-shape design is proposed. The dynamic operation of the motor is analyzed through 3-D FEA motion model. The prototype development process and static testing are demonstrated.
Experimental investigations have revealed issue of low inductance ratio due to higher leakage flux in this type of machine. Subsequently, three different novel approaches based on segmented grain-oriented steel core and magnetic shielding are proposed to mitigate the leakage flux, and then tested individually using 3-D FEA. In addition, comparative performance analysis of the original machine model and the machine with each of these approaches is carried out and improvement in the inductance ratio is observed. Overall, the proposed ASRM, with all aforementioned design improvements is found to satisfactorily address the major challenges.
DEDICATION

To My Family
ACKNOWLEDGEMENT

I am thankful to God Almighty for giving me the opportunity to pursue the doctoral program and the strength and patience to successfully complete it. I am grateful to my wonderful parents whose love, encouragement and sacrifice has made me what I am today. My heartfelt thanks go to my wife, Aya, and to my adorable children, Tasneem, Mohammad, and Malik for tolerating me and contributing toward my doctoral degree in all ways possible.

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NOMENCLATURE

Generally symbols have been defined locally. The list of principle symbols is given below.

\( A_c \) Area of coil cross-section

\( A_s \) Specific electric loading

\( A_{sp} \) Area of stator pole

\( A_w \) Area of wire cross-section

\( B \) Magnetic loading, magnetic flux density

\( B_c \) Magnetic flux density in the core

\( B_g \) Magnetic flux density in the air gap

\( B_s \) Magnetic flux density at which saturation begins

\( D \) Bore diameter

\( e \) Back electromotive force

\( F \) Per phase magnetomotive force

\( F_c \) MMF drops in c-core stator

\( F_g \) MMF drops in the air gap

\( F_r \) MMF drops in the rotor cube

\( f_s \) One phase excitation switching frequency

\( f_{max} \) Maximum frequency for 100% skin depth

\( g \) Air gap length

\( H_c \) Magnetic coercivity

\( h_c \) Core height

\( I \) Phase current
\( i \) Instantaneous phase current

\( I_{max} \) Maximum current supplied by the SRM drive

\( I_r \) Rated current

\( i_s \) Current at which saturation begins

\( K_{sf} \) Fill factor

\( K_u \) Constant relates the inductance ratio

\( k_d \) Duty cycle constant

\( k_e \) Efficiency constant

\( L \) Winding inductance

\( L_a \) Aligned inductance

\( L_u^a \) Aligned unsaturated inductance

\( L_u \) Unaligned inductance

\( l_c \) Coil length

\( l_{core} \) Length of flux path through the core

\( l_r \) Rotor cube line along the radial axis

\( l_t \) Rotor cube line along the tangential axis

\( mmf \) Magnetomotive force

\( m \) Number of phases conducting simultaneously

\( N \) Number of turns per coil

\( N_{ph} \) Number of SRM phases

\( N_r \) Rotor speed (rpm)

\( N_{rep} \) Number of phase repetition

\( N_{st} \) Number of strands per coil
\( P \) \hspace{1cm} \text{Output power}

\( P_r \) \hspace{1cm} \text{Number of rotor poles}

\( P_s \) \hspace{1cm} \text{Number of stator poles}

\( R, \ R_{ph} \) \hspace{1cm} \text{Phase resistance}

\( r \) \hspace{1cm} \text{Radius, Rotor cube to shaft distance}

\( r_{ir} \) \hspace{1cm} \text{Rotor inner radius}

\( r_{or} \) \hspace{1cm} \text{Rotor outer radius}

\( r_{os} \) \hspace{1cm} \text{Stator outer radius}

\( T \) \hspace{1cm} \text{Torque}

\( T_{av} \) \hspace{1cm} \text{Average torque}

\( T_{\text{Instmax}} \) \hspace{1cm} \text{Maximum instantaneous torque}

\( T_{\text{Instmin}} \) \hspace{1cm} \text{Minimum instantaneous torque}

\( TR \) \hspace{1cm} \text{Torque ripple}

\( t_r \) \hspace{1cm} \text{Rotor thickness}

\( t_c \) \hspace{1cm} \text{Core thickness}

\( T_s \) \hspace{1cm} \text{Per phase switching period}

\( V \) \hspace{1cm} \text{Single phase terminal voltage}

\( V_s \) \hspace{1cm} \text{Voltage of the source}

\( W_c \) \hspace{1cm} \text{Core width}

\( W_{co} \) \hspace{1cm} \text{Coenergy}

\( W_m \) \hspace{1cm} \text{Mechanical work}

\( \alpha \) \hspace{1cm} \text{Ratio of aligned to unaligned inductances}

\( \beta_r \) \hspace{1cm} \text{Rotor pole arc}

\( \beta_s \) \hspace{1cm} \text{Stator pole arc}
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
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<tbody>
<tr>
<td>$\varepsilon$</td>
<td>Step angle</td>
</tr>
<tr>
<td>$\mathcal{R}$</td>
<td>Magnetic reluctance</td>
</tr>
<tr>
<td>$\mathcal{R}_{c-back}$</td>
<td>Reluctance of the back core of a c-core</td>
</tr>
<tr>
<td>$\mathcal{R}_p$</td>
<td>Reluctance of a pole of single c-core</td>
</tr>
<tr>
<td>$\mathcal{R}_c$</td>
<td>Reluctance of the rotor cube core</td>
</tr>
<tr>
<td>$\mathcal{R}_f$</td>
<td>The air-gap fringing reluctance</td>
</tr>
<tr>
<td>$\mathcal{R}_{pc}$</td>
<td>Pole-to-cube reluctance</td>
</tr>
<tr>
<td>$\mathcal{R}_l$</td>
<td>Leakage flux reluctance</td>
</tr>
<tr>
<td>$\rho$</td>
<td>Electrical resistivity of conductor material</td>
</tr>
<tr>
<td>$\tau_r$</td>
<td>Rotor pole pitch</td>
</tr>
<tr>
<td>$\tau_s$</td>
<td>Stator pole pitch</td>
</tr>
<tr>
<td>$\tau_{spp}$</td>
<td>Stator pair-pair pitch</td>
</tr>
<tr>
<td>$\mu$</td>
<td>Magnetic permeability of the core material</td>
</tr>
<tr>
<td>$\mu_0$</td>
<td>Magnetic permeability of the free space</td>
</tr>
<tr>
<td>$\phi$</td>
<td>Magnetic flux</td>
</tr>
<tr>
<td>$\phi_r$</td>
<td>Remnant flux</td>
</tr>
<tr>
<td>$\theta$</td>
<td>Angular rotor position</td>
</tr>
<tr>
<td>$\theta_i$</td>
<td>Current conduction angle</td>
</tr>
<tr>
<td>$\omega_m$</td>
<td>Motor angular speed</td>
</tr>
<tr>
<td>$\omega_b$</td>
<td>Motor angular base speed</td>
</tr>
<tr>
<td>$\psi$</td>
<td>Single phase flux linkage</td>
</tr>
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Chapter 1

1. INTRODUCTION

1.1 Overview

The advancement of electric vehicle (EV) and hybrid electric vehicle (HEV) technology has been significant in the past decade owing to improvements in motor design, power electronics, and control and battery technologies. EVs and, to an extent, HEVs are alternatives to gasoline and diesel automobiles, which are known to cause pollution, including particulate matter and other smog-forming emissions. Automobiles contribute to more than half of the CO and NOx, and almost a quarter of the hydrocarbons emitted into our air [1]. This air pollution carries substantial risks for human health and disturbs the ecosystem balance. Through clean vehicle and fuel technologies, air pollution and its subsequent effects can be reduced significantly, while cutting projected oil use in half within the next 20 years. Switching from gasoline cars to plug-in hybrids that can be recharged by existing utility grids would lead to 42 percent national average reduction in CO\textsubscript{2} emissions by 2020, even though the majority of the utility grid uses fossil fuels [2], suggesting that usage of EV, which are zero emission vehicles, will reduce the emissions even further.

Growing transportation costs and upcoming regulations have increased the focus on vehicle fuel efficiency and emissions control, emphasizing a need for more aggressive research addressing the complex interactions of advanced powertrain technologies.
Electric traction systems account for 16% of overall losses in all-electric vehicles over an Urban Dynamometer Driving Schedule (UDDS) drive cycle, which is used for testing city driving conditions for light-duty vehicles. Most of these losses arise from the traction motor, according to an ORNL report [3]. Therefore, improvements are focused on motor design and motor materials that have the potential to significantly improve motor efficiency and performance considering constraints such as size, weight, and cost.

Research objectives in the field of automobiles are consistent with the world’s energy initiative to reduce dependency on oil and deliver consumers with affordable and environmentally-friendly transportation choices. The increase in transportation efficiency is the best place to start efforts to reduce emissions and emission related issues.

1.2 Background Literature on EV Motor Technology

A combination of new design variations, new materials, and sophisticated electronic controls are making electric motors much more agreeable for EV duty. Compared to industrial motors, EV motors require wider speed range and high torque and power densities [4]. The design of an appropriate electric motor for EV depends on the following factors: a) vehicle type, b) the driving profile and type of route followed, in terms of the acceleration demand, climbing and braking performance, maximum speed and range, c) vehicle constraints such as weight, and volume based on vehicle type, d) efficiency, power density (power output/volume), reliability and cost, and d) type and specifications of the energy storage unit. These objectives have to be achieved through essential modifications in terms of design, material selection, innovative component shapes and designs, and control strategies for electric motors to improve efficiency and
decrease cost and NVH (noise, vibration and harshness), while maintaining compact size and light weight.

Performance of an EV relies on the design and appropriate control of its electric propulsion system that consists of a traction motor, its power electronic drive, motor controller, and an energy storage element, such as a battery. Recent efforts are directed towards developing an improved propulsion system for EV applications.

Developing a high-performance electric motor strongly relies on its design. Few factors that optimum design is required to possess are high efficiency, high power density, wide speed ranges, and low torque ripple and noise. Consideration of machine losses leading to lower efficiency is important for three reasons: losses appreciably influence the machine operating cost; determine the heating of the machine and, hence, the rating or power output that may be obtained without undue deterioration of the insulation; and the voltage drops associated with supplying the losses. The materials used in the design should have extremely low magnetization losses, a high degree of ability to integrate, lead to high thermal capacity of the motor that result in longer motor life and higher loading, and less and faster production steps.

Selection of a suitable electric motor to fulfill all the requirements for EV application continues to pose a challenge [4]. Several motors have been analyzed for traction purposes [5]-[8] amongst which induction motors (IMs), permanent magnet synchronous motors (PMSMs) and switched reluctance motors (SRMs) are considered the most fitting candidates.
Focus Electric 2013 shares Ford’s global C-car platform and is based on the glider of the third generation Ford Focus. The electric car is powered by a PMSM rated at 100 kW and uses a 23 kWh capacity lithium-ion battery pack, which together delivers 245 Nm of torque.

Nissan utilizes an 80 kW AC synchronous motor in its all-electric vehicle known as the “Leaf”. The Leaf has a driving range of 100 miles/charge based on the US EPA City Cycle. Nissan has also developed a high power density 3D axial-flux type PMSM that is capable of producing a maximum power of 135 kW [9].

The Chevy Volt plug-in HEV employs a 110 kW, 9,500 rpm, 12-pole PMSM and a 55 kW, 6,000 rpm, 16 pole permanent magnet AC synchronous generator in the vehicle’s drivetrain with a shaft torque of 370 Nm and 200 Nm, respectively. Both the machines of the Voltec electric drive system are interior permanent magnet machines with the motor having a distributed winding and the generator having a concentrated winding in their respective stators.

The BMW i3, previously Mega City Vehicle (MCV), is a 5-door sedan urban electric car developed by BMW. Under the New European Driving Cycle the official range is 130 to 160 km and up to 200 km in the most efficient driving mode. The three-door coupe, like the five-door i3 electric sedan, is propelled by an electric motor developed by the BMW Group, with a maximum output of 125 kW and peak torque of 250 Nm.

The Tesla Roadster employs an AC induction motor (IM) that produces high torque at very low rpm and delivers constant acceleration up to 6,000 rpm. Raser’s
Symetron P-42 IM [10] can deliver high efficiency and high torque over a wide operating speed range. This motor has a pancake shape for easy integration.

Honda vehicles are focused on surface PM and inset PM machines with segmented stator structures and fractional-slot concentrated windings. Segmented structures not only have the potential for increasing copper slot factor and reducing manufacturing costs, but they also compromise the stator back iron rigidity. The Honda designs tend to have low winding factor of 0.866. Honda has developed a thin 15.5 kW PMSM for its Civic Hybrid that uses inset permanent magnets in its rotor, concentrated windings in its stator, and displays a very high efficiency. Also, it can be noted that the Honda designs utilize low DC bus voltage. The machines are designed for tight integration with the ICE-drivetrain housing.

Although permanent magnet machines seem to be the most common motor technology used by major automakers, demerits such as the high current for field-weakening operation, temperature sensitivity, costly magnet, demagnetization of the permanent magnets due to high temperatures, and fabrication difficulties are leading to researchers looking for better options, preferably non-magnet designs [11].

The Squirrel cage IMs are one of the magnet-free motor contenders for the electric propulsion of EVs for their reliability, ruggedness, low maintenance, low cost, and ability to operate in hostile environments. IMs are considered suitable for HEVs owing to their reliability, low cost, and wide speed ranges at constant power. Low efficiency of IMs, mainly owing to copper losses from stator and rotor and subsequent increase in heat is a major limitation in traction application for which several optimal
flux and loss minimization control techniques are proposed [12]-[14]. Other drawbacks include low power factor, and low inverter-usage factor, which is more serious for the high speed, large power motor. One of the most critical problems posed by IM is the presence of breakdown torque limit while operating in extended constant-power operation where at critical speed the breakdown torque is reached. Modifications at the design level and fault tolerant, high performance control techniques are proposed to improve the speed and torque response of IM drives [12]-[15].

It can be observed from the above content that all of the automakers are trying to develop better traction motors through extensive research and development activities as the search for the optimized vehicle traction motor is still continuing. More robust and cost efficient machines, such as switched reluctance motors (SRMs), have potential for the future of automotive applications.

1.3 SRM Research Background

As a magnet free motor with a simple and rugged construction, high speed operation and extended constant power range, SRMs are gaining interest as a potential alternative in EV traction [11], [16]-[18]. SRM is a doubly salient machine with no winding, magnet, or conductive bars on the rotor. It features a simple structure, ruggedness, fault-tolerance ability, high-speed operation capability, high power density, and low manufacturing cost. Owing to the absence of rotor excitation, they are not susceptible to rotor winding failures or demagnetization of the magnets and are excellent candidates for automotive applications, especially for heavy duty electric buses [19]-[21].
Also, in electric drivetrains, the machine has a niche performance over a wide speed range on accurate design and control.

Despite these advantages, there are challenges. For example, torque ripple and acoustic noise need to be addressed through fundamental design improvements in order to develop a viable SRM-based electric propulsion system. Extensive research has focused on the design of conventional radial-flux SRMs, such as the new configuration with a higher number of rotor poles than stator poles tending to increase the torque per unit volume [22]. Desai et al., 2010, observed that machines with increased number of poles without increasing the number of stator phases have shown superior performance with higher torque per unit volume, lower manufacturing costs and comparable torque ripple when compared to a conventional SRM. A novel two-phase SRM with E-core stator was also proposed in [23]. The E-core stator consists of three poles with two poles at the ends having windings and a center pole containing no copper windings. During the operation, the center stator pole is shared by both phases. The uniqueness of the design is that while the air gap around the common stator pole has constant and minimum reluctance, irrespective of rotor position, the two remaining stator poles at the ends experience variable reluctance with respect to the rotor position. Its major advantages were in saving the stator core and eliminating the flux reversal in the stator, hence reducing core losses. More novel designs were presented in [24] and [25], which focused on the radial SRM configuration. In [24], a novel SRM was designed for low-cost production that possessed high slot space for easy coil winding and was found to have better characteristics in terms of torque and efficiency and in [25], a multi-objective optimization procedure and an analytical model derived from reluctance network method
was developed to design a 6/4 type SRM to be employed in EV drive. Research was also conducted on linear SRMs [26], [27]. Linear SRMs gained interest due to the fact that unlike the rotating machine, with only the tangential component force, usually called torque, the linear machine has propulsion, guidance, and normal forces that affect its dynamic motion in three degrees of freedom (3-DOF). However, very little attention has been given to the design of axial-flux switched reluctance motors (ASRMs) [28]. The design in [28] has a complicated structure requiring the use of ferromagnetic material, and hence the design will not be efficient for high power rating.

Further background literature related to research performed in this thesis has been provided separately in corresponding chapters.

### 1.4 Research Objective

The recent interest by the automotive industry in finding an electric motor that has high power density, rugged construction with high temperature adaptability, lower cost, and fault tolerance ability has changed the map of electric motors and revived the research on switched reluctance motors (SRMs). Due to the aforementioned characteristics and several other features [20], SRMs are considered to be an attractive alternative to traditional electric motors in electric and hybrid electric vehicle applications.

In spite of the said advantages, there are, however, a few challenges such as torque ripple and acoustic noise that need to be addressed in order to develop a viable SRM-based propulsion system. Hence, the research objectives of this dissertation are to analyze, investigate, and address the challenges of torque ripple and acoustic noise through
fundamental design improvements keeping the aforementioned industrial targets in parallel.

1.5 Research Contributions

1) Designing and analyzing of a short-flux path outer rotor SRM for an in-wheel drive application aimed for reduction in torque ripple. The proposed machine had reduced torque ripple and higher efficiency than that of the conventional SRM.

2) Designing, prototyping and analyzing a novel 5-phase pancake shaped axial flux SRM to address issues in acoustic noise, vibration and torque ripple while keeping merits of the previously investigated short-flux path configuration. The comprehensive design procedures included a new systematic approach for deriving the output power equation of a unique and newly targeted topology of axial flux machine. It also provides a winding design approach for the process of coil optimization.

3) Investigation of leakage flux in axial flux SRM through novel flux tubing approach and experiments incorporating the 3-D effect and leakage component.

4) Solutions based on new methods of utilizing segmented grain-oriented silicon steel core material for mitigating issues related to leakage flux.

5) Shielding and winding configuration based methods to alleviate issues related to leakage flux.

1.6 Dissertation Outline

The remainder of this dissertation is organized as follows.
Chapter 2 provides a brief background study of the fundamental operating principle and introduction to radial and axial flux SRMs. Thereafter, a case study based on finite element analysis is presented in this chapter in detail to analyze and quantify the major well-known issues of SRM, namely the torque ripple and acoustic noise. The case study deals with two types of SRM designs, one is a commercially available conventional SRM and the other is a new design of in-wheel outer-rotor SRM for EV application. This chapter illustrates the design aspects and discusses the results from comparative performance analysis of both the conventional SRM and the outer rotor SRM.

Chapter 3 firstly illustrates the concept of an axial-flux SRM and describes the key features of the novel five phase pancake shaped SRM proposed. Thereafter, the motor design procedure employed is explained in details along with coil design and optimization tasks undertaken in this thesis.

Chapter 4 presents results obtained from static and dynamic finite element analysis performed on the designed motor. It also provides a comprehensive comparative performance analysis between the 2D and 3D models of the machine. An optimal model of C-core is obtained.

Chapter 5 presents the steps involved in the prototyping process of the designed axial flux SRM. It also presents and discusses results obtained from experimental investigations performed on the prototyped machine.

Chapter 6 illustrates the issues related to leakage flux elicited from experimental and theoretical analysis conducted on the proposed machine through a flux tubing technique.
Thereafter solutions based on materials, shielding and winding configurations has been presented in order to reduce the leakage flux.

*Chapter 7* provides conclusions and findings of this dissertation.
Chapter 2

2. DESIGN AND INVESTIGATION OF RADIAL FLUX SWITCHED RELUCTANCE MACHINES

2.1 Chapter Objectives:

In addition to the background literature review presented in Chapter 1 illustrating previous research activities performed in the area of SRMs, a case study based on finite element analysis is presented in this chapter in detail to analyze and quantify the major well-known issues of SRM, namely the torque ripple and acoustic noise. The case study deals with two types of SRM designs, one is a commercially available conventional SRM and the other is a new design of in-wheel outer-rotor SRM for EV application. Both of the designs have radial flux patterns and equivalent power ratings; however, they differ in their individual flux paths. The current chapter illustrates the design aspects and discusses the results from comparative performance analysis of both the conventional SRM and the outer rotor SRM.

2.2 Introduction and Operating Principle of SRM:

The switched reluctance motor is a doubly-salient singly-excited synchronous machine. It has the simplest motor construction which offers benefits over the conventional motor technologies. The laminated stator stack is analogues to that of the DC motor, where concentrated coils are wound around the stator poles. The rotor is
simply a core of laminated soft magnetic steel with no permanent magnets, conducting bars, or windings on it; hence, brushes are not required as well. Owing to the latter, SRM rotor has a low moment of inertia, thus, giving a large acceleration rate for the motor and superior dynamic response [29].

The reluctance torque is generated through the tendency of the rotor to move to its minimum reluctance position with respect to the excited stator pole. The minimum reluctance is usually reached when the excited stator pole is in full alignment with the rotor pole. Generally, the rotor poles are shaped in such a way as to maximize the variation of inductance with position.

In order to maintain the generated torque of the motor in one direction (positive torque), the windings of successive stator poles should be excited in sequence so that the magnetic field of the stator leads the rotor pole, pulling it forward.

2.3 SRM Classification and Configuration

Based on the orientation of their electromagnetic flux path with respect to the axis of their shafts, SRMs are classified as “Radial SRMs” if the magnetic flux path penetrates the rotor along the radius of it, while traveling from one stator pole to the opposing one. When the flux path through the rotor body is along the axial direction, the machine is then called an “Axial-Flux SRM.”

Depending on the length of the flux path through the rotor, there are short-flux path SRMs versus the long flux path ones that are conventional SRM. The flux path is determined by the winding coils layout design. In the long flux path, the two coils
forming one phase are placed around the diametrically opposite poles. While in the short flux, the two coils are usually wound around adjacent stator poles.

Short-flux path SRMs have the advantage of lower core losses due to the fact that the flux reversals do not occur in stator back iron. They are ideal for applications where the total length may be constrained, such as in a ceiling fan or in a propulsion application.

The variety of combinations of number of phases with stator and rotor number and shapes of poles led to a wide range of possible designs of the SRM. Fig. 2-1. shows two topologies of SRM; a conventional radial 8/6 SRM, and a 12/8 short flux radial SRM. There are various configurations of SRM designs, which were proposed to improve the overall performance of the machine, e.g. a c-core stator was proposed in [30].

Fig. 2-1. Switched reluctance motor topologies. (a), conventional radial 8/6 SRM, (b) short flux radial 12/8 SRM,(c) isometric view of 8/6 SRM, and (d) the actual stator of the 8/6 SRM.
2.4 Case study: Outer Rotor SRM versus 8/6 Conventional SRM

2.4.1 Objective

In this era of electrified transportation, switched reluctance motor (SRM) is emerging as a prospective replacement to traditional electric motors especially for large heavy duty vehicles such as the electric bus. Previous research indicated that the in-wheel outer rotor motor has an edge over the conventional motor designs for the electric vehicle application as it saves substantially large space previously occupied by the necessary mechanical components such as the transmission, speed reducer shafts and differential [31]-[34]. Bullis 2009, indicated that the in-wheel motor was instrumental to improve the efficiency of the electric bus as the bus could travel twice as far as a conventional bus on a liter of diesel. The in-wheel motors conferred additional savings by eliminating the need for a transmission, differential, and related mechanical parts. As a result, both the overall weight of the bus and energy losses due to friction are reduced. The in-wheel motors also improved traction by allowing precise control over each wheel, and they allowed greater flexibility in vehicle design since there was no need to mechanically link the wheels to an engine [35].

This case study proposes the design and analysis of a novel outer rotor in-wheel SRM (IOSRM). The integration of the motor housing inside the wheel rim saves significant space and eliminates the need for additional mechanical parts used in the centralized drive. The procedures of deriving the output power equation as a function of the motor dimensions and parameters are explained in [36]. Comparative finite element
analysis (FEA) is performed between the developed machine and a commercially available conventional SRM.

2.4.2 Concept of the Proposed In-wheel Outer Rotor SRM and its Features

Figure 2-2 shows the isometric perspective illustration of the proposed SRM integrated in a wheel of electric bus. The shaft of the stator core is rigidly fixed to a beam of the rear suspension system. The outer rotor is mounted on the stationary components by a set of bearings that facilitate the spinning of the rotor. The rotor core is firmly fastened to the wheel’s rim by an arrangement of bars and two end rings as shown in Fig. 2-3. This design has 18 rotor teeth and 16 stator poles. The rotor teeth are evenly distributed with 20 degrees spacing. The stator poles are formed of 8 pairs as shown in Fig. 2-4. The angular distance between the axes of each adjacent pair is 45 degrees. Moreover, the angular distance between the poles within the pair is 20 degrees.

The windings of each phase are split on 4 concentrated coils wound in series around 4 stator poles, two on each side diametrically. Upon excitation, the magnetic circuit of each phase is formed by a pair of stator poles facing two aligned rotor teeth as shown in Fig. 2-4. The torque production in this design relies on the tendency of the excited stator poles to pull the nearby rotor teeth into alignment. An important feature of this design topology is that it offers very short flux path thus minimizing the iron loss without compromising the high power capability of the motor. In addition, a minimum of four poles are energized at any given time, which renders this machine at least double the torque capability of the conventional SRM. Additional features of the proposed design are summarized as follows:
The main advantage of using an integrated in-wheel outer rotor electric drive is that it saves substantial space previously occupied by the necessary mechanical components such as the transmission, speed reducer shafts and differential.

The flux path is independent of the radius of the rotor. This particular feature gives the designer the capability of increasing the torque by increasing the radius of the rotor without having to increase the flux path \( (\tau = F \times r) \), where F is the reluctance force generated in the air-gap and r is the radius of the air-gap.
In conventional SRM, one magnetic circuit is shared by all phases, and the windings of two adjacent phases are fitted together in one slot. These contribute to the mutual coupling between adjacent phases. This disadvantage is minimized in this design since each phase has an independent magnetic circuit, and the winding of different phases are further separated.

The direction of the flux in the stator poles is always the same. In other words, the flux reversal does not occur, hence lowering the core losses compared to conventional SRM [37].

Enough spacing between the adjacent magnetic circuits on the stator makes it possible to include liquid cooling tubes as shown in Fig. 2-2. This pacifies most of the heat generated by the stator coils, and, therefore, permits a higher current rating, which in turn increases the output power rating of the machine.

The insertion of cooling tubes has an additional important advantage as it serves as a flux barrier that limits the mutual coupling and the leakage flux between two magnetic
circuits excited at the same time. Therefore, it increases the overall efficiency of the machine and improves its performance.

2.4.3 Geometrical Design of the Proposed IOSRM

Considerable research has been done to alleviate the problem of torque ripple in SRM by proposing intelligent control schemes [38]. Yet, the geometrical structural solution is preferred over the control scheme. Many literatures have suggested large number of phases or poles [39]. The motor in this case study is designed with outer rotor and large number of poles. Consequently, the number of strokes per revolution increases, and the torque ripple problem could be alleviated. The increased number of poles requires a larger diameter, resulting in a greater flux-path length which in turn raises the losses and reduces efficiency. The solution for the problem is addressed by adopting a shorter flux-path as shown in Fig. 2-4.

A small step angle of 5 degrees is achieved by adopting the configuration explained in this chapter. The step angle is calculated using the following:

\[
\varepsilon = \frac{360^\circ}{N_{ph} \times N_r}
\]  

(2.1)

where, \(N_{ph}\) and \(N_r\) are the number of phases and the number of rotor poles, respectively.

Generally, the initial design process goes through several iteration steps. Several geometries have been calculated with varying pole numbers and pole dimensions, keeping in mind that a number of requirements need to be fulfilled such as; minimizing the step size (\(\varepsilon\)), the self-starting capability, and the optimum pole arcs. The stator arc (\(\beta_s\))
should be greater than the step size in order to satisfy the self-starting requirement. The optimum pole arcs, which are a trade-off between various conflicting requirements, should be made as large as possible to maximize the aligned inductance and the flux linkage. However, if they are too wide the clearance between the rotor and stator pole-sides in the unaligned position will be not enough. This restriction can be represented by:
Table 2-1. Dimensions of the Proposed IOSRM

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of phases $N_{ph}$</td>
<td>4</td>
</tr>
<tr>
<td>Air-gap length $g$</td>
<td>0.9 mm</td>
</tr>
<tr>
<td>Stator-rotor configuration 16/18</td>
<td></td>
</tr>
<tr>
<td>Motor axial length $l_m$</td>
<td>180 mm</td>
</tr>
<tr>
<td>Rotor pole pitch $\tau_r$</td>
<td>$20^\circ$</td>
</tr>
<tr>
<td>Stator outer radius $r_{os}$</td>
<td>134.1 mm</td>
</tr>
<tr>
<td>In-pair stator pole pitch $\tau_s$</td>
<td>$20^\circ$</td>
</tr>
<tr>
<td>Rotor inner radius $r_{ir}$</td>
<td>135 mm</td>
</tr>
<tr>
<td>Stator pair-pair pitch $\tau_{app}$</td>
<td>$45^\circ$</td>
</tr>
<tr>
<td>Rotor outer radius $r_{or}$</td>
<td>180 mm</td>
</tr>
<tr>
<td>Stator pole arc $\beta_s$</td>
<td>$8^\circ$</td>
</tr>
<tr>
<td>Number of turns $N$</td>
<td>160</td>
</tr>
<tr>
<td>Rotor pole arc $\beta_r$</td>
<td>$8^\circ$</td>
</tr>
<tr>
<td>Step angle $\varepsilon$</td>
<td>$5^\circ$</td>
</tr>
</tbody>
</table>

\[
\frac{2\pi}{N_r} = \beta_r \beta_s
\]  

(2.2)

The optimum pole arcs are somewhere between these extremes [39]. An adequate choice for this design was to have both stator and rotor poles’ arcs equal. The main dimensions are presented in Table 2-1, and illustrated in Fig. 2-3.

Further details including the output power derivation are provided in [37].

2.4.4 Comparative Finite Element Analysis of the IOSRM and the Developed Conventional SRM

The FEA model of the proposed IOSRM design is built using MagNet Infolytica software. Details on finding the optimum magnetomotive force and winding design are provided in details in [36]. The static analysis is used to derive the operating current and to validate the correctness of the design. The sequential excitation of the motor phases
causes rotation of the outer rotor in counter clockwise direction. The color coded solution plot given in Fig. 2-5 shows satisfactory level of local saturation in the regions of the overlapped poles. The magnetization characteristics, the most descriptive illustration of the motor performance and efficiency, are obtained and presented in Fig. 2-6. It can be seen that the value of inductance ratio at the rated current (ratio of aligned flux linkage to the unaligned one at a specific current) is relatively high for a motor with large number of poles based on comparison with the results in [40], and [41]. High inductance ratio is directly reflected in the output power, hence validating the design with respect to the efficiency improvement. The IOSRM power ratings are listed in Table 2-2.

An FEA model of 8/6 conventional SRM with the same power rating and identical related geometrical dimensions to the IOSRM is built for comparative analysis purpose. Both models phases are current driven with the same values. The field solution of the developed conventional SRM is as shown in Fig. 2-7. The output torque obtained during single stroke of operation by both the machines is demonstrated in Fig. 2-8. Since their poles arcs are not equal, the rotor positions scales are taken as per unit quantities of each machine. The comparison in Fig. 2-8 illustrates the improvement in the output torque and, hence, in the efficiency. Fig. 2-9 shows the output torque obtained by the individual consecutive phases. The obvious large overlapping of 50% suggests that there are no dead torque zones at the output. This is due to the small step angle in this design, which is in turn governed by the large number of poles and special configuration design of the poles. It can be concluded here, based on the large overlap between consecutive phases torque profiles, this design can minimize the torque ripple to a very low level, by applying basic control techniques that boost of the current at the low torque regions.
Fig. 2-5. FEA solution for the sequential excitation of all phases and the corresponding motion of the outer rotor.

Fig. 2-6. The saturation characteristics of the proposed motor obtained while varying the rotor angular position over one stroke by step of 2 degrees.
Fig. 2-7. The field solution of an 8/6 conventional SRM FEA model built with similar size and power rating to the IOSRM.

Fig. 2-8. Output torque obtained by both the machines over one single stroke at rated current of the machines.

Table 2-2. Rating of the Proposed IOSRM.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Values</th>
<th>Parameter</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Power</td>
<td>24 kW</td>
<td>Maximum power</td>
<td>40 kW</td>
</tr>
<tr>
<td>Rated torque</td>
<td>120 N.m</td>
<td>Maximum torque</td>
<td>200 N.m</td>
</tr>
<tr>
<td>Rated current</td>
<td>35 A</td>
<td>Base speed</td>
<td>2,000 rpm</td>
</tr>
</tbody>
</table>
The developed FEA models solutions have also shown the radial forces experienced by the motors structure. Radial forces are the main cause to the noise and vibration in SRM [27], [42]. As observed in Fig. 2-10, the model of the conventional SRM is rotated by 15 degrees in order to align its radial forces along the Y axis. Similarly, the IOSRM is also rotated for the same purpose, as seen in Fig. 2-11. The radial force seen at Y axis is provided in Fig. 2-12 for the conventional SRM and in Fig 2-13 for the IOSRM. These are static FEA tests under rated current and full alignment position, and are sufficiently suitable for the purpose of comparison. While both designs showed significant radial forces, the IOSRM showed slightly higher values.

Fig. 2-9. Output torque showing large overlapping with no dead torque zones.
Fig. 2-10. The conventional SRM model rotated by 22.5 degrees in order to align its radial forces along the Y axis.

Fig. 2-11. The IOSRM model showing the excited winding aligned with the Y axis inorder to capture the radial forces.

Fig. 2-12. Static FEA showing the radial force experienced by the conventional SRM model.
Fig. 2-13. Static FEA showing the radial force experienced by the IOSRM model.

2.5 Conclusion

The results obtained through FEA investigations show good output torque profile in term of low torque ripple, observed for a motor without applying any control technique. The motor efficiency is also validated based on the satisfactory inductance ratio’s value of approximately (4). The low torque ripple can be attributed to the increased number of poles, and the satisfactory efficiency can be attributed to the short-flux path configuration in the machine. However, the radial forces which accounts for the acoustic noise and vibration are still high.

Hence, a major design modification in SRM technology is necessary in order to keep the merits such as efficiency and reduced torque ripple and simultaneously overcome challenges such as acoustic noise and vibration in the machine. This is one of the key motivating points to design and investigate axial-flux SRMs whose design and analysis is presented in the forthcoming chapters.
3. DESIGN OF A NOVEL 5-PHASE PANCAKE SHAPED AXIAL FLUX SRM (ASRM)

3.1 Introduction

Electrified automotive industrial research on axial flux motors has significantly increased in the last few years and been adopted by few automakers [9] due to its higher torque density by both volume and weight as compared to the radial-flux configuration.

Axial flux machines utilize the active radial part effectively, as required in high power density applications. For higher aspect ratio (outer diameter to overall axial length), power density of axial flux machines is higher than that of radial flux machines. For in-wheel EV applications, the torque requirement is high, but the size of the motor is restricted by the wheel rim. Nissan Motors is investing largely in research for enhancing the new configuration of axial flux PMSM that has high number of phases [9].

In [43], the authors have developed an axial flux segmented toroidal type winding SRM for EV applications. The new design offered advantages of low copper losses and end winding volume compared to conventional SRM.

Axial and radial SRMs generally share the same operating concepts and features, except the orientation of the flux as it passes through the rotor; parallel to the axis of
rotation in axial motors and perpendicular in radial machines. Thus, the design procedures vary due to this geometric dissimilarity.

In this chapter, a new design of axial-flux switched reluctance motor (ASRM) is proposed. The motivation behind this novel machine design is to develop a viable axial flux SRM for EV applications by addressing issues such as acoustic noise and vibration, less torque ripple and yet keeping the efficiency and power density equivalent or better than conventional radial SRMs.

### 3.2 Axial-flux SRM Concept and Description

The conceptual diagram for the proposed motor is shown in Fig. 3-1. The stator is composed of 15 c-core, each of which has individually wound coil. The rotor which has a disc shape could be made of any low magnetic permeability material. Through the rotor disc, 12 square holes are created to be filled by 12 cubes of high magnetic permeability material such as the silicon steel. The torque production in this design relies on the tendency of any of these cubes (which can be considered as the rotor’s poles) to align with the two poles of an energized c-core, providing the minimum reluctance path to the magnetic circuit of one c-core. The motor has 5 phases and 3 repetitions as shown in Fig. 3-2, where the 15 cores on the stator are divided into 3 sets displaced by 120 mechanical degrees for better forces distribution on the structure of the motor.
Fig. 3-1. The proposed switched reluctance motor. (a) Whole motor. (b) Rotor.

Fig. 3-2. Overview of the proposed motor.

3.2.1 Torque Ripple Reduction

Torque ripple is defined as

\[ TR = \frac{T_{\text{Inst}_{\text{max}}}-T_{\text{Inst}_{\text{min}}}}{T_{\text{Ave}}} \]  

(3.1)

According to [44], the majority of torque ripple occurs in the phase overlap region where the torque producing responsibility is commutated from one energized phase to
another. Poor overlap ratio (less than 10%) is the main cause of the torque ripple. The overlap region is greatly influenced by the step size that is defined as follows

\[ \varepsilon = \frac{360^\circ}{N_{ph} \times P_r} \]  

(3.2)

where, \( N_{ph} \) and \( P_r \) are the number of phases and the number of rotor poles, respectively.

It can be noted that both the number of phases and the number of poles are required to be maximized for the purpose of reducing the torque ripple. However, to adopt high number of phases or poles the diameter of the motor has to be increased which in turn increases the flux path length and hence the losses increase, this is for the case of conventional radial SRM.

In the proposed design, the flux path is independent of the rotor diameter. Therefore, the latter can be increased to allow more space for higher number of poles without affecting the magnetic loading. As a result of having a higher number of poles, the step size is reduced which in turn results in minimizing the torque ripple.

### 3.2.2 Vibration and Acoustic Noise Minimization

According to literatures [27], [42], the acoustic noise in reluctance motors is mainly generated by the electromagnetic radial force produced by excitation as seen in Fig. 3-3 By examining the proposed design in Fig. 3-4, it can be seen that the reluctance force developed at the rotor’s pole has only tangential components with no radial component produced.; therefore the vibration and the acoustic noise in this design are brought to a low level.
Fig. 3-3. Reluctance force components of radial field motor.

Fig. 3-4. Reluctance force component of proposed motor.
3.2.3 Key Features of the New Design

This design has additional features as compared with typical SRMs, they are summarized as follows:

1. Higher torque and power density:

   \[ P = \frac{\pi}{120} k_e k_d k_u B A_s D^2 l_m N_r \]  
   \[ (3.3) \]

   where,
   - \( k_e \): the efficiency constant,
   - \( k_d \): the duty cycle constant,
   - \( k_u \): constant relates the inductance ratio \( K_u = 1 - \frac{L_u}{L_a} \),
   - \( L_a \): the aligned saturated inductance per phase,
   - \( B, A_s \): are the magnetic and electric loading respectively,
   - \( D \): the bore diameter,
   - \( l_m \): the length of motor,
   - \( N_r \): the rotor speed (rpm).

   From (3.3), the standard output equation of SRM [37], it can be seen that for a given magnetic and electric loading, the torque and power are proportional to the square of the diameter. The proposed design has the ability to increase the torque by increasing the diameter of the rotor, without increasing the flux path length.

2. Larger space available for the coils gives the designer more flexibility in determining the number of turns so the resistance and copper losses are reduced unlike the conventional SRMs that are limited by the slot space. This is illustrated in the specific electric loading formula [38]:
\[ A_s = \frac{2N \cdot I \cdot m}{\pi D} \] (3.4)

where, \( N \) is the number of turns per phase, \( I \) is the current, and \( m \) is the number of phases conducting simultaneously.

3. Lower core losses than that in conventional SRM due to the fact that flux reversals do not occur in the stator back iron in addition to having short flux paths.

4. All cores are electrically and magnetically isolated from each other so that they can be wound individually without complex winding equipment and by automated process.

5. The inertia of the rotor is small since it can be made of a material that has much lesser mass density than that of the steel which is used in conventional SRM’s rotor. In addition, this inertia could be made smaller by creating some openings in the rotor body keeping in mind not to exceed its solidarity limits.

6. Due to its unique pancake shape, this design can be augmented for higher power rating.

7. This design provides a maximized thermal dissipation due to the following:
   - Thin structure of the motor has a better thermal dissipation factor
   - The heat generation source in motors is usually the coils. In this design, they are located on the outer circumference which
facilitates a faster and more efficient heat transfer to the outer ambient

- This special structure also enables it to adopt advanced cooling systems such as water jacket around the motor case.

3.3 Motor Design Procedures

3.3.1 Geometry Design

The aim of the design is to provide a feasible solution to the well-known drawbacks of SRMs namely the acoustic noise and torque ripple. This demerit is particularly undesirable for the vehicle applications. A considerable work done in the analysis of the acoustic noise suggests that the radial forces account for most of it as mentioned earlier in this thesis. Therefore to meet this constraint, we designed a motor with axial flux so that the magnetic forces are only in the tangential plane to the rotor circumference, thus reducing the radial forces to a very low level.

And regarding the torque ripple, a possible solution according to [30] is to consider higher number of phases so that the number of strokes per revolution increases; as a result the torque dip problem could be alleviated. This solution is harder to achieve in case of conventional SRM because increasing the number of phases requires larger diameter resulting in a greater flux-path length which in turn raises the losses and reduces efficiency. The solution for this was addressed in this design by adopting a shorter flux-path as shown in Fig. 3-5.
The subsequent step of the design procedure was with the hand calculation of several geometries with varying pole numbers and pole dimensions keeping in mind that several requirements need to be fulfilled such as; minimizing the step size (ε) also known as stroke in some literature, the self-starting capability, and the optimum pole arcs. In this design, a small step angle of 6° was achieved, and calculated as in (3.2).

To achieve the self-starting requirement, the stator arc (βs) should be greater than the step size (ε). The optimum pole arcs are a compromise between various conflicting requirements. On the one hand they should be made as large as possible to maximize the aligned inductance and the maximum flux linkage. However, if they are too wide there is not enough clearance between the rotor and stator pole-sides in the unaligned position. This restriction can be represented by:

\[
\frac{2\pi}{P_r} - \beta_r \geq \beta_s
\]

(3.5)

The optimum pole arcs are somewhere between these extremes. Generally, for very high efficiency designs, the slot area needs to be maximized and this leads to a narrower pole arc [39]. However, in the proposed design this problem does not exist as there is enough space for the windings around the stator cores owing to its unique topology as can be seen from the top view of the motor in Fig. 3-6. An adequate choice for this design was to have both stator and rotor poles’ arcs equal. Fig. 3-7 shows the ideal inductance profile for all phases. It can be noted that the overlap region was increased to offer stronger self-starting capability, large output torque, and little torque ripple. The main parameters and dimensions are presented in Table 3-1, and illustrated in Fig. 3-6, and Fig. 3-8.
Fig. 3-5. Short magnetic flux-path.

Fig. 3-6. Geometry parameters of the proposed SRM.

Fig. 3-7. Five phases inductance profile.
Fig. 3-8. 3-D sketch for one stator c-core of the proposed motor, showing the dimensions and the orientation of laminations.

Table 3-1. Motor Parameters.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>The number of phases</td>
<td>5</td>
</tr>
<tr>
<td>The stator-rotor configuration</td>
<td>15/12</td>
</tr>
<tr>
<td>The rotor pole pitch $\tau_r$</td>
<td>30°</td>
</tr>
<tr>
<td>The stator pole pitch $\tau_s$</td>
<td>24°</td>
</tr>
<tr>
<td>The rotor pole arc $\beta_r$</td>
<td>9.6°</td>
</tr>
<tr>
<td>The stator pole arc $\beta_s$</td>
<td>9.6°</td>
</tr>
<tr>
<td>The outer diameter of the rotor $D$</td>
<td>320 mm</td>
</tr>
<tr>
<td>The radius of the centre of a rotor cube $r$</td>
<td>144 mm</td>
</tr>
<tr>
<td>The air gap length $g$</td>
<td>0.3 mm</td>
</tr>
<tr>
<td>The core height $h_c$</td>
<td>93 mm</td>
</tr>
<tr>
<td>The core width $W_c$</td>
<td>72 mm</td>
</tr>
<tr>
<td>The core thickness $t_c$</td>
<td>24 mm</td>
</tr>
<tr>
<td>The rotor thickness $t_r$</td>
<td>24 mm</td>
</tr>
<tr>
<td>The stator pole depth $d_s$</td>
<td>30 mm</td>
</tr>
<tr>
<td>The coil length $l_c$</td>
<td>45 mm</td>
</tr>
<tr>
<td>Number of turns $N$</td>
<td>40</td>
</tr>
</tbody>
</table>
3.3.2 C-core Magnetic Circuit Design Analysis

In the proposed motor design, the fact of having completely separated magnetic circuits represented by the c-cores simplifies the designing procedures as compared to the conventional SRM. Considering one c-core along with one rotor cube that completes the magnetic circuit as shown in Fig. 3-5, we can write the required magnetomotive force (MMF) to produce the flux density in the air gap as well as the stator core and the rotor cube, as follows

\[
F = F_g + F_c + F_r
\]  

(3.6)

where, \(F\) is the total MMF per phase applied and \(F_g, F_c,\) and \(F_r\) are the MMF drops in the air gap, c-core stator, and rotor cube respectively. In this design, the cross-section and material of the stator and the rotor are identical. Therefore \(F_c\) and \(F_r\) can be together considered as a part of the circuit, and equation (4.12) can then rewritten as

\[
F = N \frac{B_g}{\mu_0} \cdot g + \frac{B_c}{\mu} l_{\text{core}}
\]  

(3.7)

where,

\(N\) : number of turns,
\(B_g\) : magnetic flux density in the air gap,
\(B_c\) : magnetic flux density in the core,
\(g\) : air gap length,
\(l_{\text{core}}\) : flux path length in the core and rotor cube,
\(\mu\) : magnetic permeability of the core material,
\(\mu_0\) : magnetic permeability of the free space.
In the aligned position the phase inductance is at its maximum value because the magnetic reluctance of the flux path is at its lowest. At low current levels, before the saturation starts, most of the reluctance is in the air gap because the magnetic permeability of the core material is much higher than that of the air ($\mu > \mu_0$). In addition to it, the flux path length in this design is relatively short, as a result, the second term on the right-hand side of equation (3.7) can be neglected and the minimum required current can be determined.

On the other extreme, working in the heavily saturated region is also not recommended as it increases the iron losses which in turn reduce the efficiency. The current at which this occurs will be obtained by utilizing the data from the $B-H$ characteristics of the core material used (M19-29), without neglecting the second term in (3.7) at this time. It was found that the minimum current required is 17 A and the maximum allowable is 275 A. The optimum current can thus be derived on the basis of the maximum increment of co-energy later in this thesis.

In spite of having a completely different structure topology, the new motor has the fundamental torque production mechanism similar to the conventional SRMs. The general expression for the torque produced by one phase at any rotor position is

$$T = \left[ \frac{\partial W_{co}}{\partial \theta} \right]_{i=\text{const.}}$$  \hspace{1cm} (3.8)

where, $W_{co}$ is the coenergy, $\theta$ is the angular rotor position. At any position the coenergy is the area below the magnetization curve so it can be defined as follows
where, \( \psi \) is the flux linkage at any rotor position as a function of the current.

### 3.3.3 Derivation of output power equation

Similar to AC machines, SRM design starts with deriving its output power equation, which is a function of the specific electric loading, magnetic loading, motor speed, and the dimensions of the machine [37], [45]. The axial SRM design procedure presented in [46] is suitable for their specific geometry, but cannot be applied to the design proposed in this dissertation. However, the initial steps are the same. Assuming that the phase current is flat-topped during the phase conduction period and the phase winding resistance is negligible, the voltage equation for one phase can be given as:

\[
V = I \left[ \frac{L_u - L_a}{\Delta t} \right]
\]

(3.10)

where \( L_a \) and \( L_u \) are the inductances in the aligned and unaligned positions, respectively. \( \Delta t \) is the time taken for the rotor to move from the unaligned to aligned position. \( \Delta t \) can be related to the angular speed of the rotor, \( \omega_b \), and the stator arc, \( \beta_s \), as follows:

\[
\Delta t = \frac{\beta_s}{\omega_b}.
\]

(3.11)

The ratio, \( \alpha \), of the aligned and unaligned inductances is defined as:

\[
\alpha = \frac{L_a}{L_u}.
\]

(3.12)

By inserting (3.11) and (3.12) into (3.10), (3.13) can be obtained.
The flux-linkage $\psi$ at the aligned position is given by:

$$\psi = L_a I = B A_{sp} N.$$  

(3.14)

where $A_{sp}$ is the stator pole area, $N$ is the number of turns per phase, and $B$ is the average flux density at the stator pole face. The value of this average flux density is obtained from the $B$-$H$ characteristics of the material used by choosing the conventional SRM operating area as the window.

To derive the output power and voltage equations for this design, the relationship that links all the relevant variables and geometric parameters has to be found. The cross-sectional area for the stator pole $A_{sp}$ can be derived considering the geometric parameters in Fig. 3-6 and Fig. 3-8 as:

$$A_{sp} = \beta_s r l_r.$$  

(3.15)

where $r$ is the rotor cube to shaft distance and $l_r$ is the rotor cube line length. Substitution of (3.14) and (3.15) in (3.13) gives the voltage equation for the proposed motor as shown below:

$$V = \omega_b B r l_r N (1 - 1/\alpha).$$  

(3.16)

The phase current can be found from the specific electric loading, which is defined as the stator ampere-conductors per meter of stator periphery at the air-gap. The specific electric loading is mainly governed by the available space for winding and by the methods of heat dissipation or cooling systems. It is derived for this design as follows:
where $N_{rep}$ is the number of phase repetitions, and $m$ is the number of phases which are conducting simultaneously. The output power for an SRM is defined as follows [37]:

$$P = m k_e k_d V I$$

(3.18)

where $k_e$ is the efficiency factor, usually in the range of 0.8 and 0.93, and $k_d$ is the duty cycle which can be defined as:

$$k_d = \frac{\theta_i N_{ph} P_r}{2\pi}$$

(3.19)

where $\theta_i$ is the mechanical angle during which the current is flowing through one phase winding. Due to the large number of poles, the phase-overlapping ratio is relatively high, so the step-angle is relatively small ($6^\circ$). As a result, the value of $k_d$ becomes equal to 1.

Finally, the output power equation for the proposed SRM design can be found by substituting the voltage equation in (3.16) and the current equation in (3.17) into (3.18):

$$P = k_e k_d A_s \omega B_l r^2 \pi r^2 (1 - 1/\alpha).$$

(3.20)

### 3.3.4 Coil Design and Optimization

In conventional SRMs, windings are concentrated around each of the stator poles. Each pair of diametrically opposite stator pole windings are connected either in series or in parallel to form an independent machine phase winding. However, the ASRM design in this dissertation has each phase winding formed by a single cylindrical coil wound around one c-core.
The targeted power rating of the machine dictates the required magnetomotive force to saturate the magnetic circuit. Therefore, the winding in an SRM, as the current-carrying component, serves to produce the required flux density in the air-gap. One main challenge winding designers face, is how to create enough volume within the stator core in order to accommodate the winding coils.

The winding design objective is to achieve a certain level of flux density at the air-gap with minimal copper and iron losses. The parameters that play a role in the coil design can be confined to winding slot space, number of turns, coil dimensions, switching frequency, skin effect, maximum permissible current density, insulation and the cooling methods to be used. Most of these parameters have been well documented in literature for the conventional radial-flux SRM. In this work the analysis will mainly focus on the parameters that are important from the perspective of the proposed motor design.

3.3.4.1 Determination of a Single C-core Coil Inductances

The accurate calculation of aligned and unaligned inductances has always been an essential step in the SRM design process, as it can predict the output power and performance of the machine. Different methods have been adopted to analytically calculate these parameters, such as the dual energy method [47] which is one of the techniques used in finite element analysis. The proposed design in this dissertation has the magnetic circuit (c-core and cube) for each phase separated from the other phases. This particular feature offers multiple advantages, such as: 1) a reduction in the mutual inductance between phases; and 2) a simplified design process. Due to this simplicity of having individual and separated magnetic circuits, the inductances (aligned and unaligned) of a single c-core coil can be derived by adapting the magnetic equivalent
circuit (MEC). The straightforward (MEC) lumped-parameter method is used here to model a c-core along with two cubes in different positions, in an electrical network representation. The flux tube technique is used to derive the reluctances in the air-gap region. This approach has been used widely in electric machine analysis [48]-[50] and is being incorporated here.

Assuming specific flux tube geometries depend to some extent on the skill and care by which it is applied. And the observation of the flux distribution plot obtained by FEM can provide considerable assistance [51]. Although FEA is considered the most accurate method, it is still desirable to produce geometrical representation and analytical formula, for inductances estimated by other techniques, as will be seen in the next sections.

For flux tube of \( l \) as length, \( A \) as cross-sectional area, and \( \mu \) as permeability, the reluctance is given as:

\[
\mathcal{R} = \int_0^l \frac{1}{\mu A(x)} dx
\]  

(3.21)

In Fig. 3-9, the flux lines are confined within specific paths or regions called flux tubes. Their reluctances are functions of their geometrical dimensions and permeabilities. Having the latter derived, the inductance can be found from the general expression in (3.22):

\[
L = \frac{N^2}{\mathcal{R}_{\text{total}}}
\]  

(3.22)
3.3.4.1.1 *Inductance Derivation for the C-core Unaligned Position:*

The unaligned inductance is considerably difficult to determine analytically because of the complexity of the magnetic flux paths in the large air-gap. However, the core material is not susceptible to saturation due to the large air-gap. The MEC is developed with respect to the c-core geometric parameters. Fig. 3-10 shows the MEC model for a single c-core and two cubes in a full unaligned position. Four main different reluctances are observed. The appropriate flux tubes are proposed in order to calculate the total reluctance of the magnetic circuit as follows:

1. The pole-to-pole reluctance ($\mathcal{R}_{pp}$): is derived using a rectangular cross-sectional area through the large air gap as seen in Fig. 3-9. The integration in (3.21) is applied in order to find the formula of this segment’s reluctance, as follows:

$$\mathcal{R}_{pp} = \frac{g_{pp}}{\mu_0 l_r l_t}$$  \hspace{1cm} (3.23)

where $g_{pp}$ is the pole-to-pole air-gap length, and $l_r$ and $l_t$ are the pole dimensions in the radial and tangential directions respectively.
Fig. 3-9. The magnetic flux lines in the air-gap represented by flux tube. (a) For full aligned rotor position. (b) For full unaligned position. $\mathcal{R}_{p1}$ and $\mathcal{R}_{p2}$ represent the reluctance of the two poles of single c-core, $\mathcal{R}_{c\text{-back}}$ is the back core reluctance, $\mathcal{R}_{c1}$ and $\mathcal{R}_{c2}$ represent the reluctance of the corresponding cube.

Fig. 3-10. The magnetic circuit model for single c-core and two cubes in a full unaligned position.

2. The air-gap fringing reluctance ($\mathcal{R}_f$): the flux tube is approximated to be a semicircle wedge. Fig. 3-9 shows two wedges in the 2-D plane. However, the 3-D effect is
incorporated by deriving reluctances for the four wedges at all sides of the stator pole. Their reluctances are derived as follows:

\[
\mathcal{R}_f = \left\{ \begin{array}{l}
\frac{\mu_0 g_{pp} l_r}{2 + g_{pp}} \\
\frac{1 + \pi/2}{\mu_0 l_r}
\end{array} \right. \\
\mathcal{R}_f(1/2) = \frac{1 + \pi/2}{2\mu_0 l_r}
\]

(3.24)

Similarly, the fringing reluctance in the third dimension is given as:

\[
\mathcal{R}_f(3/4) = \frac{1 + \pi/2}{2\mu_0 l_t}
\]

(3.25)

3. The pole-to-cube reluctance (\(\mathcal{R}_{pc}\)): as demonstrated in Figs. 3-9, there are four reluctances that fall under this classification; \(\mathcal{R}_1, \mathcal{R}_2, \mathcal{R}_3, \) and \(\mathcal{R}_4\). The reluctances of the cubes are negligible in comparison to the large air-gap reluctance because they will not be saturated, in case of unaligned position. As a result, the pole–to-cube reluctance is reduced to \(\mathcal{R}_1\) only. Furthermore, \(\mathcal{R}_1\) can be divided into two flux tubes in series; \(\mathcal{R}_{pc1}\) and \(\mathcal{R}_{pc2}\). Therefore, the overall reluctance is produced as:

\[
\mathcal{R}_{pc} = \frac{t_{pc} + \pi l_t/2}{\mu_0 l_t}
\]

(3.26)

where \(t_{pc}\) is the pole to cube distance in the tangential direction.
4. The leakage flux reluctance ($\mathcal{R}_l$): as the reluctance of the winding slot region, $\mathcal{R}_l$ has a significant share of the total reluctance. Moreover, this design has greater flux leakage, due to the fact that the winding is less surrounded by steel than in radial flux machines. Taking the coil cross section, $A_c$, as the flux tube cross section, and the coil length, $l_c$, as the flux path length, the leakage flux reluctance can then be given as:

$$\mathcal{R}_l = \frac{l_c}{\mu_0 A_c}$$  \hspace{1cm} (3.27)

Having defined all the reluctances in the MEC, and by neglecting the core reluctances, the equivalent reluctance in the case of an unaligned position is reduced to the latter four main reluctances in parallel. Thus, the total reluctance is calculated accordingly and then substituted in (3.22) to give the unaligned inductance as follows:

$$L_u = \frac{N^2 \mu_0}{\mu} \left[ l_c l_c (2 + \pi) \left( 2 t_{pc} + \pi l_c + 2 g_{pp} \right) + \left( 2 g_{pp} t_{pc} + \pi g_{pp} l_t \right) l_c (l_c + l_t) + A_c (2 + \pi) \right] \left( l_c g_{pp} (2 + \pi) (2 t_{pc} + \pi l_t) \right) \right] (3.28)

### 3.3.4.1.2 Inductance Derivation for the C-core Aligned Position:

For the fully aligned position, the task is relatively easier because only two types of flux tubes are present. At the rated current, the magnetic circuit is saturated. So, the $B$-$H$ curve for the core material is utilized to obtain its permeability. The inductance at full alignment with the fringing effect is accordingly developed as follows:

$$L_a = \frac{N^2}{\mu A_c} \left[ \frac{g_a}{\mu_0 A_g} \left( \frac{\pi}{2} \right) \left( \frac{g_a}{\mu_0 A_g} + \frac{\pi}{2} \right) \right]$$  \hspace{1cm} (3.29)
where \( N \) is the number of turns, \( g_a \) and \( A_g \) are the air-gap length and the air-gap cross-sectional area respectively for fully aligned position. For per-phase rated current of 27 A, the calculated values for \( L_a \) and \( L_u \) of a single c-core are 2.6 mH and 0.9 mH respectively.

### 3.3.4.2 Relationship of Skin Depth with Switching Frequency

The skin effect is another important factor that needs to be addressed when selecting the conductor diameter or number of strands. The skin effect gets worse for high excitation frequencies [52]. The phase switching frequency for the designed motor at the maximum operating speed of 6,000 rpm can be derived by determining the time it takes one rotor pole to move from one phase to the adjacent one, as follows:

\[
T_s = \frac{N_{ph} \varepsilon}{\omega_b} \left\{ \frac{1}{\omega_b P_r} \right\}
\]

\[
= \frac{2\pi}{\omega_b P_r}
\]

(3.30)

For a step angle of 6°, the phase switching frequency is calculated to be 1,200 Hz. By referring to the *Handbook of Electronic Tables and Formulas for American Wire Gauge*, a 7-gauge conductor seems to be appropriate for the proposed design. For up to 1,300 Hz switching frequency, it can utilize 100% of the conductor area. Alternatively, use of a multiple strands conductor, such as a five-strand 14-gauge wire, will be equivalent in terms of its resistance and maximum allowable ampacity, as it further permits a higher switching frequency. Note that the maximum permissible current density rule is considered in the design. Depending on the insulation and the method of cooling to be used, the current density is chosen to be 6 A/mm².
3.3.4.3 Winding Slot Space

Figure 3-11 shows the rectangular cross-section (winding space) of one phase coil. The optimized area for this cross section would be a square, but since there are mechanical restrictions, this area was kept as close to a square as possible. It is designed to enclose an area of 1,200 mm², which will be filled by round conductors.

The fill factor $k_{sf}$ for a circular cross section conductor can be derived from the ratio of the area of the circle to the area of the square that surrounds tangentially the same circle. Thus, the theoretical fill factor value is 78%. However, the practical fill factor is found to be as high as 40% [53].

It is assumed that the available winding slot area is enough to accommodate the required copper so that the current density through the coil conductors does not exceed the maximum permissible value.

![Fig. 3-11. The available area for winding (slot area) of one c-core.](image)
A flow chart describing the complete process of the coil optimization is presented in Fig. 3-12. In the first stage, the design objective is to find the minimum number of turns at which the per coil current does not exceed $I_{\text{max}}$.

In the second stage, the number of turns is determined so that the back-EMF equals the supply voltage at the base speed.

The third stage determines the number of strands so that for the maximum switching frequency, 100% skin depth of each conductor is utilized.

Multiple 3-D FEA models of the c-core are built each with a different number of turns while keeping the same supply voltage of 140 V. The calculation was performed for a fixed speed of 2,400 rpm. The results in Fig. 4-1 demonstrate the inverse relationship between the number of turns and both of the current and back-EMF, and verify the first and second stages of the flow chart.
Motor power rating, core dimensions and material determine the required magnetomotive force $F$. Assume initial values for $V_s$, $I_{\text{max}}$, $\omega_b$.

Initial number of strands, $N_{st}=1$

\[ f_s = \frac{\omega_b \times P_r}{2\pi}, \quad A_w = \frac{A_c \times k_{sf}}{N \times N_{sl}} \]

Obtain $f_{\text{max}}$ for the conductor's cross-sectional area $A_w$

\[ f_s < f_{\text{max}} \quad \text{Yes} \]

\[ N_{st} = N_{st} + 1 \quad \text{No} \]

Store $N_{st}$

Fig. 3-12. Stages of coil design optimization.
Fig. 3-13. The current through one e-core coil obtained by 3-D FEA models at fixed speed of 2,400 rpm, and for different number of turns.
Chapter 4

4. FINITE ELEMENT MODELS AND ANALYSES OF THE DESIGNED AXIAL FLUX SRM

4.1 Optimal design and static 2D/3D analysis of the proposed ASRM

Although axial field motors require 3-D FEA modeling, a conversion method was developed previously by the authors [54], to model the new design in 2-D. When modeling in 2-D, there are some effects that should be considered in order to obtain a model that signifies the real performance of the machine. Different correction methodologies have been presented in literature to compensate the effects that 2-D modeling cannot take into account. The common idea is to use empirical correction factors [55], [56]. If the machine has a conventional design, the empirical factors may be quite well known; hence, having a 2-D FEA is sufficiently accurate. Conversely in the case of new motor design concepts, there is a considerable need for accurately determining the effect of the geometry changes and saturation on motor efficiency and other parameters related to the magnetic field.

For this study, the end winding effect and radial flux fringing constitute a considerable portion among the other effects, as will be seen later in this chapter.
4.1.1 Development Process of the 3-D Machine Model

The SRM under investigation features a set of isolated magnetic circuits; each represented by a c-core and one cube.

Due to this feature, along with the symmetry in the 3-D model, only one c-core with one rotor cube is enough to generate the entire machine's characteristics over the range from the unaligned to the fully aligned position.

The 3-D model is analyzed employing the MagNet version 7 2-D/3-D electromagnetic field simulation software from Infolytica Corporation. Fig.4-1(a) shows the 3-D model of the c-core wound by one phase coil, and a slice of the rotor disc carrying one rotor pole. It was built by sweeping the base lamination of the c-core with the same dimensions of that for the 2-D model. The material used is the non-oriented silicon steel (M19-29 gauge). The 3-D simulation process consumes significant computational time; therefore, special care should be taken when assigning the mesh to the model. Maximum element size (MES) is the tool used to refine the mesh only in particular areas such as the air-gap and poles. The 3-D motion models with transient analysis further increases the computational time cost. In order to reduce the model size and address the computational time concerns, a horizontal (XZ) plane is used to divide the c-core model into two symmetric halves, upper and lower parts, and then eliminate the lower one keeping only the smallest portion required for 3-D modeling. The fields will be everywhere normal to the symmetry plane; therefore, a field normal boundary condition must be assigned to this plane. In order to validate the latter process, both
Fig. 4-1. 3-D model. (a) One c-core wound by a phase coil and a slice rotor disc enclosing the rotor cube. (b) 3-D model solution for the flux density for 75 A at 62% alignment.

Complete and half c-core models were built and simulated under similar conditions and the results confirmed that they are exactly equivalent.

The coil of the model is to be energized by ideal flat-topped current over the whole stroke range. Fig. 4-1(b) presents the arrow and shaded plot for the flux density solution when the excitation current is 75 A and number of turns is 40 at 62% alignment. It clearly shows that the model can generate the needed MMF to saturate the areas of interest.

The SRM magnetization characteristics and the single-coil output torque calculated by the 2-D and 3-D models are presented in Figs. 4-2 and 4-3. The torque is calculated according to: \( \tau = F \times r \), where \( F \) is the reluctance force obtained from the FEA solution, and \( r \) is the distance between the rotor’s cube and the shaft. The noticeable discrepancies between the 2-D and 3-D models solutions affirm the need for 3-D modeling, especially at high excitation current levels.
Fig. 4-2. Comparison of the magnetizing characteristics of the proposed design at aligned and unaligned positions by 2-D and 3-D FEA models ($N=40$).

Fig. 4-3. Shaft torque by single c-core at two different current excitations calculated by 2-D and 3-D FEA models ($N=40$).

### 4.1.2 End Winding and Radial Flux Fringing in ASRM

In most conventional machines, the end winding portions usually constitute a small fraction of the whole winding of one pole depending on the geometrical ratio of the stator pole arc to the pole stack length, which usually ranges between the ratios 1:8 and 1:4. However, the c-core design at this stage has the ratio of 1:1”, due to the square cross-
section of its poles, as shown in Fig. 4-4. By referring to Fig. 4-2, the 3-D flux-linkage in the unaligned position has approximately twice the value of that obtained from the 2-D modeling. This is justified by realizing that the end winding portions are out of the 2-D model and it can only be modeled by the 3-D one.

4.1.3 C-core Geometrical Modifications for Diverting the Fringing Flux

The fringing flux at the four edges of one pole tooth can be divided into two sets; one of which is desired (tangential fringing flux), and can be seen on the side surface (S) of the c-core, as shown in Fig. 4-5(a). This flux will result in additional force in the tangential direction (the direction of the rotor motion). The other set of fringing flux can be seen on the front side (F) of the c-core. This set does not contribute to the tangential force. Instead, it will be deducted from the MMF causing a significant reduction in the output torque that is noticed in the 3-D results in Fig. 4-3. This flux is known as the radial fringing flux.
Various modifications on pole shapes were analyzed in an effort to reduce the radial flux fringing and, alternatively, divert it to be with the tangential fringing flux. Firstly, the improved pole shape, as shown in Fig. 4-6(a), is designed to minimize the radial fringing. A wedge shape is created on the edge of the front side (F), while the (S) side is kept flat. Furthermore, this design is modified to obtain the superior design by making the cross-section of the pole a rectangle instead of a square, in which the length of the rectangle is the side where the fringing flux is desired, as shown in Fig. 4-6(b). The analysis of the gain obtained from these changes is explained in the magnified drawing in Fig. 4-5(b). As we compare this modified shape with the original one, in Fig. 4-5(b) and Fig. 4-5(c) respectively, it can be concluded that the fringing flux in the modified shape would experience a longer path (more than the semi-circle that is in the original shape) and consequently have a higher reluctance. The equation that relates to the magnetomotive force and the reluctance is given by:

\[
F = NI \quad \Rightarrow \quad \phi = FR \tag{4.1}
\]

It can be observed from (4.1) that for a given value of MMF, if the reluctance is increased, the fringing flux will decrease which is in agreement with the FEA results. The improvement in the magnetization characteristics, as shown in Fig. 4-7, is significant in the aligned position where the saturation and fringing occur. This proves that the modifications have achieved the goal of reducing the undesired fringing (radial fringing flux). The improvements can also be seen in the comparison of the output torque for the original, improved and superior models, as depicted in Fig. 4-8.
Fig. 4-5. The proposed pole shape modification. (a) 3-D model of the c-core showing the flux fringing on the front and side surfaces. (b) The close-up of modified shape shows longer flux path. (c) The close-up of original shape shows shorter flux path.

Fig. 4-6. Modifications on the 3-D model. (a) The improved pole shape. (b) The superior design with rectangular cross-section of the c-core.
4.2 Analysis of the Operation of the Proposed ASRM Employing 3-D Dynamic FEA

For the design and development of SRMs and their drives, analysis and simulation of their dynamic behavior are essential. The dynamic performance of an SRM is described by the following equations:

\[
\begin{align*}
\frac{\partial \psi}{\partial t} &= v(t) - Ri(t) \\
\frac{\partial \theta}{\partial t} &= \omega \\
\frac{\partial \omega}{\partial t} &= \frac{T_d(\theta, i)}{J}
\end{align*}
\]  

(4.2)

where \( \theta \) is the rotor angular position, \( \omega \) is the angular speed, \( J \) is the moment of inertia, and \( T_d \) is the developed torque. As can be seen, the flux linkage \( \psi \), \( \theta \) and \( i \) are governed by a non-linear relationship due to the saturation in the magnetic circuit. Several non-linear SRM models have been developed and presented in literature [57], [58].

Fig. 4-7. The magnetization characteristics for the original design, the improved pole shape design, and the superior design (N=40).
Fig. 4-8. Shaft torque provided by single coil, calculated by 3-D FEA for the original pole geometry, the improved and the superior designs for two different current values.

FEM is found to be the optimum solution as it provides a precise solution through accurate computation of the magnetic field for a developed model. The switched reluctance motor operates in a series of strokes or transients and does not have steady state in which all its variables are constant. Therefore, it is reasonable to evaluate its dynamic operation by looking at individual strokes through 3-D FEA simulation. Here, 3-D FEA based motion model is employed at discrete speeds chosen from low, moderate and high speed regions in order to investigate the dynamic performance over the full speed range operation. This has been performed owing to the extremely large time consumption of 3-D FEA over the entire speed region.

It should be noted that the calculations, results and analyses presented are based on a single-coil computation. This is reasonable due to the complete separation of any c-core magnetic circuit from the other c-cores. The mutual inductances between adjacent phases or coils are negligible due to the sufficient spacing between them. The electric
excitation is provided to the c-core winding through single phase asymmetric bridge converter, as in Fig. 4-9 and is linked to the FEA model. The operation is divided into low speed region and high speed region. This separation is based on the method of control that needs to be applied to the motor.

4.2.1 Low Speed Range Operation

When the two power switches in Fig. 4-9 turn on, the full DC bus voltage is applied across the phase coil and the single phase voltage equation will govern the voltage, current and back-EMF as follows

\[ V = Ri + e \]  

(4.3)

where \( e \) is the back-EMF which is also defined as

\[ e = \frac{\partial \psi}{\partial t} = \omega_m \frac{\partial \psi}{\partial \theta} \]  

(4.4)

Accordingly, at low speeds, the back-EMF is small and most of the bus voltage appears in the first term of the right-hand side of (4.3). In other words, there will be

Fig. 4-9. Single phase asymmetric bridge converter is linked to the 3-D FEA model to energize one c-core coil.
sufficient voltage to allow for the current to rise up to the rated current and above. Furthermore, the current has to be regulated or limited to its rated value in order to generate the rated torque. The two main methods that can be applied to regulate the current are voltage PWM, and current hysteresis method. Fig. 4-10 shows the typical waveforms for both the current and voltage during one stroke of operation.

From the efficiency point of view, this region has two virtues: 1) the current can go to high levels and therefore the torque can reach the highest possible level, and 2) since the speed is low, there would be enough time for the current to rise up, reach the desirable level, and then fall down, all within the positive torque region (from the unaligned position to the fully aligned one). This will be reflected in a maximized co-energy area enclosed by the current trajectory in the magnetization characteristics. It can be seen in Fig. 4-11 that the current trajectory at the rated speed of 2,400 rpm encloses a larger area than those at higher speeds.

![Fig. 4-10. Voltage and current waveforms of a single coil, obtained at speed below the base speed. The current is regulated by current hysteresis method.](image)
4.2.2 Rated and High Speed Range Operation

Above the base speed, the back-EMF is increased until it becomes higher than the DC supply voltage, thus eliminating the need to limit the current. On the contrary, one needs to advance the switching in order to allow the current to reach its rated value before being limited by the back-EMF that tends to increase significantly with the increase in the overlapping area between the poles. Fig. 4-12 shows the current traces for different speeds obtained by advancing the firing angle for increasing the speed as shown in Table 4-1. Although the current is inherently limited, in the case of rated and high speed operations, it exponentially increases for speeds lower than the rated value, thus requiring a precise current regulation.

Fig. 4-11. The flux linkage trajectories against current obtained by 3-D FEA models under transient (motion) analysis, for three different speeds.
Table 4-1. Firing Angle Look-up Table for Low and High Speed Ranges.

<table>
<thead>
<tr>
<th>Speed (rpm)</th>
<th>1,200</th>
<th>1,800</th>
<th>2,400</th>
<th>3,000</th>
<th>3,600</th>
<th>4,200</th>
</tr>
</thead>
<tbody>
<tr>
<td>Turn on (mm)</td>
<td>-24</td>
<td>-27</td>
<td>-30</td>
<td>-33</td>
<td>-35</td>
<td>-38</td>
</tr>
<tr>
<td>Turn off (mm)</td>
<td>-6</td>
<td>-10</td>
<td>-12</td>
<td>-13</td>
<td>-14</td>
<td>-14</td>
</tr>
</tbody>
</table>

Fig. 4-12. Set of current traces obtained by varying the firing angle positions in accordance with different motor speeds.

4.3 Motor Performance Characteristics

The output torque observed at the shaft while running at rated speed for three consecutive phases is simulated in the 3-D motion model and the results are shown in Fig. 4-13. Based on the standard definition of torque ripple as in (3.1), the observed motor torque ripple is approximately 15% which is considered sufficient results, since no control technique has been used, whereas the current can be boosted in regions of low torque. With basic conventional control, this torque ripple can be brought to a negligible value.
The solution of the 3-D FEA model as shown in Fig. 4-14 provides the forces along the three coordinate axes. The radial force corresponds to the X-axis and shows negligible values. For the purpose of fair comparison with the radial SRM configuration, one would have to consider comparing the axial force of the axial SRM versus the radial force of the radial SRM. This comparison is provided in Fig. 4-15. Also, for proper comparison, the three models are current driven with the same value, and the design models are scaled in order to make all related parameters and dimensions identical, as possible. The design parameters and dimensions, such as the axial and radial dimensions and air-gap length, are identical and presented in Chapter 2. While the c-core model is scaled to equivalent values like the other two models, as close as possible. It is noticeable that the proposed ASRM has significantly lower axial forces (approximately 10 times lower) compared to the radial forces in radial SRM, leading to lesser acoustic noise, thus validating one of the intended purposes of the novel design.

Lastly, the ASRM ratings that were obtained by 3-D FEA simulation under multiple speed operation are summarized in Table 5-1. The predicted dynamic performance is described in the motor characteristics in Fig. 4-16.

Fig. 4-13. The output torque profile for three consecutive phases showing the individual contribution from each phase.
Fig. 4-14. 3-D FEA model solution for the purpose of obtaining forces along the X-Y-Z axes. Due to the symmetry in the 3-D model, only one half of the model is sufficient representation of one complete c-core.

Fig. 4-15. Comparison of the radial forces of the in-wheel radial SRM, the conventional radial SRM, and the axial force of the axial SRM.
Fig. 4-16. The ASRM power and torque-speed characteristics.
5. ASRM PROTOTYPE DEVELOPMENT AND TESTING

5.1 Development of the ASRM Prototype

Prior to the prototype hardware implementation, several design optimization and improvement stages have been completed. Comprehensive static and dynamic analyses in 2-D and 3-D environments have been performed. From a construction point of view, the motor prototype consists of the following components:

1. The c-cores and the cubes: In accordance with the c-core superior design model in Fig. 4-6(b), a prototype that has the same dimensions with minor modifications is built and shown in Figs. 5-1 and 5-2. Fully processed non-oriented silicon steel laminations, M19-29, are used. The laminations are stacked together in the z-axis (tangential direction). In order to keep the thin laminations together and to suppress any harmonic-base vibration among the steel sheets themselves, several laser welding locations are distributed evenly around the volume of the c-core as shown in Fig. 5-2(b). Evidently, there is no welding on the internal edges of the c-core due to manufacturing capability constraints. However, the welding distribution on the outer sides has achieved the desired mechanical strength. It should also be noted that the welding locations were chosen to be as far as possible from the regions of high flux density, which can be seen in the FEA-flux density solution in Fig. 5-1(a).
Fig. 5-1. C-core development. (a) Final version of the 3-D FEA model showing the shaded plot of the magnetic flux density of single c-core at full alignment. (b) The corresponding prototype during winding process.

Fig. 5-2. Prototype core components. (a) Proposed c-cores arrangement of the novel ASRM design. (b) An individual c-core with dimensions. Welding locations are marked by arrows.
2. The winding coils: This design has five phases, three repetitions, and concentrated winding around each of the 15 c-cores. The number of turns, strands, and conductor diameter has been determined according to the coil design algorithm that was explained previously in chapter 3, and based on the current constraint and voltage source availability \(N\) is found to be 60. The actual measured area of the cross-section of the coil is found to be 1,227 mm\(^2\) which results in a high practical winding fill factor of 40.9\%. This is due to the special geometric shape that allows for individual winding of each coil prior to its final assembly into the housing of the motor. Figures 5-1 and 5-6(a) demonstrate the availability of large room for extra winding around the c-core which will reduce the copper loss in the machine unlike other conventional machines where the amount of winding is limited by the slot space. Copper loss is the principle loss in SRM, and hence the overall efficiency can be optimized as efficiency improvement remains a major bottleneck in electric vehicle drivetrain. The resistance of the individual winding has been calculated depending on the conductor length and area. Also, the actual resistance has been measured using a micro ohm meter and both the calculated and measured results are provided in Table 5-1. Talking about the fault tolerance capability, this machine can withstand any fault in the c-core winding since the cores are wound individually, and hence will remain operational until the c-cores are re-placed. These significant characteristics recommend the motor to top the list of electric machines used in the drivetrain.

3. The rotor, which has a disc shape, is built using aluminum material (Al 6061-60T). The slot in the aluminum rotor disc was designed to be a bit larger than the length of
the cube. Once the cube was inserted into the slot, it was pushed towards the circumference of the disc and firmly fixed by the two locks as shown in Fig 5-3.

Table 5-1. RATINGS OF THE PROPOSED ASRM.

<table>
<thead>
<tr>
<th></th>
<th>Output power</th>
<th>Base speed</th>
<th>DC voltage</th>
<th>Maximum speed</th>
<th>Rated current</th>
<th>Maximum torque</th>
<th>Rated torque</th>
<th>Maximum power</th>
<th>Calculated single coil resistance</th>
<th>Measured single coil resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>9 kW</td>
<td>2,400 rpm</td>
<td>140 V</td>
<td>5,500 rpm</td>
<td>80 A</td>
<td>56 N.m</td>
<td>37 N.m</td>
<td>14 kW</td>
<td>29.6 mΩ</td>
<td>31 mΩ</td>
</tr>
</tbody>
</table>

Fig. 5-3. Rotor construction and assembly. (a) Isometric view of the final assembly. (b) The cross-section view of the rotor disc
In order to ensure safe operation at high speeds, a mechanical strength study is essential for the body of this disc at the critical regions, specifically at the circumference, outward of the rotor cube as shown in Fig. 5-4. The study objective is to verify that the direct shear stress $\tau$, at $A$, which is caused by the centrifugal force of the rotor cube during high speed operation, does not exceed the Ultimate Shear Strength ($S_{uu}$) of the aluminum material. The direct shear stress is given by:

$$\tau = \frac{F}{2A}$$

where $A=ht$. The “2” in the denominator indicates the double shear that exists. The centrifugal force at the cube centre relates the cube mass, angular speed, and rotor radius as follows:

$$F=mw^2r.$$  

(5.2)

The calculation results show that for the specific dimensions and for speeds up to 6,000 rpm, the direct shear was (31.8 MPa) which is significantly below the ultimate shear strength of (207 MPa).

4. The housing is built from the same aluminum material that was used in constructing the rotor disc. The slots provided by the housing sides and segments restrain the c-
cores from delocalization in the tangential axis against any type of vibration. Additional moveable locking wedges are designed to prevent delocalization in the radial axis and to allow for quick and easy replacement of any c-core in case of a damaged coil. Two pairs of tapered roller bearings are used to accommodate the axial thrust in either direction, as well as the radial loads. This design allows the motor to maintain a constant air-gap in spite of vehicular vibrations.

The Mechanical CAD drawing showing all components of the motor is presented in Fig. 5-5. The final motor assembly is shown in Fig. 5-6(a). In the present work, the prototype has been subjected to static testing in order to measure the magnetizing characteristics, machine parameters and to verify the correctness of the design.

5.2 Experimental Determination of the Static $\psi$-i Characteristics of the Developed Prototype

The magnetization characteristics are considered as the fingerprints of SRMs as they provide the essential information required to predict the behavior of these machines. Thus, we obtained these characteristics experimentally by holding the rotor at the two extreme positions; the aligned and unaligned ones. An indirect method, as in [59], is used for this purpose. The experimental setup is shown in Fig. 5-6(b). An isolated channels oscilloscope records the voltage and current data over the time frame that begins at the instant of applying voltage across the winding and terminates once the current reaches steady state. These data is transferred to a PC where it will be employed in performing the integration as follows:
Fig. 5-5. The Mechanical CAD drawing of the assembled prototype (dimensions are in mm). (a) Side view. (b) Front view.

Fig. 5-6. Prototype development. (a) Final assembly showing one c-core removed. (b) Experimental setup for obtaining the magnetization characteristics.
\[ \psi = \int \{v(t) - Ri(t)\} dt. \] (5.3)

The experimentally obtained magnetization characteristics along with 3-D FEA model results are presented in Fig. 5-7. The discrepancy in the unaligned position is more at the higher level of current. The main cause of this deviation is the flux leakage which is usually higher for high excitation current levels. While, the deviation in aligned position can be attributed to the uncertainties in the \( B-H \) characteristics of the core material employed while manufacturing and the subtle deviation of the air-gap length in the manufactured prototype, compared to the value taken in the 3-D FEA simulations.

Comparison of the calculated inductances through analytical equations (3.28) and (3.29), 2D FEA and 3D FEA for both aligned and unaligned cases have been summarized in Table 5-2 along with the error in percentage provided between parentheses.

<table>
<thead>
<tr>
<th>Method</th>
<th>Flux tube</th>
<th>2-D FEA</th>
<th>3-D FEA</th>
<th>Experiment</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L_u ) [mH]</td>
<td>0.9 (18%)</td>
<td>0.48 (56%)</td>
<td>1 (9%)</td>
<td>1.1</td>
</tr>
<tr>
<td>( L_a ) [mH]</td>
<td>2.6 (8%)</td>
<td>2.3 (4%)</td>
<td>2.5 (4%)</td>
<td>2.4</td>
</tr>
</tbody>
</table>

Fig. 5-7. Comparative analysis of the magnetization characteristics as obtained from experiments and 3-D FEA simulations (N=60).
5.3 Heat Run Test Performed on the Designed Prototype

The objective behind this test is to verify that at the rated power loss, the temperature rise of the machine is within limit, as dictated by the class of insulating materials used. No standard procedures for heat run test on an SRM exist at present, to the best knowledge of the authors. In order to perform the heat run test on an electrical machine in general, the rated voltage is applied and the machine is loaded to ensure rated current. This ensures the rated loss (core plus copper losses) to occur inside the machine on a continuous basis and the temperature rise at steady state is noted or calculated. Borrowing this idea, since the drive for this prototype is still in the process of being developed, a DC voltage source has been used to supply a continuous DC current through the winding that is higher than the rated current by about 25%. This additional amount of current has been deliberately passed on a continuous basis to take care of the heat that would have been generated due to the iron loss at rated voltage and other stray losses. For

Fig. 5-8. Heat run test setup; deep-cycle lead-acid batteries supplying 40 A to the prototype winding, thermometer is showing 45°C after 30 minutes.
Fig. 5-9. Experimentally-measured temperature rise of the prototype winding starting from the ambient temperature, and recorded every minute till thermal equilibrium is reached. The excitation current is 40 A.

For this purpose, all the c-core coils are connected in series to a deep cycle battery, as shown in Fig. 5-8. The steel cubes are removed to eliminate any undesired or harmful vibrations during the long test process. A thermocouple is inserted near the surface of one coil, and the temperature is recorded by a Fluke 89IV meter. Fig. 5-9 shows the experimental temperature-time curve, wherefrom the steady state temperature has been found to be about 47°C. The hot spot temperature of the winding is estimated to be 20°C higher than the measured one. As the insulating materials used for building the prototype belonged to Class B, the entire design is thus validated. The reserve thermal capacity of this design is keeping in view the short time overloads that this SRM is expected to undergo under fast acceleration/deceleration for an electric vehicle application.
5.4 Experimental Torque/Current vs. Rotor Position Angle Characteristics

Experiments were performed using the proposed ASRM coupled to a programmable loading machine through a torque transducer. The ASRM was run at the rated speed and the motor was tested by applying one pulse to a single coil. This test serves as an initial step towards the drive development which is ongoing. The current was limited by the back emf, since the machine was run at rated speed. The experimental setup and the measured results of torque/current are shown in Figs. 5-10 and 5-11 respectively. It can be noticed from the figures that the current is plummeting down due to the low inductance of the coils. Thus, the results are found to be satisfactory when compared to that of the FEA.

Fig. 5-10. Experimental setup of the ASRM coupled to the loading machine.
Fig. 5-11. Experimentally-measured shaft torque and corresponding current provided by single coil.

5.5 Conclusion

The motor prototype construction was explained in this chapter. A pair of tapered roller bearings was necessary to accommodate the axial thrust in either direction, as well as the radial loads. The thermal design testing showed steady state temperature of about 48°C. The hot spot temperature of the winding is estimated to be 20°C higher than the measured one, hence class B insulating materials is used. Lastly, the experimental torque measurement, contributed by only one c-core, comes in sufficient agreement with the ASRM ratings that were obtained by 3-D FEA simulation under multiple speed operation, since the FEA results was for the whole motor which energizes three c-cores coils at a time (the motor has a repetition of three).
Chapter 6

6. NOVEL APPROACHES TOWARDS LEAKAGE FLUX REDUCTION IN AXIAL FLUX SWITCHED RELUCTANCE MACHINES

The finite element analysis performed on the axial flux SRM in chapter 3 showed discrepancy between its 2-D and 3-D results in the unaligned position. Experimental investigations performed on the developed prototype have shown a relatively low inductance ratio. Furthermore, it is verified that the unaligned inductance is high which causes the low inductance ratio [60] that leads to a low torque to ampere ratio in axial flux SRM. Hence, in this chapter, firstly, a flux tubing model along with a flux distribution plot obtained by 3-D FEM was employed to analyze the factors that are causing the poor inductance ratio. The above issues have been thoroughly studied and three different approaches have been proposed in this chapter in order to reduce the leakage flux and, therefore, improve the inductance ratio. One approach is related to modification in winding, the second approach suggests the utilization of grain oriented steel segments and the third one proposes a novel method for using permanent magnet material as a shield to the leakage field. Each of these approaches are modeled and analyzed in detail through 3-D FEA and the results are discussed.
6.1 Investigation of Leakage Flux in Axial Flux SRM through Flux Tubing/Magnetic Equivalent Circuit

The torque producing capability of an SRM mainly depends on the rate of change of the inductance vs. rotor position. An indicator for that is the inductance ratio $La/Lu$, where $La$ and $Lu$ are the aligned and unaligned inductances respectively. The accurate calculation of these inductances has always been an essential step in the SRM design process. A lumped-parameter magnetic equivalent circuit (MEC) has been developed in pervious sections, where the magnetic circuit is represented by a lumped parameter electrical network. Since the motor under study has the magnetic circuit (c-core and cube) for each phase separated from the other phases, the author would like to state that the following investigation will only consider a single c-core along with a cube, as the rotor pole.

The flux tube technique was used to derive the reluctances in the air-gap region. Fig. 6-1 shows the MEC model for a single c-core and two cubes in a full unaligned position. For the unaligned position case, and due to the large airgap, the c-core material will not saturate, hence its reluctances can be neglected. $R_{c\text{-}back} = R_{p1} = R_{p2} = R_{c1} = R_{c2} = 0$. While, all air gap reluctances are large, as always. As a result, the leakage reluctance dominates in the unaligned position. On average, this leakage constitutes about 25% of the total flux. This is also verified by 3-D FEA simulation [60]. Simulation for the same c-core under exact conditions except the volume of the air box was repeated. The findings
prove that as the air box volume grows, the leakage flux increases, to a certain limit. This validates our analysis, since most of the leakage occurs in the air surrounding the coil.

6.2 Proposed Leakage Flux Reduction Methods

6.2.1 Reduction through modified winding configuration

Taking the coil cross section, $A_c$, as the flux tube cross section, and the coil length, $l_c$, as the flux path length, as shown in Fig. 6-2(a), the leakage flux reluctance can then be given as:

$$ R_l = \frac{l_c}{\mu_0 A_c} \quad (6.1) $$

From (6.1) it is observed that in order to increase the leakage reluctance, we should either increase the length of the coil, reduce the coil cross section, or both. However, the first option is limited by the length of the c-core itself. And the second option conflicts with the desire to have more copper, so that higher current can be allowed through the winding, and hence higher torque can be achieved. An optimized case is proposed and
Fig. 6-2. A new winding method. (a) Coil dimensions. (b) Conventional winding wire vs. square cross-section winding wire for optimum fill factor.

tested by using square single strand winding wire. As can be seen in Fig. 6-2(b), the utilization of square cross-section magnetic wire significantly increases the winding fill factor; consequently, cross-sectional area of the coil is reduced for the same amount of copper, and hence, the leakage reluctance increases. Fig. 6-3 shows a test structure constraining a c-core, with the coil cross section area smaller than that of the original coil by 25%. The c-core magnetization characteristics are obtained experimentally by holding the rotor cube at two extreme positions; the aligned and unaligned ones. The method used in [59] is adopted here for this purpose. An isolated channels oscilloscope records the voltage and current data over the time frame that begins at the instant of applying voltage across the winding and terminates once the current reaches steady state. This data is transferred to a PC where it will be employed in performing the integration as follows:

\[
\psi = \int \left[ v(t) - Ri(t) \right] dt.
\]  

(6.2)
Fig. 6-3. Test structure constraining a c-core with the coil cross section area smaller than that of the original coil by 25%.

![Image](image.png)

Fig. 6-4. Comparative analysis of the magnetization characteristics of a single c-core with special winding configuration as obtained from experiments and 3-D FEA simulations.

The aligned and unaligned flux linkage is measured and the saturation characteristic is provided in Fig. 6-4. The noticeable reduction of the unaligned flux linkage validates this proposed method. This winding configuration seems to improve the
reduction in leakage flux by a factor of 5% when compared to the original winding configuration.

6.2.2 Reduction through segmented grain oriented steel core

Another approach to minimize the leakage flux can be achieved by selecting the core material that has some control on direction of flux lines inside it. Such material is called Grain-Oriented (GO) electrical steel. In general, GO electrical steel has been limited to transformers and high power rotating machines, while Non Grain-Oriented (NGO) steel to rotating machines. That is because the GO steel is highly anisotropic; it has an axis of easy magnetization and low power loss \([001]\), coincided with the rolling direction, and an axis of hard magnetization and high loss \([111]\) [61].

The c-core can be divided into three segments as shown in Fig. 6-5. The middle segment is particularly important for reducing the leakage flux. Its orientation along the coil axis increases the core permeability along the same axis, and constrains the flux lines in one direction. The other two segments are useful to curve the flux lines in the desired

---

Fig. 6-5. Orientations of the segmented grain oriented c-core used in the design.
direction. These segments’ configuration increases the overall permeability in the core and hence reduces the losses. Also, it is particularly suitable for this topology of axial SRM since its magnetic circuits are isolated from each other as explained before, and the coil’s excitation current is always unidirectional. 3-D FEA results of a segmented c-core model that was built using grain oriented silicon steel, Alphasil 35 M6, have showed slight improvement in term of reducing leakage flux as well as improving the inductance ratio, as can be seen in Fig. 6-9.

6.2.3 Inductance ratio improvement through novel shielding strategy using permanent magnet material

The general definition of electromagnetic shielding revolves around the techniques of reducing the electromagnetic field in a space by blocking the field with barriers made of conductive or magnetic materials. Electromagnetic shielding has been used in a very wide range of different applications. Their major function was either to protect the devices or the human beings that are in proximity to a source of a magnetic radiation. Three main techniques of shielding have been well documented in literature: 1) passive shielding with high permeability materials or superconducting sheets [62]; 2) active shielding with compensation coils which produce a cancelling field; and 3) a combination of the above.

Electromagnetic shielding has rarely been used in electric machines. The Authors in [63] have proposed active shielding for leakage flux reduction in a transformer where the shielding is provided by two compensating coils being wound on the same core as for
the primary and secondary coils, but no work has been done to study the effects of electromagnetic shielding in electric motors.

As the leakage magnetic fields contribute to several undesirable effects in many applications [64], and more specifically in axial flux SRM [60], there is considerable interest in reducing such fields [65]. In this section the author proposes a novel concept in which permanent magnet material is used to provide a field that opposes the leakage fields, and hence reduces the leakage flux significantly. Although the concept is being proposed here for reducing the leakage flux in SRM, the concept can be applied in different kinds of machines wherever it is feasibly justified.

By referring to the flux tubing model and the MEC in Fig. 6-1, it can be deduced that in order to reduce the leakage flux, the branch that has $\mathcal{R}_l$ should contain a magnetomotive force (mmf) in the opposite direction to that of the original mmf generated by the coil, so that the flux in that branch becomes zero. The FEA field solution in Fig. 6-8(a) shows that most of leakage occurs near the two ends of the coil, and that any leakage from the side of the coil can be neglected. Therefore, it is suggested to design the leakage shield at these two ends. Two permanent magnets (PMs) are designed to suppress the leakage flux at both ends of a c-core coil as shown in Fig. 6-8(b).

These magnetic shields should be capable of at least producing equivalent magnetic field to the leakage field, and in the opposite direction. These magnets were initially designed depending on the estimated value of the leakage field by using the
developed mathematical model based on Biot-Savart law. Thereafter, the magnet design was optimized after several iterations through FEA.

6.2.3.1 Mathematical model of the C-core end winding flux:

Based on Biot-Savart law, and assuming a square cross section core, and a four layers coil of \( N \_L \) turn-levels along the Z-axis, for \( L=(1,2,3,\ldots,N) \), as demonstrated in Fig. 6-6, the magnetic flux density due to a finite length of wire at a point P is derived as follows:

\[
\vec{B} = \frac{\mu I}{4\pi} \int \frac{dl \times \vec{r}}{r^3}
\]

Where, \( dl = dx \hat{i} \) and \( \vec{r} = x \hat{i} + y \hat{j} \)

\[
\vec{B} = \frac{\mu I}{4\pi} \int \frac{ydx}{\left(x^2 + y^2\right)^{\frac{3}{2}}} (\hat{k}) = \frac{\mu I}{4\pi} \left(\cos \theta_2 + \cos \theta_1\right) \hat{k}
\]

Fig. 6-6. Isometric view of the square cross-section coil wound around the c-core.
In order to calculate the magnetic flux due to four layers, the defined angles $\theta_{n(1)}$ and $\theta_{n(2)}$ for $n=(1,2,3,4)$ and the displacement vector ($\vec{r}$) are incorporated according to Fig. 6-7.

For layer 1:

$$\vec{B}_{1,N_1} = \frac{\mu l}{4\pi y} (\cos \theta_{1(1)} + \cos \theta_{1(2)}) \hat{k}$$

$$\vec{B}_{1,N_1} = \frac{\mu l}{4\pi (y + a)} (\cos \theta_{1(1)} + \cos \theta_{1(2)}) (-\hat{k})$$
Due to the placement of the magnetic core, the flux density from the far side of coil is negligible.

The flux from sides 3 and 4 cancel each other: \( \vec{B}_{1(4,N_1)} + \vec{B}_{1(3,N_1)} = 0 \)

Similar analysis can be applied to layer 2, 3, and 4, resulting in the following,

\[
\vec{B}_{2(1,N_1)} = \frac{\mu I}{4\pi(y-4)}(\cos \theta_{2(1)} + \cos \theta_{2(2)})\hat{k}
\]

\[
\vec{B}_{3(1,N_1)} = \frac{\mu I}{4\pi(8-y)}(\cos \theta_{3(1)} + \cos \theta_{3(2)})\hat{k}
\]

\[
\vec{B}_{4(1,N_1)} = \frac{\mu I}{4\pi(12-y)}(\cos \theta_{4(1)} + \cos \theta_{4(2)})\hat{k}
\]

\[
\vec{B}_{(Total-at-P)N_1} = \vec{B}_{1(1,N_1)} + \vec{B}_{2(1,N_1)} + \vec{B}_{3(1,N_1)} + \vec{B}_{4(1,N_1)}
\]

For the levels of turns \( N_2, N_3 \ldots N_N \), the magnetic flux will have a component along X and Z axis, as can be seen in Fig. 6-6, and the displacement vector (\( \vec{r} \)) and magnetic flux calculation for second level-turn and first layer will be as follows,

\[
dl = dx\hat{i}
\]

\[
\vec{r}_2 = x\hat{i} + y\hat{j} + z_{N_2}\hat{k}
\]

\[
\vec{B}_{1(1,N_2)} = \frac{\mu I}{4\pi} \int \frac{dl \times \vec{r}}{r^3} = \frac{\mu I}{4\pi} \int \frac{(ydx)\hat{k} + (z_{N_2}dx)(-\hat{i})}{\left(x^2 + y^2 + z_{N_2}^2\right)^{3/2}} =
\]

\[
\frac{y}{y^2 + z_{N_2}^2} (\cos(\arctan\frac{\sqrt{y^2 + z_{N_2}^2}}{x_1}) + \cos(\arctan\frac{\sqrt{y^2 + z_{N_2}^2}}{x_2}))\hat{k} +
\]

\[
\frac{z_{N_2}}{y^2 + z_{N_2}^2} (\cos(\arctan\frac{\sqrt{y^2 + z_{N_2}^2}}{x_1}) + \cos(\arctan\frac{\sqrt{y^2 + z_{N_2}^2}}{x_2}))(-\hat{i})
\]

Similar procedures are followed for the second level-turn and second layer:
\[ \vec{B}_{2(1,N_2)} = \frac{\mu_0}{4\pi} \int \frac{\vec{d}l \times \vec{r}}{r^3} = \frac{\mu_0}{4\pi} \int \frac{((y-4)dx)\hat{k} + (z_{N_2}dx)(-\hat{i})}{\left(x^2 + (y-4)^2 + z_{N_2}^2\right)^{\frac{3}{2}}} \]

\[ \frac{(y-4)}{(y-4)^2 + z_{N_2}^2} \left(\cos(\arctan(\sqrt{(y-4)^2 + z_{N_2}^2})) + \cos(\arctan(\sqrt{x_1^2 + (y-4)^2 + z_{N_2}^2}))\right)\hat{k} + \]

\[ \frac{z_{N_2}}{(y-4)^2 + z_{N_2}^2} \left(\cos(\arctan(\sqrt{(y-4)^2 + z_{N_2}^2})) + \cos(\arctan(\sqrt{x_2^2 + (y-4)^2 + z_{N_2}^2}))\right)(-\hat{i}) \]

Similarly for second level-turn and third layer:

\[ \vec{B}_{3(1,N_3)} = \frac{\mu_0}{4\pi} \int \frac{\vec{d}l \times \vec{r}}{r^3} = \frac{\mu_0}{4\pi} \int \frac{((8-y)dx)\hat{k} + (z_{N_2}dx)(-\hat{i})}{\left(x^2 + (8-y)^2 + z_{N_2}^2\right)^{\frac{3}{2}}} \]

\[ \frac{(8-y)}{(8-y)^2 + z_{N_2}^2} \left(\cos(\arctan(\sqrt{(8-y)^2 + z_{N_2}^2})) + \cos(\arctan(\sqrt{x_1^2 + (8-y)^2 + z_{N_2}^2}))\right)\hat{k} + \]

\[ \frac{z_{N_2}}{(8-y)^2 + z_{N_2}^2} \left(\cos(\arctan(\sqrt{(8-y)^2 + z_{N_2}^2})) + \cos(\arctan(\sqrt{x_2^2 + (8-y)^2 + z_{N_2}^2}))\right)(-\hat{i}) \]

For different number of level-turns, the Z axis component will be replaced in above equations. And displacement vectors for all turns will be as follows:

\[ \vec{r}_2 = x\hat{i} + y\hat{j} + z_{N_2}\hat{k} \]

\[ \vec{r}_3 = x\hat{i} + y\hat{j} + z_{N_3}\hat{k} \]

\[ \vec{r}_4 = x\hat{i} + y\hat{j} + z_{N_4}\hat{k} \]

\[ \vdots \]

\[ \vec{r}_N = x\hat{i} + y\hat{j} + z_{N_N}\hat{k} \]

And, the total magnetic flux at point P will be calculated as follows:

\[ \vec{B}_{(Total\text{-at\text{-point}\text{-P})}} = \sum_{i=1}^{N} \vec{B}_{(Total\text{-at\text{-point}\text{-P})Ni} = \sum_{i=1}^{N} \vec{B}_{1(1, Ni)} + \vec{B}_{2(1, Ni)} + \vec{B}_{3(1, Ni)} + \vec{B}_{4(1, Ni)} \]
The analytical approach along with the flux tubing model provides the preliminary value of the leakage field that needs to be suppressed by the magnets, as a shield. The end results of a design procedure for any PM are selecting the suitable magnetic material and dimensions of the magnets. The rule of thumb is that the magnetic material should tolerate the maximum flux density without losing its magnetism. To ensure proper selection, the operating point on the material demagnetizing curve should be studied. Since the magnets are located in a low permeability media (air), therefore, the magnet is operating at a flux below the remnant flux $\phi_r$, as the mmf drop across the large airgap is seen by the magnet as negative demagnetizing mmf. Additional demagnetizing mmf is affecting the magnet which is caused by the leakage field from the c-core. This addition drives the operating point more to the left of the material demagnetizing curve, toward the coercive mmf point, $F_c$. Therefore, Neodymium Iron Boron 28/23, (Nd$_2$Fe$_{14}$B), is chosen since it has a very high value of coercivity, ($H_c$=813kA/m). The coercivity is related to $F_c$ through the equation ($F_c = H_c \cdot T_M$), where $T_M$ is the thickness of the magnet. Hence, the dimensions of the magnet are designed accordingly. The two

![Fig. 6-8. Magnetic solutions of the C-core models with and without the magnetic shields.](image)
magnets have a flat ring shape covering the two ends of the c-core coil, with 5 mm thickness, and uniform magnetization in the axial direction.

3D FEA simulation is performed on the developed design. Fig. 6-8(a) shows the designed c-core without the magnetic shielding. Magnetic field solutions show the high leakage at both ends of the coil. Fig. 6-8(b) shows the c-core with magnetic shielding. Its magnetic field shown in Fig. 6-8(c) illustrates the reduction in leakage flux in the vicinity of the coil.

Fig. 6-9 shows the 3D FEA flux linkage characteristics of the axial-flux SRM for all three leakage flux reduction approaches proposed in this chapter. It is clear that a significant improvement in the values of unaligned inductance is achieved using the magnetic shielding approach. This improvement is attributed towards the reduction of leakage flux. The approaches utilizing grain oriented steel in the core and special winding configuration also leads to a smaller improvement in the aligned and unaligned inductance values as desired. Fig. 6-10 shows the corresponding 3D FEA torque profiles of each of the machine models including the approaches. The torque profile for the machine having a magnetic shielding seems to be bigger when compared to the original c-core design. This increment in the torque can be attributed towards the increased inductance ratio through magnetic shielding. The torque profiles of the other two approaches also have shown an increment compared to the original machine design however to a limited extent.
Fig. 6-9. Flux linkage profiles of different C-core designs incorporating the proposed leakage flux reduction approaches.

Fig. 6-10. Torque profiles of different C-core designs incorporating the proposed leakage flux reduction approaches.
7. CONCLUSION

- The comparative performance analysis performed in chapter 2 between the outer-rotor SRM and the conventional radial flux SRM shows a 10 percent improvement in the output torque of the short flux SRM, and a reduction of torque ripple. This reduction can be attributed to the increased number of poles and small step angle of 5 degrees which lead to a high overlap ratio of 50 percent. While the improvement in output torque can be attributed to the short-flux path configuration in the machine. However, the radial forces which accounts for the acoustic noise and vibration are still high.

Hence, a major design modification in SRM technology is necessary in order to keep the merits such as efficiency and reduced torque ripple and simultaneously overcome challenges such as acoustic noise and vibration in the machine. This is one of the key motivating points to investigate axial-flux SRMs for electrified vehicles.

- Chapters 3, 4, 5 presents the design procedures, improvements and analysis of a novel axial flux SRM. The new design aims to reduce radial forces and torque ripples without sacrificing the efficiency. The performed analysis based on key design aspects such the step angle, overlap ratio between adjacent phases, torque equation and radial forces promotes the motor to be a potential candidate for electrified vehicle applications. The methodologies used for deriving the output power equation based
on the motor parameters provide a platform for designing other geometries of axial
flux machines. The winding designing algorithm presented in this dissertation is
validated through 3-D FEA models, the number of turns is calculated to be 60, based
on the current constraint and voltage source availability. Several 2-D and 3-D FE
models are built, and the optimum design is selected based on maximum inductance
ratio achieved. The obtained FEA results necessitate the need for 3-D FEA, especially
in the case of axial-flux SRM, as excessive end winding flux and radial flux fringing
was only noticed in the 3-D FEA results. Moreover, this dissertation proposes a
modification in the ASRM pole shape to reduce the radial flux fringing and the results
are presented. The 3-D FE motion model at multiple speeds is examined, as part of
the dynamic performance analysis. The speed at which the flux linkage trajectory
maximizes the enclosed area is considered the base speed, which is found to be 2,400
rpm. The process of prototype development and the selection of all components are
explained. The results from FEA and from various tests performed on the prototype
satisfactorily validate the design. The analytical calculations employed show good
agreement with FEA’s results. The motor output torque obtained by the 3-D motion
model running at rated speed has shown torque ripple of approximately 15% which is
considered an improvement, since no control technique has been used, whereas the
current can be boosted in regions of low torque. The performed radial and axial forces
analysis has shown significant improvement. The axial force is found to be 10 times
lower than their counterpart in the radial SRM. And the radial forces in the ASRM are
negligible. Despite the mentioned validated advantages, the performed tests on the
ASRM have revealed low inductance ratio.
- In chapter 6, investigations are performed on the developed prototype to elicit the issue of low inductance ratio corroborated by developed flux tube and magnetic equivalent circuit models. Thereafter, the proposed approaches were tested by developing three different machine models with each of these approaches. A 5% improvement in inductance ratio was obtained using the special winding configuration when compared to the original winding design in the machine. Both segmented grain oriented steel core and special winding configuration exhibited improvement in the inductance values both in the aligned and unaligned positions in the machine. However, the magnetic shielding method seems to provide significant improvements in reduction of leakage flux in the unaligned position when compared with the other two approaches and the original machine design.
REFERENCES


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