Modeling and Analysis of High Torque Density Transverse Flux Machines for Direct-Drive Applications

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Modeling and Analysis of High Torque Density Transverse Flux Machines for Direct-Drive Applications

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Dissertation

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ABSTRACT

Commercially available permanent magnet synchronous machines (PMSM) typically use rare-earth-based permanent magnets (PM). However, volatility and uncertainty associated with the supply and cost of rare-earth magnets have caused a push for increased research into the development of non-rare-earth based PM machines and reluctance machines. Compared to other PMSM topologies, the Transverse Flux Machine (TFM) is a promising candidate to get higher torque densities at low speed for direct-drive applications, using non-rare-earth based PMs. The TFMs can be designed with a very small pole pitch which allows them to attain higher force density than conventional radial flux machines (RFM) and axial flux machines (AFM).

This dissertation presents the modeling, electromagnetic design, vibration analysis, and prototype development of a novel non-rare-earth based PM-TFM for a direct-drive wind turbine application. The proposed TFM addresses the issues of low power factor, cogging torque, and torque ripple during the electromagnetic design phase.

An improved Magnetic Equivalent Circuit (MEC) based analytical model was developed as an alternative to the time-consuming 3D Finite Element Analysis (FEA) for faster electromagnetic analysis of the TFM. The accuracy and reliability of the MEC model were verified, both with 3D-FEA and experimental results. The improved MEC model was integrated with a Particle Swarm Optimization (PSO) algorithm to further enhance the
capability of the analytical tool for performing rigorous optimization of performance-sensitive machine design parameters to extract the highest torque density for rated speed.

A novel concept of integrating the rotary transformer within the proposed TFM design was explored to completely eliminate the use of magnets from the TFM. While keeping the same machine envelope, and without changing the stator or rotor cores, the primary and secondary of a rotary transformer were embedded into the double-sided TFM. The proposed structure allowed for improved flux-weakening capabilities of the TFM for wide speed operations.

The electromagnetic design feature of stator pole shaping was used to address the issue of cogging torque and torque ripple in 3-phase TFM. The slant-pole tooth-face in the stator showed significant improvements in cogging torque and torque ripple performance during the 3-phase FEA analysis of the TFM. A detailed structural analysis for the proposed TFM was done prior to the prototype development to validate the structural integrity of the TFM design at rated and maximum speed operation. Vibration performance of the TFM was investigated to determine the structural performance of the TFM under resonance.

The prototype for the proposed TFM was developed at the Alternative Energy Laboratory of the University of Akron. The working prototype is a testament to the feasibility of developing and implementing the novel TFM design proposed in this research. Experiments were performed to validate the 3D-FEA electromagnetic and vibration performance result.
DEDICATION

I dedicate my work to my parents.
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CHAPTER I
INTRODUCTION

1.1 Direct-Drive Applications

Direct-drive electrical generators offer a reliable alternative to gearbox drivetrains. Elimination of the gearbox has several benefits such as reduced noise levels, simplification of the drive train that increases reliability, reduced losses due to the reduced steps in energy conversion and lower maintenance cost [1-2]. Costly gearbox maintenance issues that can cause long downtime periods like oil or gearbox replacement are avoided [3-5]. Furthermore, the decreased number of bearings and moving parts required for direct-drive systems provide a more reliable machine with increased lifetime [6].

Electrical excitation of the rotor poles of a direct-drive machine, brushless or not, can lead to resistive heat losses in the system. Due to these losses, a complicated cooling mechanism is required for heat dissipation and the overall system’s efficiency also decreases. A number of manufacturers have turned to machines with permanent magnets (PM) for their attractive characteristics, such as compact machine structure, the absence of electrical excitation in the rotor, and the increased efficiency [7].

In a Permanent Magnet Direct-Drive (PMDD) machine, the magnetic field in the rotor is provided by the PM material. This attribute eliminates the rotor excitation losses from the generator and decreases the heat developed in the system. The energy yield and the
overall efficiency are thus increased while the absence of slip rings increases the reliability of the machine. However, rare-earth PM materials are expensive and difficult to manufacture [8]. The stator of a PMDD generator is usually identical to that of an Electrically Excited Direct-Drive (EEDD) generator, but alternative stator topologies have been developed as well [9-10]. An efficient power converter along with the PMDD generator can be connected to the grid to produce a clean power output [11-12].

The rotor poles of a PMDD machine are made of rare-earth PM materials such as Samarium-Cobalt (SmCo) or Neodymium-Iron-Boron (NdFeB) that has high magnetic energy densities. SmCo magnets are mainly used in high-temperature applications. The work in [13] concluded that NdFeB magnets produce a greater remnant flux density (1.2T) and can reduce the overall mass and price of the PMDD generator even further. PM excitation produces a compact and reliable machine with superior efficiency and torque density when compared to EEDD machines. The advancement of PM materials and power converters during the last decade, encouraged a number of developers to turn to this excitation type for direct-drive machines [14-15]. Due to the rapid commercial and military interest that PM technology has experienced over the last decade, the industrial demand for high power PM machines has also increased. This can be also verified by the increasing number of PMDD manufacturers who have established the PM synchronous generator as a prime candidate for direct-drive wind turbine applications [16].

The design parameters, which affect the generator speed, are dependent on the power of the wind turbine. This is illustrated in Figs. 1.1 and 1.2, which show that as the power rating of the turbine increases, the shaft speed decreases. Thus, machines with high power output at low speeds are ideal for wind generator applications. Short pole-pitch electric
machines are capable of delivering high power and torque density at low speeds without using rare-earth magnets making transverse flux machines (TFM) an ideal candidate for direct-drive wind turbine application.

![Figure 1.1: Shaft speed of turbine for different power rating and wind speed.](image1)

![Figure 1.2: Rotor radius of the turbine for different power ratings and wind speed.](image2)

1.2 Classification of Permanent Magnet Machines

Brushless DC (BLDC), PM synchronous motors (PMSM), and PM step motors can all be categorized as brushless PM motors. They have identical operating principles and are used in different applications based on their performance requirement. The overall classification of electric motors is shown in Fig. 1.3, which puts into perspective, how all the motor technologies fit in the overall motor ecosystem. BLDC motors are characterized by their trapezoidal back-electromotive force (EMF) waveform, which can be driven by rectangular pulse currents. A PMSM is characterized by its sinusoidal back-EMF which can be driven by sinusoidal currents. Step motors require multiple periods of pulsating current excitation for one complete mechanical shaft revolution as they have a higher pole count, compared to BLDC and PMSM.
Several different PMDD topologies have been proposed in the literature in an attempt to produce machines with high torque density at low cost. They can be categorized based on the orientation of the magnetic flux crossing the air gap between the rotor and the stator, leading to radial flux (RF), axial flux (AF) and transverse flux (TF) topologies. Depending on the stator’s core design, the machine can be slotted or slotless [10]. Another way to categorize PMDD generators is by the existence of iron in the stator’s core or not, leading to iron-cored or “air-cored” machines [17-18].

1.2.1 Radial flux machines

The radial flux (RF) configuration is the most common configuration of PMSM. This type of machines is widely used in various applications, such as traction, ship propulsion systems, wind power generation, robotics, and many others [19-21]. The magnetic flux lines are in the radial plane, while the current flows in the axial direction. The stator of the RF PMSM is similar to that of a conventional AC machine, which makes it easier to build due to existing and well-proven stator manufacturing technology.
Depending on the way the PMs are placed in the rotor, there are several design possibilities. The most common rotor designs used in the RF-PMSM are the surface-mounted, inset, and buried [110].

Figure 1.4: Magnet configuration on rotors on radial flux machines.

Figure 1.4 shows a variety of the most common inner rotor types. Four of the rotors shown, Fig. 1.4 (a-d) illustrates variations of surface-mounted magnets (SPM). The traditional radial arc magnet shape is shown in Fig. 1.4(a). Figure 1.4(b) is similar to Fig. 1.4(a), except the sides of the magnet are parallel, rather than radial. Another iteration is shown in Fig. 1.4(c), which is often called bread loaf. A solid ring of magnet material as shown in Fig. 1.4(d) can also be designed where the magnet poles are generated by magnetizing the rotor after developing the rotor yoke. The remaining two rotor cross-sections in Fig. 1.4 show two common interior permanent magnet (IPM) rotors. The rotor is shown in Fig. 1.4(e) is known as the spoke configuration. This configuration promotes flux concentration because the magnet surface area is greater than the rotor surface area. This rotor type is useful for gaining better performance from low energy ferrite magnet material and has the benefit of using rectangular block magnets. The final rotor design
shown in Fig. 1.4(f), has buried magnets. This construction is beneficial for high-speed operation since the rectangular magnets are entirely enclosed in a solid rotor envelope. While the IPM rotors support the use of rectangular magnets, the presence of ferromagnetic material at the rotor surface dramatically increases the air gap inductance. Furthermore, it adds a reluctance component to the torque produced.

The RF machine with an iron-cored stator and SPM rotors the most common topology for PMDD generators for its structural stability and reliable design. The slotted RF PMDD machine is the most conventional one, as it integrates the structural characteristics of an EEDD machine with the advanced magnetic characteristics of the PM. The reduced mass for high power ratings has established RF machines as the most common option for industrial PMDD generators for wind applications [2, 11, 12, 14, and 17].

1.2.2 Axial flux machines

The axial flux (AF) configuration has a similar electromagnetic design to its RF counterpart with cylindrical rotors but the thermal design and structural assembly are more complex. The AF PMSM is also referred as the disc-type machine due to its pancake shape rotor and has attractive features like compact construction, and higher power density, compared to RF PMSM. AF PMSM is particularly suitable for electrical vehicles, pumps, fans, valve control, centrifuges, machine tools, robots and industrial equipment [21]. AF PMSM can be designed with a large number of pole for a given diameter, making it ideal for ideal for low-speed applications, such as traction drives, small to medium power wind generators. The unique disc-type rotor profile and several stator core topologies allow for developing diverse and interchangeable AF PMSM designs. AF PMSM can be designed as single or dual air gaps machines, with slotted, slotless or even completely ironless stator.
SPM rotor and slotless stator topology are favored for low power AFPM machines. As the output power requirement increases, the area of contact between the rotor and the shaft in proportion to the power becomes smaller [21]. Careful attention is required for the design of the rotor-shaft mechanical joint as this is usually the prime cause of failure of disc type machines [21].

Development of new PM and lamination materials, improvements in manufacturing technology and cooling techniques, resulted in further increase in the power density for conventional PM electrical machines. However, there is a limit to this improvement for conventional RFPM machines because of [19-23]:

- The bottle-neck feature for the flux path from the rotor to the stator limits the scope for increasing the magnetic loading for a given machine outer diameter.
- Most of the rotor yoke does not contribute to the active magnetic circuit.
- Heat from the stator winding is radiated to the stator core and then to the frame which is not enough for high current density designs without forced cooling arrangements.

These limitations are inherent to the RF structural geometry and cannot be mitigated. The AFPM machine is recognized as having a higher power density and better thermal cooling ability than the RFPM machine [19-22]. In general, the special properties of AFPM machines, which are considered advantageous over RFPM machines in certain applications, can be summarized as follows [20-21]:

- AFPM machines have a much larger diameter to stack length ratio.
- AFPM machines have a planar and adjustable air gap.
- AFPM machines can be designed to possess a higher power density with reduced core material.
• The topology of an AFPM machine is modular where the number of the modules can be adjusted to meet the power or torque requirements.

• The larger the outer diameter of the core, the higher the number of poles that can be accommodated, making the AFPM machines a suitable choice for high frequency or low-speed operations.

Several publications made quantitative investigations of RFPM and AFPM machine configurations in terms of sizing and power density equations [20, 24-27]. The performance comparison between a conventional RF PMSM and a number of AF PMSM with different topologies at five different power levels is given in [28], which concluded that the AFPM machine can be designed with a smaller volume and less active material mass for a given power rating than the RFPM machine.

1.2.3 Transverse flux machines

The major difference between transverse flux machine (TFM) and RFM and AFM is that the TFM allows an increase of the space for the armature windings without decreasing the available space for the main magnetic flux [38]. The TFM can be also made with a very small pole pitch compared with other machines [9]. This feature of TFM allows it to have higher force density than RFM and AFM. The copper winding of the TFM is simple, and the end winding is significantly smaller than on other machines. Thus, the active mass of the TFM machine required to produce the desired torque can be much smaller than that of other machines. In other words, TFM can attain a higher torque density than both RFM and AFM. The disadvantages associated with TFM includes a complicated construction assembly and low power factor. In [29], through extended comparison among a number of suggested direct-drive technologies, it was concluded that the iron-cored PM TFM
topology offered the greatest potential in terms of power density and cost/torque ratio. A fundamental issue for PM TF generators is their low power factor (typical between 0.35 for surface and 0.55 for buried magnets) due to large armature leakage fields [30]. This can be overcome by using an appropriately rated power converter and active control. In [31], a DSP controller board was used for each phase converter to optimize the power factor of a C-core TF PMDD generator with flux-concentration using a normalized open-circuit voltage as a current reference signal, giving a power factor of one. Other optimization methods for the power factor include magneto-static and transient 3D-finite element analysis (FEA) for obtaining the best magnetic circuit to minimize the leakage paths of the machine [32]. Overall, the PM TFM was concluded to be a good fit for direct-drive applications due of its high specific torque compared to RF and AF PMDD machines, even though it has a complex manufacturing process and assembly along with low power factor [33].

1.3 Research Motivation

Direct-drive machines are preferable in wind generator applications due to increased efficiency and system reliability. The efficiency and reliability are further increased by mitigating failures in the gearbox and lowering maintenance problems. Due to the direct-drive configuration, the machine needs to operate at low speeds while producing high torque output. Furthermore, direct-drive generators with rare-earth PM excitations offer an added advantage of eliminating excitation losses and a reduction in the active weight of the machine. State-of-the-art, high-efficiency, high power density electric machines typically utilize rare-earth PMs which is confined within a small region of the World, resulting in a very volatile price for this commodity. Due to the six-fold increase in the price of NdFeB,
the production cost of a Toyota Prius permanent magnet electric machine has increased from $200 to $600 from 2006 to 2011, and the entire increase in cost is attributed to the price increase in the NdFeB magnets. PM wind generators are successfully competing with doubly-fed induction generators (DFIG) for their high torque and power density and superior control features. However, the amount of magnet required in wind generators is much more than in electric vehicles, and it has become more than essential to find alternatives to rare-earth PM based electric machines for this application.

1.4 Research Objective

The objective of the proposed research is to develop a novel, modular, high torque density, rare-earth PM free machine for direct-drive wind turbine application using the concept of transverse flux (TF) paths. The proposed PMTFM would be based on an innovative pole configuration around the ring windings of a transverse flux path that significantly reduces the leakage flux while providing modularity and robustness in a manufacturable structure. The flux focusing technique would be used in the rotor to allow the use of non-rare-earth PM materials to overcome the cost and availability issues of rare-earth PM. The proposed research aims to design and optimize the transverse flux concept through electromagnetic, structural and thermal studies, and then fabricate a 1 kW (peak) prototype machine for experimental analysis. The principles and techniques developed through this research for TFMs will also be applicable for various other applications, such as in automotive and aerospace systems.
1.5 Dissertation Outline

This dissertation presents the modeling, design and prototype development of a novel non-rare-earth based PM-TFM, designed for direct-drive wind turbine application. The proposed TFM addresses the issues of low power factor, cogging torque, and torque ripple during the electromagnetic design phase.

Chapter 2 provides a literature review on the existing TFM topologies highlighting their advantages and disadvantages. The critical TFM characteristics are explained in detail and some recently developed TFMs are also critiqued. Based on the literature evaluation, a novel TFM topology is proposed.

Chapter 3 lays out the initial design procedure for the TFM proposed in Chapter 2. The design consideration based on the application requirement, as well as lab testing and prototyping limitations is discussed. The critical machine parameters required for optimization were identified and certain design parameters are optimized.

In Chapter 4, an improved Magnetic Equivalent Circuit (MEC) based analytical model was developed as an alternative to the time-consuming 3D-FEA for faster electromagnetic analysis of the TFM. The MEC model, developed with detailed modeling for air gap and stator core leakage flux path, could also capture the effect of magnetic saturation. The accuracy and reliability of the MEC model were verified with 3D-FEA.

In Chapter 5, the improved MEC model was integrated with the Particle Swarm Optimization (PSO) algorithm to further enhance the capability of the analytical tool for performing rigorous optimization with performance-sensitive machine design parameters to extract the highest torque density for rated speed. The overall effectiveness of the analytical model-based optimization was compared using 3D-FEA based optimization, and
it was found that the analytical model leads to the design of a similar sized machine with drastically reduced computation time.

A novel concept of integrating the rotary transformer within the proposed TFM design was explored in Chapter 6 that would completely eliminate the use of magnets from the TFM. The proposed structure allowed for improved flux-weakening capabilities of the TFM for wide speed operations.

The electromagnetic design feature of stator pole shaping was used to address the issue of cogging torque and torque ripple in the 3-phase TFM in Chapter 7. The slant-pole tooth-face in the stator showed significant improvements in cogging torque and torque ripple performance during the 3-phase FEA analysis of the TFM. Vibration response under resonance for the baseline TFM and the TFM with slant-pole was also investigated.

The TFM prototype was developed at the Alternative Energy Laboratory of the University of Akron. Detailed mechanical prototyping procedure and experimental analysis for the proposed TFM are presented in Chapter 8.

Finally, the research contributions and scope for future work with this proposed TFM is discussed in Chapter 9.
CHAPTER II
TRANSVERSE FLUX MACHINES

2.1 Introduction

This chapter will discuss the general characteristics of the Transverse Flux Machine (TFM). The most common types of generic TFM configurations, the working principle and other challenging aspects of the TFM will be discussed.

The concept of TFM was first mentioned in 1895 [34] but did not attract much research interest at that time. In 1970’s, some papers on a linear variant of the TFM for railway motored vehicles application [35, 36] were published, which gave a new life to the concept. In the mid-80s, the publication of [24] kick-started the development of different types of TFMs.

2.2 TFM Working Principle

The geometry depicted in Fig. 2.1 is a basic arrangement referred to as single-sided TFM [37]. The geometry depicted in Fig. 2.1 is a basic arrangement referred to as single-sided TFM [37]. It consists of a rotor yoke with SPM rotor poles. The stator is made up of C-shaped iron cores of laminated steel. The winding is placed in the stator slots. The magnets are magnetized with alternating polarity, thereby producing an alternating flux in the stator. The toroidal winding links the fluxes produced by each pole pair.
Figure 2.1: Single-sided TFM topology [41].

Unlike in conventional RFMs, the flux lines in this topology lie in the perpendicular or, in other words, transversal plane to the direction of movement and that of current flow [38]. The major differences between PM based TFM with RFM and AFM were mentioned in Chapter 1. The TFM machine allows an increase of the space for the windings without decreasing the available space for the main magnetic flux. The TFMs can be also made with a very small pole pitch compared with other machines. This feature of the machine results in higher force density than RFMs and AFMs.

2.3 TFM Characteristics

TFMs are known for their high power density, where at least in theory, the power rating of the machine can be increased by increasing the number of poles.

2.3.1 Significance of pole number

Increasing the number of poles for a given machine envelope and current loading will increase the machine VA rating and consequently, higher values of specific torque density can be achieved [38]. This concept can be illustrated using the TFM shown in Fig. 2.1, where the number of poles in the lower machine is doubled compared to the upper one.
The pole pitch has been reduced to half along with the length of stator stack, as well as the length of the magnets. This indicates that the magnetic loading in the machine has not been altered. The amount of the flux that encircles the winding is still the same, for a constant current loading. However, the rate of change of the flux is doubled which results in the doubling the induced EMF according to (2.1). Thus, decreasing the pole pitch in the machine by half, for the same amount of iron, copper, and current, the VA rating in the machine can be doubled. However, the pole pitch has a lower bound as it affects the power factor and mechanical rigidity of the iron cores. The EMF in a TFM machine can be written according to:

\[
EMF = k_f N \varnothing \frac{\omega}{2\tau}
\]  

(2.1)

where \( k_f \) is the waveform factor, \( N \) is the number of turns \( \varnothing \) is the flux encircling the winding, \( \omega \) is the speed of the mover and \( \tau \) is the pole pitch.

2.3.2 Independent current and magnetic loading

The specific output torque of an electrical machine can be improved either by increasing the air gap flux density or by increasing the current loading of the armature. Increasing the air gap flux density requires a compensation in the teeth width to prevent the excessive level of saturation in the iron core. Increasing the current loading requires increasing the winding area to maintain the level of the current density based on the thermal requirements. The strong competition for the same space by the stator teeth and the armature conductors sets restrictions to achieve high power densities in the RFM.

One of the attractive features of the TFM is that the current and magnetic loading can be set almost independently. The machine axial length sets the magnetic loading, whereas
the lateral width of the machine determines the current loading [38]. This allows for a more favorable construction, as the rotor magnetic circuit and the stator armature winding are not competing for the same space.

2.3.3 Power factor

One of the major drawbacks of a TFM is its low power factor where values in the range of 0.35-0.55 are typical [30]. Since the rating of the drive inverter is inversely proportional to the power factor, there will be a substantial increase in the power rating of the inverter. Hence, the drive circuit costs for the TFM are expected to be higher than for conventional machines. Furthermore, according to [30], the scope for the improvement in this area is very limited, and a low working factor is something inherent in the TFM topology. The problem with the power factor in a TFM is strongly dependent on the leakage flux. In [30], values of about 50% magnet leakage flux and about 70% armature flux leakage have been reported. Due to the high leakage (poor power factor), a higher inverter (VA) rating would be required (per phase) for a given power (kW) output, as compared to the case for conventional machines.

2.3.4 End winding

The absence of physical end winding is an attractive feature in TFMs since no extra space is required for those. However, in a conventional TFM as shown in Fig. 2.1, only 50% of the winding is active at any particular time. As the winding between two stator core is surrounded by air, and not by iron, the cooling of the machine will be aggravated. Furthermore, these parts of the winding contribute to increased weight, leakage, and copper losses in the machine and should, therefore, be treated as end windings [38]. The leakage
flux that encircles the portion of the inactive winding will interact with the flux from the
magnets just underneath. The flux orientation in the inactive winding would be the same
as it is in the stator, however, the flux orientation of the magnets is in the opposite direction
compared with the torque producing magnets under the stator teeth. Therefore their
interaction may cause a braking torque.

2.3.5 Cogging torque

Cogging torque can be a major issue for TFMs since the stator tooth span is
approximately equal to a pole span. Hence, cogging reduction by conventional techniques
such as skewing is inefficient, and fractional-pitch winding is unavailable. The amount of
the cogging in a TFM can be reduced by designing a machine with a lower magnet loading
and a higher current loading. However, this will, in turn, reduce the power factor as the
amount of armature flux leakage in a TFM is dependent upon the amount of the current
loading.

2.3.6 Multi-phase TFM

TFMs, in general, have a single-phase structural unit. The separate units are
subsequently attached to each other in the lateral direction in order to produce a multi-
phase machine. The three-phase winding from three individual sing-phase units does not
produce a common rotating field. For a three-phase machine, three independent alternating
fields shifted by 120 electrical degrees is created by the mechanically displacing the
magnets in the rotor. The electromagnetic properties of a single unit are therefore inherited
by the entire machine. Since there is less interaction between phases, a fault in one phase
does not bring the operation of the machine to a standstill. However, there will be less specific torque and more torque ripple in case of a fault.

2.3.7 Manufacturing complexity

A significant disadvantage of the TFM is the complicated mechanical structure of the magnetic circuit. The TFM rotor and stator consists of a large number of separate small-size components, which results in a weak construction and more complex assembly.

2.4 Different TFM Topologies

It is possible to create different topologies of TFM both in terms of stator and rotor configurations. To maintain the basic operating characteristics of TFM, the magnetic flux and the torque producing current must be perpendicular to each other and the magnetic flux has to cross the airgap vertically in order to produce a tangential force component. In [39], different type of windings: Gramme, Drum, Pole, and Ring, as shown in Fig. 2.2, have been presented as possible winding solutions to ensure the transverse flux characteristics of the TFM.

![Figure 2.2: Four variants of armature winding topologies.](image)

An overview of the existing TFM topologies based on the possible rotor and stator configurations has been presented in [40]. Based on its magnet location, the PM-TFM can be classified as Surface-Mounted (SM) and Flux-Concentrating (FC). Depending on the location of the stator, PM-TFM can be single-sided (SS) or double-sided (DS). PM-TFM can
can have an outer rotor and inner rotor configurations as well. The shape of the stator cores is also changing based on the need for the design engineer to optimize it for less flux leakage. U-Core, C-Core, E-Core, Claw pole and Z-Core are some of the stator core types reported in the literature. U-shaped stator core is the most popular type due to its simple construction. The advantage and disadvantage of the TFM topologies will be discussed in the following section.

2.4.1 Surface mounted PM-TFM

Single-sided surface mounted PM-TFMs (SSSM PM-TFM) as shown in Fig. 2.3 are the simplest TFM in terms of construction. The drawback for this configuration is that it utilizes only every other magnet. As a result, half of the machine magnetic potential remains completely unused.

![SSSM PM-TFM assembly with single winding.](image1)

![SSSM PM-TFM with iron bridges and single winding.](image2)

The magnets, which are completely exposed to air and does not see the stator core will attract a leakage flux from the adjacent iron parts. In [39, 41], the authors proposed a remedy to this issue, which required inserting a short iron bar to provide shunt paths for unused magnets as shown in Fig. 2.4. This prevented the non-torque-productive magnet flux to link to the stator winding in an adverse manner. However, the presence of iron bridge increases the active weight of the machine and reduces the slot area used for winding. This means that the machines with bridges need to operate at a higher current.
density in order to achieve the same performance [30]. The bridges may also attract significant leakage flux from the adjacent stator cores.

Another way to increase the magnet utilization is to have a SSSM PM-TFM with Claw Pole [42] stators which guide the flux in such a way that all magnets are utilized at the same time as shown in Fig. 2.5. The drawback for this configuration is that only half of the available pole area of the stator tooth is being utilized. This is done intentionally for reducing the leakage of the armature flux between the twisted stator teeth. Shortening the tooth span also tend to saturate the stator tip, resulting in a lower magnetic flux linkage.

SSSM PM-TFM with isolated poles in the rotor is shown in Fig. 2.6. This will prevent a low reluctance path between the adjacent poles, which will reduce the pole-pole leakage.
However, the drawback of this configuration is that the mover will be more complicated to build and mechanically more unstable.

Double Sided (DS) SM PM-TFM can be developed using the SSSM PM-TFM, by having a stator with the teeth facing the inactive magnets under the rotor of the SSSM PM-TFM as shown in Fig. 2.7. This allows for complete utilization of the magnets when compared to the SSSM PM-TFM. The DSSM PM-TFM has higher VA rating for the same volume and relatively small pole pitches are possible [41] compared to SSSM PM-TFM. However, the drawback is the increased complexity associated with constructing the active rotor parts as they have to be supported in a cantilevered arrangement [42].

2.4.2 Flux concentrated (FC) PM-TFM

Having the rotor magnets arranged in a flux-concentrating geometry, ensure complete utilization of the peripherally-magnetized magnets. The magnets are buried in the rotor, and the magnetic flux can be focused in the rotor to increase the air gap flux density. FC PM-TFM is reported to have higher power factor and higher torque density when compared to SM PM-TFM [43]. The drawback for this configuration are the active components of the rotor need to be supported in the cantilevered arrangement, which makes the construction complex. Double-sided and single-sided arrangements for the FC PM-TFM are shown in Fig. 2.8. Examples of different FC PM-TFM can be found in [44-47].
2.4.3 Consequent-pole TFM

There is a concept TFM that had been patented in 1992 but not scientifically-probed [48], which utilizes all magnets simultaneously resulting in unity magnet utilization factor. It is a single-sided surface-mounted PM-TFM with intermediate poles and single winding as shown in Fig. 2.9.

Figure 2.8: Flux-concentrated PM-TFM with single-sided and double-sided stator topology.

Figure 2.9: Single-sided surface mounted Z-TFM.
In this topology, all magnets are active and contribute to the flux linkage. This is achieved by introducing a certain complexity in the stator where the flux from intermediate poles, of opposite polarity, is guided in a way that it encircles the winding in the same direction. Theoretically, simultaneous utilization of all magnets will almost double the induced voltage and the power rating of the machine [49]. Furthermore increased number of the iron parts in the stator will reduce the end winding effect. Other advantages with the topology are reduced magnet flux leakage, the improved thermal behavior of the machine and reduced flux fringing. However, all benefits from the introduction of the extra iron stack in the stator lose significance because of the poor force production. In [49], it was reported that the 3D FEA simulated value of the force produced in a 10 kW machine was only about 10 % of the force calculated analytically. A detailed analysis revealed a very high 3D leakage effect of the armature reaction flux between the adjacent stator cores. Almost no armature reaction flux entered the air gap which caused a huge loss in the torque production. Insertion of the extra stator stack introduces an overlapping iron area carrying oppositely oriented fluxes. The reluctance between the adjacent stator cores was much smaller, due to the short space between the poles, compared to the reluctance in the air gap. Hence, this resulted in high armature flux leakage (around 90%). In order to decrease high pole to pole leakage flux a larger space between the poles is necessary. This on the other hand, for the same machine dimensions, will lower the induced EMF due to higher pole pitch. It was concluded in [49] that the Z-TFM will have a larger diameter compared to the conventional TFM for the same machine rating. The Z-TFM will also have a higher cogging torque, as it utilizes twice the number of magnets compared to a conventional TFM.
2.5 Recent TFM Designs

Some recently published work with different TFM topologies (with prototypes developed) will be discussed in this section. In [50-52], a single-sided claw pole TFM with the outer rotor is designed for electric bike application, shoulder joint motor in a service robot arm and aerospace application respectively. The TFM cores were made with Soft Magnet Composites (SMC) in [50-52]. This is a big disadvantage as the TFMs will have much lower torque producing capability when compared to TFMs built with laminated steel. In [50] and [52], the outer rotor had flux-concentrated rare-earth based NdFeB. In [51], the outer rotor had a surface mounted rare-earth based NdFeB. In [50], the TFM at a rated speed of 250 rpm achieved 6.9 Nm/kg torque density in the simulation. The prototype is under development. In [51], the prototype TFM was experimentally tested at a rated speed of 1100 rpm and had a power factor of 0.65 and an efficiency of 80%. In [52], from the prototype experiments at a rated speed of 450 rpm, the 3-phase machine had a torque density of 24.5 Nm/kg.

In [53], two single-sided U-Core outer rotor TFMs, with and without magnetic shunts has been developed. Both had an outer rotor with surface mounted rare-earth based NdFeB. The TFM with magnetic shunts showed better performance, due to the reasons explained in Section 2.4.1. Both the TFMs were experimentally tested at a rated speed of 600 rpm and achieved a torque density of 1.42 kW/kg without a magnetic shunt and 2.30 kW/kg with a magnetic shunt.

In [54], single-sided consequent pole TFM with E-Core stators was developed. The inner salient pole rotor has no magnets, and the rotor was divided into 3 segments with the inner and outer segments having a half pole pitch shift. The magnet was mounted on the
center pole of the E-Core stator. Double-winding was used for the armature excitation. No prototype has been built, but through 3D-FEA, the authors claim to have significantly reduced cogging torque through the innovative rotor arrangement. The main drawback with this topology is the use of PM in stator cores, as it increases the chance of demagnetization due to the heat generated from the double winding structure.

All the aforementioned TFMs were designed with rare-earth based PM and does not take into consideration the issue of supply and price volatility associated the NdFeB.

2.6 Summary

A thorough literature review of the prior art and the recently developed TFMs has been presented this chapter. The advantages and drawbacks associated with each design are discussed, laying the foundation for selecting a novel topology for the proposed TFM, presented in Chapter 3.
CHAPTER III

DESIGN OF A TFM FOR DIRECT-DRIVE APPLICATIONS

3.1 Introduction

The design considerations for the proposed TFM topology are based on extensive literature review of prior TFMs (presented in Chapter 2), and on the National Renewable Energy Laboratory’s (NREL) SWIFT wind turbine specification requirements. This chapter will give a comprehensive overview about the TFM mechanical, power converter and electromagnetic design considerations, from the Swift Wind Turbine specifications and prototyping point of view. The case for choosing the proposed TFM topology for further research and prototyping to meet the performance specifications are explained in this chapter.

3.2 Direct-Drive Wind Turbine Application

Wind generator systems are slowly moving away from geared single-speed drives to a direct-drive configuration that increases the system efficiency by avoiding the mechanical losses in the gears [55]. It also increases system reliability by reducing failures in the gearbox and lowering maintenance downtime. Direct-drive generators with permanent magnet excitation offer the added advantage of eliminating excitation losses and reducing the active weight of the machine. Based on the NREL Swift test turbine configuration [56]
given in Table 3.1, a TFM would be designed to generate a 1 kW output power at a rated speed of 400 rpm.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>No. of blades</td>
<td>5</td>
<td>Rated wind speed</td>
<td>11 m/s</td>
</tr>
<tr>
<td>Blade Radius</td>
<td>2.1 m</td>
<td>Rated rotor speed</td>
<td>400 rpm</td>
</tr>
<tr>
<td>Cut-in speed</td>
<td>3.4 m/s</td>
<td>Rated power</td>
<td>1 kW</td>
</tr>
</tbody>
</table>

3.3 Baseline TFM Design Considerations

The motivation of the proposed research is to develop a compact, high torque density, energy-efficient, rare-earth-free permanent magnet (PM) electric machine for direct-drive Swift Wind Turbine specification using the concept of transverse flux (TF) paths. The design will be optimized for minimum weight and maximum torque/power density and high efficiency over a wide operating area while keeping practical limitations of manufacturability in context.

3.3.1 Machine envelope

The housing and packaging of the motor dictates the TFM envelope for the baseline geometry. Based on an existing dyno test bench at the Alternate Renewable Energy Laboratory at UA, as shown in Fig. 3.1, to perfectly position and mount the prototype TFM for experimental validation testing, the baseline machine outer diameter (MOD) and machine axial length (MXL) have been set at 225mm and 105mm respectively.
3.3.2 Machine axial air gap length ($l_g$) and rotor split ratio ($\lambda$)

Since only a single module of the TFM would be prototyped, the machine will have no starting torque but there will be a very high cogging torque with short air gaps. Along with prototyping limitations to create and maintain a uniform short axial air gap, a 1mm air gap length ($l_g$) was selected and deemed to be practically achievable while prototyping.

The rotor split ratio ($\lambda$) is defined as the ratio of rotor inner diameter ($D_i$) to rotor outer diameter ($D_o$) as shown in Fig. 3.2. For AFMs and TFMs, the rotor split ratio ($\lambda$) is a major design parameter that significantly affects the machine performance characteristics [57]. The optimal value for $\lambda$ depends on the optimization goal and varies for different electrical and magnetic loading conditions. Researchers [58, 59] have extensively worked on deriving an optimum value of $\lambda$ and found that for machines with an axial air gap, $\lambda$ should be selected between 0.57 to 0.63. For the initial baseline TFM design, $\lambda$ is selected as 0.6, from which we get the shaft diameter of 137mm since the MOD was already set at 225mm.
3.3.3 Current density and power converter limits

Higher current density implies increased electrical loading and increased average torque output until the machine starts to saturate, however, it does come with the penalty of increased copper loss and excess heat generation. The value of current density is limited only by the ability to cool the machine and the maximum allowable temperature in the machine. Exposure to high temperature in the stator cores would lead to winding insulation break down and degrades the “Hot Strength” of the epoxy holding the stator cores to the housing. Typical values for current densities used in motors cooled by different methods are given in Table 3.2 [60]. For the baseline TFM design, the current density is set at 5 A/mm², to eliminate the need for an additional cooling method. Based on this current density limit, the number of turns for the coil conductor and winding slot cross-sectional area will be optimized in Chapter 4.

Table 3.2: Typical values of current-density with cooling method.

<table>
<thead>
<tr>
<th>Cooling Method</th>
<th>Current Density (A/mm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Totally Enclosed, Not Ventilated (TENV)</td>
<td>4.7 -5.4</td>
</tr>
<tr>
<td>Air over; fan-cooled</td>
<td>7.8-10.9</td>
</tr>
<tr>
<td>External blower; through-cooled</td>
<td>14.0-15.5</td>
</tr>
<tr>
<td>Liquid-cooled</td>
<td>23.3-31.0</td>
</tr>
</tbody>
</table>
The power converter that will be used for driving the TFM, will be operating with a DC Bus voltage of 48 V and a peak current carrying capability of 200 A. As a result, during the electromagnetic design, it must be ensured that the voltages induced in the coil conductor of the TFM should not exceed 48 V at rated load. Based on the current density, number of turns, slot area, and peak current handling capability of the power converter, the wire gauge suitable for the TFM winding will be selected.

3.3.4 Choice for permanent magnet material

Several PM materials are available in the market including alnico, ferrite (ceramic), rare-earth samarium-cobalt (Sm-Co), and rare-earth neodymium-iron-boron (NdFeB). Ferrites were introduced in 1953 and have now become a matured cost-effective PM material for many types of motors. Sm-Co, introduced in the mid-1970’s, has a much higher energy product than ferrites and is thermally more stable. However, they are very expensive and only used in applications that demand high performance over an extended temperature range. In the mid-1980’s, a new generation of NdFeB was developed which has a higher energy density than Sm-Co but it is less expensive. However, it was not as thermally stable as Sm-Co and suffers from corrosion problems.

The most suitable magnets for the brushless motors (RFM, AFM, and TFM) used in direct-drive applications are the ferrite magnets and the high-energy rare-earth NdFeB magnets. Both magnets have straight characteristics throughout the second quadrant and are classified as hard magnets because of their high resistance to demagnetization. Other magnets, particularly Alnico magnets, have a high remnant flux but very low coercive magneto motive force (MMF) and low resistance to demagnetization. Table 3.3 and Fig. 3.3 give a comparative analysis of the aforementioned magnetic materials.
At room temperature, NdFeB has the highest energy product of all commercially available magnets. The high remanence and coercivity allow for a significant reduction of the motor frame size for the same power output compared to motors using ferrite magnets. However, ferrite magnets are considerably cheaper and are not subject to supply and price volatility. Since the objective of the research was to design a low cost rare-earth-free PM based TFM, ferrite magnets became the obvious choice for the baseline design.

Table 3.3: Magnet properties (20ºC) [60].

<table>
<thead>
<tr>
<th>Property</th>
<th>Units</th>
<th>Alnico</th>
<th>Anisotropic Ferrite</th>
<th>Sintered Sm-Co</th>
<th>Sintered NdFeB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Remanence $B_r$</td>
<td>T</td>
<td>0.63 – 1.35</td>
<td>0.35 – 0.43</td>
<td>0.7 – 1.05</td>
<td>1.0 – 1.3</td>
</tr>
<tr>
<td>Intrinsic Coercivity $H_{ci}$</td>
<td>kA/m</td>
<td>40 - 130</td>
<td>180 - 400</td>
<td>800 - 1500</td>
<td>800 – 1900</td>
</tr>
<tr>
<td>Recoil Permeability $\mu_{rec}$</td>
<td></td>
<td>1.9 - 7</td>
<td>1.05 – 1.15</td>
<td>1.02 – 1.07</td>
<td>1.04 – 1.1</td>
</tr>
<tr>
<td>$(BH)_{max}$</td>
<td>kJ/m$^3$</td>
<td>20 - 100</td>
<td>24 - 36</td>
<td>140 - 220</td>
<td>180 – 320</td>
</tr>
<tr>
<td>Magnetizing force</td>
<td>kA/m</td>
<td>200 - 600</td>
<td>600 - 1700</td>
<td>1600 - 4000</td>
<td>2000 – 3000</td>
</tr>
<tr>
<td>Resistivity</td>
<td>$\mu\Omega$cm</td>
<td>47</td>
<td>&gt;10$^4$</td>
<td>86</td>
<td>150</td>
</tr>
<tr>
<td>Thermal Expansion $B_r$ temperature coefficient</td>
<td>$^\circ$C</td>
<td>(-0.01) – (-0.02)</td>
<td>-0.2</td>
<td>(-0.0045) – (-0.05)</td>
<td>(-0.08) – (-0.15)</td>
</tr>
<tr>
<td>$H_{ci}$ temperature coefficient</td>
<td>$^\circ$C</td>
<td>-0.02</td>
<td>0.2 – 0.4</td>
<td>(-0.2) – (-0.25)</td>
<td>(-0.5) – (-0.9)</td>
</tr>
<tr>
<td>Max. working temperature</td>
<td>$^\circ$C</td>
<td>500 - 550</td>
<td>250</td>
<td>250- 350</td>
<td>80 - 200</td>
</tr>
<tr>
<td>Curie temperature</td>
<td>$^\circ$C</td>
<td>850</td>
<td>450</td>
<td>700 - 800</td>
<td>310 - 350</td>
</tr>
<tr>
<td>Density</td>
<td>kg/m$^3$</td>
<td>7300</td>
<td>4900</td>
<td>8200</td>
<td>7400</td>
</tr>
</tbody>
</table>
Figure 3.3: (a) Typical B/H curves for PM materials [6], (b) Typical characteristic for permanent magnets.

3.3.5 Stator and rotor design considerations

The decision to select ceramic ferrite magnets over rare-earth based NdFeB for designing the TFM means there is already a significant electromagnetic design penalty in terms of magnetic energy from PMs. It is quintessential that the choice for the stator and rotor topology be such that maximum torque density can be extracted from the TFM through efficient magnetic and electrical loading optimization.

In Chapter 2, several rotor and stator topologies have been explored and their advantages and disadvantages were identified through an extensive literature review. After careful consideration, based on electromagnetic performance and prototyping complexity, the down-selected TFM topology consisted of a

- Stator with intermediate poles (Zwegbergks Stator).
- Rotor with flux-concentrating (FC) buried magnets.
- Double-sided stator with a single rotor configuration.

The Zwegbergks stator ensures complete utilization of all rotor magnets simultaneously, which means that all magnets contribute to the winding flux linkage. The advantages and disadvantages of this stator topology were discussed in Chapter 2. Their
drawbacks would be addressed in this section. A rotor with a flux-concentrating set up has been selected to address the issue of low power factor, and most importantly to use the flux-focusing features to amplify the air gap flux density. Since low energy ferrite magnets are to be used, there is no alternative to using an FC rotor topology to generate high torque output within a compact motor structure. A double-sided topology was selected to allow for active utilization of the magnets from both sides of the rotor, which increases the VA rating of the machine within the same MOD. The concept of using intermediate stator poles also increases the number of iron parts in the stator which will reduce the length of the winding facing the air. The heat generated in the winding is easily cooled away through conduction in the iron parts, along with air convection. This would also improve the thermal behavior of the proposed TFM.

3.3.5.1 Stator design parameters

The proposed stator topology for a pole pair is shown in Fig. 3.4 and the parameterized geometry of the stator is shown in Fig. 3.5.

![Figure 3.4: Magnetic flux flow through the proposed stator topology.](image1)

![Figure 3.5: Parameterized stator core geometry.](image2)

The individual stator cores are identical in shape but are alternately turned 180° with respect to the PM axis. The magnetic flux in the stator back iron for all the stator cores is
in the same direction as shown in Fig. 3.4, so a toroidal winding setup will be used in both the upper and lower stator. The main drawback of using the Zwegbergks stator topology comes in the form of pole-pole leakage, which makes the selection of pole number (p) and stator pole width (Wₕ) critical to the electromagnetic design of the TFM. The design parameters for the stator core that need to be optimized are listed in Table 3.4.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Parameter</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>LP</td>
<td>Pole length</td>
<td>SO</td>
<td>Slot opening</td>
</tr>
<tr>
<td>Wₕ</td>
<td>Stator Pole width</td>
<td>SH</td>
<td>Slot height</td>
</tr>
<tr>
<td>SBI</td>
<td>Back iron thickness</td>
<td>SW</td>
<td>Slot width</td>
</tr>
<tr>
<td>HS1</td>
<td>Center pole height</td>
<td>HS2</td>
<td>Center pole trunk</td>
</tr>
</tbody>
</table>

3.3.5.2 Rotor design parameters

The initial design for the rotor with only two flux-focusing magnets per pole is shown in Fig. 3.6. Two flux-focusing magnets, one with positive radial and the other with negative radial magnetization, are alternately focusing the magnetic flux towards the center pole and away from it. This was not enough to produce the required rated power output of 1 kW at a rated speed of 400rpm.

Figure 3.6: Initial rotor design with only flux-focusing magnets.
To meet the performance specification requirement, additional flux-focusing magnets were added to the rotor as shown in Fig. 3.7. The new rotor setup helped to reduce the pole-pole flux leakage in the rotor. The additional magnets with positive and negative orthogonal magnetization were used to aid in focusing the magnetic flux flow to and away from the alternating rotor poles. Use of these additional magnets would add to the cost and complexity associated with developing the overall machine, but they are necessary to meet the required performance specification for the NREL Swift wind turbine application.

![Figure 3.7: Improved rotor design with additional flux-focusing magnets.](image)

The design parameters for the rotor core that needs to be optimized are listed in Table 3.5.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Parameter</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>LP</td>
<td>Pole length</td>
<td>LM</td>
<td>Magnet Length</td>
</tr>
<tr>
<td>W&lt;sub&gt;R&lt;/sub&gt;</td>
<td>Rotor Pole width</td>
<td>HRY</td>
<td>Rotor/Magnet Thickness</td>
</tr>
<tr>
<td>τ</td>
<td>Pole pitch</td>
<td>W&lt;sub&gt;R&lt;/sub&gt;/τ</td>
<td>Pole embrace</td>
</tr>
</tbody>
</table>
3.3.5.3 Selecting optimum pole width and pole number

The flux path in the stator cores allows for an increase in the VA rating of the machine by increasing the pole number \( p \), for a given machine geometry (MOD and MXL), current, and magnetic loading. This can be illustrated by the Equation for electrical loading \( (A_m) \) in (3.1),

\[
A_m = p \frac{\sqrt{2} I_{\text{rms}} N}{\pi D_g} \quad (3.1)
\]

\[
f = \frac{N_s p}{120} \quad (3.2)
\]

\[
\tau = \frac{2\pi (\text{Shaft radius})}{(p/2)} \quad (3.3)
\]

where \( D_g \) is the average air gap diameter, \( I_{\text{rms}} \) is the phase RMS current and \( N \) is the number of turns. In single phase TFMs the number of stator and rotor poles is kept the same to ensure a balanced back-EMF in both the top and bottom stators. The pole number controls the peripheral length of the poles and sets the pole pitch (\( \tau \)) length as shown in (3.3). With increasing \( p \), the amount of flux linking the stator conductor remains the same (neglecting leakage flux), but the rate of the change of flux linkage increases, thereby increasing the back-EMF of the machine for the same mechanical speed. However, increasing the pole number increases the pole-pole flux leakage, which degrades the power factor, thus setting a limit on the number of poles. There is also a limit imposed on increasing the pole number based on the electrical frequency \( f \) according to (3.2), at which the machine can operate at because of the motor drive's power electronics and iron loss.

Although it is generally agreed that the choice of pole number should be high to achieve high torque densities, selecting the pole number is a choice firmly linked with the pole pitch (\( \tau \)) and pole-pole overlap area.
3.3.5.3.1 Optimizing pole width for a constant pole pitch

A fixed 30-pole machine has been selected to optimize the pole width for both the rotor and the stator. For a 30-pole machine, each pole pair occupies a 24° mechanical span as shown in Fig. 3.8, which sets the limit for the maximum circumferential length at 28.7 mm for the rotor inner periphery (RIP) and 47.1 mm and a pole pitch (τ) of 14.3 mm.

Figure 3.8: Pole pair cross-section of the proposed TFM.

In order to optimize the rotor and stator pole width, two new geometric ratios a) Pole embrace and b) Pole-pole overlap are introduced, defined by Equations (3.4) and (3.5).

\[ P_{embrace} = \frac{W_R}{\tau} \]  \hspace{1cm} (3.4)

\[ \lambda_{r/s} = \frac{W_R}{W_S} \]  \hspace{1cm} (3.5)

\[ P_{embrace} \] is the fraction of the pole pitch occupied by the rotor pole width, and \( \lambda_{r/s} \) is the rotor-pole to stator-pole overlap ratio. Both design ratios investigate the effect of slotting on the air gap flux distribution in the machine, which significantly impacts the back-EMF wave shape, power factor and torque, density.

The back-EMF waveform is an important design aspect for PM machines. The back-EMF waveform factor \( K_p \) in radial machines is defined by (3.6)

\[ K_p = f(L_m, K_d, K_{sl}, K_a K_w) \]  \hspace{1cm} (3.6)

\[ K_p = f(W_R, W_S, N) \]  \hspace{1cm} (3.7)

where \( L_m \) is the magnet width; and \( K_d, K_{sl}, K_a \) and \( K_w \) are the distribution, slot, skew, and winding factors, respectively. In radial machines, the winding configuration and stator slot
selection play a major role in the back-EMF waveform factor. In the proposed TFM, the rotor and stator pole numbers are the same, and ring windings are used; thus, the back-EMF waveform factor for the proposed machine would take the form of (3.7).

The first stage of the analysis is to decide on the pole-pole overlap ratio, on whether the rotor and stator pole width should be equal to each other or should there be an offset. For a fixed pole pitch the effect of changing $\lambda_{r/s}$ on the machine no-load and load performance was analyzed using 3D-FEA, and the results are shown in Fig. 3.9.

![Graphs showing the effect of $\lambda_{r/s}$ on machine performance.](image)

Figure 3.9: Effect of $\lambda_{r/s}$ on the machine performance.

From Figs. 3.10 (a)-(d), the stator pole width $W_s$ was kept constant 8.25mm, but the rotor pole width $W_R$ was varied. It can be seen that by increasing $W_R$,

- the RMS back-EMF voltage increased as more magnet flux was linked to the stator winding,
- the peak to peak cogging torque reduced due to a decrease in slotting between the rotor and stator.
However, the optimum torque density (which is the main objective for the design optimization) was reached when $W_R$ was equal to $W_S$ at 8.25 mm, and the power factor penalty was not significant compared to other design points. As a result, $\lambda_{r/s}$ was selected as 1.

The effect of $P_{embbrace}$ was investigated for a fixed pole pitch with $\lambda_{r/s}$ set to 1. The effect of $P_{embbrace}$ on torque density and the power factor is shown in, Fig. 3.10.

![Figure 3.10: Effect of $P_{embbrace}$ on the machine performance.](image)

It can be seen that by increasing $P_{embbrace}$,

- the torque density increases as more active magnet material is added to the machine,
- the power factor drops sharply as the pole-pole leakage between the consequent poles increases.

Considering the overall effects on torque density and power factor, along with prototyping considerations, the optimum $P_{embbrace}$ was selected as 0.57. Using the optimum value for $P_{embbrace}$, the optimum rotor pole width is obtained from (3.8).

$$W_R = P_{embbrace}(0.5)\left(\frac{1}{p/2}\right)(2\pi)(Shaft\ Radius) = 8.25\ mm \quad (3.8)$$
3.3.5.3.2 Selecting the number of poles

Once $P_{embrace}$, $\lambda_{r/s}$, and the pole width were optimized, the effect of number of poles on the machine performance was investigated. Changing the pole number will change the pole pitch ($\tau$), but the current and magnetic loading was kept constant. Six different pole numbers were selected for analysis as shown in Fig. 3.11 and the choice for the pole number was dictated by the operating frequency, iron loss and structural rigidity of the cores.

![Different pole configurations](image)

Figure 3.11: Different pole configuration for the Z-TFM.

The effect of increasing the pole number on the torque density and power factor is shown in Fig. 3.12. Increasing the pole number increases the torque density for reasons explained in Section 3.3.5.3. However, the power factor starts to decrease, since in consequent pole TFMs, with a larger pole number, the pole-pole flux leakage increases significantly.
Figure 3. 12: Effect of pole number on (a) torque density and (b) power factor.

The pole number is also a function of the speed of the machine and the flux frequency, which is the product of the rotor rotational frequency and pole-pair number [61]. The limitation to increasing the pole number is that the flux in the machine would alternate at a high frequency resulting in excessive iron losses which affect the machine efficiency. The thickness of the lamination also makes a significant difference in eddy-current loss [60]. For building the prototype of this TFM, 0.35 mm thick steel laminations would be used. Considering the flux level and degree of saturation in the TFM from the 3D-FEA simulation, an operating frequency of less than 150 Hz provided an acceptable core-loss while having acceptable torque density and power factor performance. Thus, a 30 pole machine operating at an excitation frequency of 100 Hz at the rated speed of 400 rpm that meets the required performance criteria for the TFM has been selected. The 36 pole machine despite having better torque density and similar power factor to the 30 pole machine, was not selected due to increased complexities for developing the prototype with the large increase in lamination parts.
3.4 Conclusion

This chapter introduced the proposed rare-earth magnet free PM-TFM, designed for direct-drive wind turbine applications. Based on the application specification, dyno test setup, power converter rating, current density, prototyping feasibility, and choice of magnets, an initial geometry for the rotor and stator was developed. Table 3.6, brings together the critical dimensions, design ratios and design limitations that were determined during the initial design phase in this chapter.

Table 3.6: Initial baseline TFM design parameters and power converter limits.

<table>
<thead>
<tr>
<th>Critical electromagnetic design parameters</th>
<th>Machine Outer Diameter (MOD)</th>
<th>225mm</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Machine Axial Length (MXL)</td>
<td>105mm</td>
</tr>
<tr>
<td></td>
<td>Shaft Diameter</td>
<td>137mm</td>
</tr>
<tr>
<td></td>
<td>Rotor Split Ratio (λ)</td>
<td>0.6</td>
</tr>
<tr>
<td></td>
<td>Number of poles</td>
<td>30</td>
</tr>
<tr>
<td></td>
<td>Air Gap Length</td>
<td>1mm</td>
</tr>
<tr>
<td></td>
<td>Pole Embrace Ratio, $P_{embrace}$</td>
<td>0.575</td>
</tr>
<tr>
<td></td>
<td>Pole Overlap Ratio, $\lambda_{r/s}$</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Pole Width, ($W_R = W_s$)</td>
<td>8.25mm</td>
</tr>
<tr>
<td></td>
<td>Pole Pitch τ</td>
<td>14.35mm</td>
</tr>
<tr>
<td></td>
<td>PM remanant flux density</td>
<td>0.4T</td>
</tr>
<tr>
<td></td>
<td>Maximum current density</td>
<td>$5A_{rms/mm^2}$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Power converter limits</th>
<th>DC Bus voltage</th>
<th>≤48V</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Input phase current</td>
<td>$≤141A_{rms}$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Output requirements at rated condition</th>
<th>Rated Average Torque</th>
<th>24 Nm</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Rated Output Power</td>
<td>1 kW</td>
</tr>
<tr>
<td></td>
<td>Rated Speed</td>
<td>400rpm</td>
</tr>
</tbody>
</table>

In order to further optimize the remaining stator and rotor design parameters from Table 3.4 and 3.5, it would be extremely difficult and time-consuming to perform a multi-dimensional and multi-objective based 3D-FEA design optimization. To address this issue, a two-stage solution will be developed in the upcoming chapters. The first stage will
include the development of a comprehensive MEC based analytical model for the proposed TFM capable of predicting the TFM electromagnetic performance under varying electrical and magnetic loading conditions. The second stage would be to use an optimization algorithm with the MEC model, to perform a multi-dimensional and multi-objective design optimization for the proposed TFM. The developed analytical design optimization tool will be computationally less intensive and would allow faster design optimization for the TFM.
CHAPTER IV
MAGNETIC EQUIVALENT CIRCUIT MODEL OF TFM

4.1 Introduction

Accurate prediction of the parameters and characteristics of electromagnetic devices with complex geometry and high saturation using advanced numerical methods like finite element solvers are time-consuming processes that require significant computational resources. Altering design parameters (machine dimensions, current density) in 3D-FEA requires the model to be reconstructed and re-meshed. Taking these factors into account, 3D-FEA is not ideal to be used during the initial design stage of the electric machine. A faster nonlinear analytical model capable of producing similar results to the finite element (FE) solvers is essential to facilitate faster analysis of the preliminary TFM design.

The magnetic equivalent circuit (MEC) model is a common analytical tool which analyzes the performance of electric machines by considering the machine’s material characteristics [62-70]. Several analytical models have been developed for different TFM configurations. In [65] magnetic charge and magnetic imaging techniques were used to predict the flux and no-load EMF for a single sided surface mounted PM based TFM with 15% error. The MEC modeling concept has been applied to different TFM configurations in [66-70] for predicting air gap flux density with less than 15% error. In [71] and [72], the MEC method has been applied to a claw pole alternator and a linear PM synchronous
machine. In both cases, magnetic saturation was addressed using the Hopkinson law while calculating reluctances. However, both the MEC models involved multiple flux loops with complex reluctance networks, which makes the model computationally intensive. The 3D magnetic flux path through the TFM requires 3D-FEA to design and optimize the proposed TFM. 3D-FEA is computationally intensive and makes the initial sizing and optimization, extremely time-consuming. In this chapter, an improved MEC model has been developed as an alternative to 3D-FEA, for faster electromagnetic analysis of the proposed double-sided flux-concentrated PM based TFM.

4.2 Improved MEC Model

Although MEC models are based on the fundamentals of solving a magnetic circuit to determine the magnetic flux flowing through the reluctance path, setting up of the reluctance network is distinctive for different machine topologies. Conventional RFMs, with radial flux flow from the rotor to the stator, have only one layer of 2D-reluctance network in the MEC model, as there is no axial magnetic flux flow variation along the stack of the machine. Machine topologies designed with axial (AFM) and transverse flux paths (TFMs), require a much more sophisticated 3D-reluctance network to completely map the flux flow through the machine. In this research, a comprehensive MEC model is developed for a double-sided TFM with a Z-Core Stator [73]. Special attention was given to leakage flux path modeling while taking into account the effects of magnetic saturation in the cores. Volumetric tubes are used to model the paths of magnetic flux through the machine. Discretization of the flux into volumetric flux tubes and defining them as reluctance elements that are a function of the geometry and permeability creates a magnetic circuit that can be solved by mesh analysis using the electrical circuit matrix method.
To develop the reluctance network in the MEC model, separate reluctances are used to represent the rotor poles, rotor back iron, stator poles, stator back iron and air gap. The current carrying coil conductor and the permanent magnets are modeled as MMF sources. The polarity of the PM alternates with the changing rotor position due to the different orientations of the magnets embedded in the rotor. Hence calculation of the flux at every rotor position must be handled carefully. The electromagnetic characteristics of the TFM are the same and repetitive for every pole pair. Therefore, the MEC of one pole pair is considered for the analytical model of Z-Core TFM, as shown in Fig. 4.1. As in FEA, magnetic saturation and leakage are obtained directly from knowledge of the material characteristics and machine geometry based reluctance network in the MEC approach. In contrast to the FEA technique, the nonlinear Equations that are solved at each step in the MEC model are less computationally intensive, thus making it ideal to be used in design optimization, control, and modeling.

The basic Equation that governs each element of the MEC model, is

\[ F = \phi R \] (4.1)

\[ R = \frac{L}{\mu_r \mu_o A_c} \] (4.2)

where \( \phi, R, \) and \( F \) are the flux, reluctance and mmf, respectively, and \( \mu_r, \mu_o, A_c \), and \( L \) are the relative permeability, the permeability of free space, the cross-sectional area, and the length of each element, respectively.
4.2.1 Modeling MMF sources

Each stator winding is modeled as an MMF source $F_c$ where $I$ is the phase current and $N$ is the number of turns. Each PM is modeled as an MMF source $F_M$ in series with an internal reluctance $R_M$, where $\mu_{rm}$ and $B_r$ are the recoil permeability and remanence, respectively. $A_M$, $L_M$, $H_M$, and $L_A$ are the cross-sectional area and thickness of the magnets in the direction of magnet field orientation, length of the magnets in the radial direction, and the axial length of the core, respectively.

$$F_c = NI \quad (4.3)$$
$$R_M = \frac{L_M}{\mu_{rm}H_oA_M} \quad (4.4)$$
$$F_M = \frac{B_rL_M}{\mu_{rm}H_o} \quad (4.5)$$
$$A_M = H_ML_A \quad (4.6)$$

Four-flux focusing magnets are oriented to either focus the flux to or away from the middle rotor pole. However in the MEC model, the $F_M$ for each magnet is quantified and superimposed into a single magnet model which helps in reducing the flux loops and establish a 2D MEC network.
4.2.2 Flux linkage and back-EMF

The total flux linking the stator winding $\lambda(\theta)$ is obtained by (4.8) where $\theta$ is the rotor position, $R_T$ is the total machine reluctance including the rotor, stator, air gap and leakages.

$$\varphi(\theta) = \frac{F_c + F_M}{R_T(\theta)} \quad (4.7) \quad \lambda(\theta) = N\varphi(\theta) \quad (4.8) \quad e(\theta) = N\frac{\partial \varphi}{\partial \theta} \quad (4.9)$$

The flux entering the core of the TFM is only a part of that present in the air gap. The remaining part of the flux in the air gap is the armature leakage. Due to armature flux leakage, not all the flux produced by the magnets enters the core and actually links the coil. The modeling for the air gap and leakage fluxes is discussed in detail in Section 4.1.4. The no-load back-EMF generated by the TFM can be derived from the flux through the core, which is calculated in (4.7). The back EMF $e(\theta)$ is given by (4.9).

4.2.3 Electromagnetic torque

The torque produced in an electrical machine is given by the rate of change of the magnetic co-energy of its winding with respect to position [74]. Here the MEC model has been used along with the virtual work method to predict the torque in electric machines using the air gap flux information for changing rotor position. The virtual work method states that for a rotating machine, torque is given by (4.10) where, $W'$ is the magnetic co-energy stored within the machine which is obtained by (4.11) where, $\lambda$ is the flux linkage in the coil. It is easier to keep the current constant in the MEC model rather than the flux linkage. Thus calculating torque using the co-energy method is preferred here. The cogging torque is the oscillatory torque of zero average value caused by the tendency of the rotor to line up with the stator in a particular direction where the permeance of the magnetic circuit “seen” by the magnets is maximized. The interaction between the rotor magnetic
flux and variable permeance of the air gap due to the stator slot opening causes the cogging torque. A simplified expression of cogging torque is given by (4.12) where \( \varphi_g \) is the magnet flux crossing the air gap and \( \mathcal{R} \) is the total reluctance through which the flux passes. In addition, cogging torque is independent of the flux direction as the magnet flux \( \varphi_g \) is squared. Again, the reluctance of the PM and the iron core are negligible compared to the air gap reluctance and thus the reluctance here exclusively refers to the air gap reluctance.

\[
T = \frac{\partial W'(\theta, i)}{\partial \theta} \quad (4.10) \quad W'(\theta, i) = \int_{i=0}^{l} \lambda(\theta, i) \, di \quad (4.11) \quad T_{cogg} = -\frac{1}{2} \varphi_g^2 \frac{d\mathcal{R}}{d\theta} \quad (4.12)
\]

### 4.2.4 Air gap reluctance modeling

A crucial component of the MEC is the flux tube in the air gap as shown in Fig. 4.2. The air gap is the primary location of energy storage in the machine. Establishing a closed form expression for air gap reluctance is difficult because the rotor changes position relative to the stator. Therefore, tube dimensions change as a function of rotor position. Tube reluctance \( R_{airgap} \), which is a function of tube length \( L \), cross-section area \( A_c \) and magnetic permeability \( \mu \) inside the tube, is given by

\[
R_{airgap} = \frac{L}{\mu_0 A_c(\theta)} \quad (4.13)
\]

\[
dR = \frac{dz}{\mu A_c(\theta)} \quad (4.14)
\]

\[
R = \int \frac{dz}{\mu A_c(\theta)} \quad (4.15)
\]

Figure 4.2: Flux tube.

In most cases, the tube volume between the planes has a non-uniform length or cross-sectional area. For tubes with non-uniform area, it is convenient to discretize the volume into differential tube reluctances as shown in (4.14) where \( dz \) is the differential tube length,
and $A_c(\theta)$ is the cross-sectional area which must be derived analytically. Integration of the differential reluctance over the entire length of the tube provides the tube reluctance as shown in (4.15). Different reluctances that are included in the air gap model: variable air gap reluctance, complete and partial pole overlap fringing reluctance, stator core fringing reluctance and winding leakage reluctance.

4.2.4.1 Variable air gap reluctance

The variable air gap reluctance across the overlapping poles have two components depending on the position of the rotor as shown in Fig. 4.3.

![Figure 4.3: Stator and rotor pole orientations.](image)

The same stator pole sees alternate flux paths as the magnet pole alternates with changing rotor position. The net flux path is the sum of the two flux paths resulting from different overlapping rotor poles. The air gap reluctance from separate poles are modeled using (4.16)

$$R_{\text{overlap}} = \frac{l_g}{(p\mu_0)A_{\theta_r}} \quad (4.16)$$
4.2.4.2 Overlapping fringing reluctance

When there is partial overlap between the rotor and stator poles as shown in Fig. 4.4, the fringing field is portrayed as 90° wedge curved flux lines coming out from the face (θ-z plane) of one pole and entering the side of the overlapping pole. The cross-section area $A_w$ of the wedge is given by

$$A_w = (r_p)(\theta_{SW} - \theta_r)(r_{lp})$$ (4.17)

where $r_p$ is the radius up to the pole faces, $\theta_{SW}$ is the stator pole pitch and $r_{lp}$ is the pole thickness in the radial direction. The average length of the wedge $L_w$ is given by

$$L_w = \frac{(r_p)(\theta_{SW} - \theta_r)\pi}{2}$$ (4.18)

The partial overlapping reluctance between a rotor and stator pole can be obtained combining (4.17) and (4.18), as given by

$$R_{pooverlap} = \frac{L_w}{(p\mu_o)A_w}$$ (4.19)

When there is a complete overlap between the rotor and stator poles as shown in Fig. 4.5, the flux tube is approximated by a 180° semicircle. The resulting length of the flux path is twice that of the partial overlap case, thus making the complete overlapping fringing to be estimated by

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\[ R_{\text{coverlap}} = 2 \, R_{\text{poverty}} \] (4.20)

4.2.4.3 Stator core fringing reluctance

![Figure 4.6: Leakage paths from the Z-Core stator.](image1)

![Figure 4.7: Fringing path model for the stator.](image2)

The possible flux paths through the air gap and the fringing paths around the air gap of
the stator core are numbered from 1 to 4 as shown in Fig. 4.6 and 4.7. The semi-circular
line number 1 denotes the fringing path between the two stator pole faces. The fringing
between the planes forms a volume of half annulus. The fringing permeance \( P_f \) for the
infinitesimal element of width \( l \) and height \( dr \), is given by

\[
P_f = \int_{r_1}^{r_2} \frac{\mu_0 l}{\pi r} \, dr \] (4.21)

and the reluctance of fringing between two planes is given by

\[
R_{s1} = \frac{\pi}{p \mu_0(l) \log \frac{r_2}{r_1}} \] (4.22)
The line number 2 denotes the fringing between two aligned edges forming a volume of a semicircular cylinder. The average length \( l_{ave} \) of the paths between two parallel lines can be obtained by

\[
l_{ave} = \frac{r_1}{2} \int_{-\pi/2}^{\pi/2} \sqrt{(4 - 3\sin^2 \theta)}
\]

(4.23)

The reluctance for fringing between these two parallel edges is given by

\[
R_{s2} \approx \frac{\left(\frac{r_1}{2}\right) \int_{-\pi/2}^{\pi/2} \sqrt{(4 - 3\sin^2 \theta)^2}}{p\mu_0(\pi)(r_1^2)\left(\frac{\pi}{2}\right)} \approx \frac{4.34}{p\mu_0(l)}
\]

(4.24)

where \( \theta \) is the angle between the two planes. The line number 3 denotes the fringing between the two vertically-aligned parallel edges which forms a quadrant of a spherical shell. The fringing reluctance between two aligned edges can be approximated by

\[
R_{s3} \approx \frac{4}{p\mu_0(r_2 - r_1)}
\]

(4.25)

The line number 4 denotes the fringing between the corners of the pole faces which forms a volume of the spherical quadrant. The fringing reluctance between corners can be approximated by

\[
R_{s4} \approx \frac{\left(\frac{r_1}{2}\right) \int_{-\pi/2}^{\pi/2} \sqrt{(4 - 3\sin^2 \theta)^2}}{p\mu_0(2\pi/3)(r_1^2)\left(\frac{\pi}{2}\right)} \approx \frac{6.51}{p\mu_0(l)}
\]

(4.26)

4.2.4.4 Winding leakage

By applying the Biot-Savart law to the current carrying conductor, the flux \( \varphi_w \), leaking the conductor can be determined as a function of the conductor radius \( R_c \). The reluctance of the winding leakage \( R_w \) is given by
\begin{align*}
  d\varphi_w &= dBA_{coil} \\
  \varphi_w &= \int_0^\pi d\theta = \frac{\mu_0 N I R}{12} \\
  R_w &= \frac{\varphi_w}{N I}
\end{align*}

where \( A_{coil} \) is the cross-section area of the coil.

4.2.4.5 Reduced order 2D MEC model

The 2D-MEC model is used to approximate the 3D flux paths through the air gap between the stator and the rotor which includes the leakage flux paths. Instead of creating a 3D network for the air gap reluctances, the air gap flux paths along the stack are discretized and modeled as parallel reluctances. These are lumped together to form a single equivalent reluctance during the final estimation of the net flux passing through the air gap between a stator and rotor pole. This helps to simplify the MEC network into a 2D plane, thus reducing the number of nodes and flux loops. By increasing the number of discretized airgap reluctance paths, the accuracy of the MEC model can be increased.

The equivalent reluctance model for the stator fringing effect \( R_{StatFringeEqv} \) can be obtained as

\[
R_{StatFringeEqv} = R_{s1} \parallel R_{s2} \parallel R_{s3} \parallel R_{s4}
\]  

Depending on the overlap between the rotor and stator poles, the equivalent variable air gap reluctance can be defined as

\[
R_{AirGapEqv1} = R_{overlap1} \parallel R_{overlap}
\]

for the complete rotor and stator pole overlap and

\[
R_{AirGapEqv2} = R_{poverlap1} \parallel R_{overlap}
\]

for partial rotor and stator pole overlap.
4.3 Nonlinear Solution for the MEC

Magnetic saturation increases losses and degrades machine performance. Due to excessive simplifications, conventional MEC models often lack the ability to accurately predict the machine performance when the machine saturates at higher current. Correction factors based on experience may be used for compensation but are not adaptable to changes in machine geometry. Thus a nonlinear model for the MEC is required to increase the accuracy of estimation result when the machine saturates.

Figure 4.8: Nonlinear analysis using the Gauss-Seidel method.

There are several methods for solving a set of non-linear equations e.g. Newton's and Gauss-Siedel methods. Newton's method requires a Jacobian matrix calculation in which its elements are a partial differential of reluctance with respect to flux. In this case, since reluctances are not an explicit function of flux, obtaining analytical expressions for the Jacobian matrix elements is not possible. Therefore, a Gauss-Siedel method, as shown in Fig. 4.8, with an accelerating factor of 300 for faster convergence has been used to solve the problem. The magnetic scalar potential drop across each reluctance element of the 2D
MEC model is calculated. The reluctance of the iron core elements changes as the machine begins to saturate. The B-H curve data for the laminated steel used in the stator and rotor cores are integrated with the iterative Gauss-Siedel algorithm used in the MEC model. This allows the MEC model to update the permeability of the reluctance network from the material B-H data when the iterative result does not satisfy the error criterion. An iterative approach is used to update the reluctance matrix when the MMF drop compared with previous iterations is within an acceptable error difference of magnitude $10^{-5}$. The updated reluctance matrix is used to determine the air gap flux through the magnetic circuit.

4.4 MEC Model Result Verification with 3D-FEA

4.4.1 TFM electromagnetic characteristics

The meshed model and the flux density distribution for the TFM under the loaded condition for completely aligned and unaligned positions are shown in Fig. 4.9. The model was meshed using a first-order mesh with 23056 nodes, 5241 line elements, 48642 surface elements, and 82462 volume elements. The sizing of the mesh used in the FEA was controlled and optimized to achieve good quality meshing while reducing the simulation time. Each design iteration in the FEA domain for one-half electrical cycle took approximately 300 seconds whereas the MEC model gave an estimate in approximately 80 secs. Solving the model with a second-order mesh drastically increases the FEA simulation time from 5 to 25 minutes. Table 4.1 documents the difference between the no-load and rated-load simulation results due to different mesh-order. For initial sizing of the machine, which requires multiple design iterations, the first-order mesh can be used for faster analysis.
4.4.2 Flux linkage distribution, EMF and winding inductance.

The MEC model for the Z-Core TFM was used to predict flux linkage under a no-load condition at different rotor positions as shown in Fig. 4.10. Flux linkages at different rotor positions under current excitations – unsaturated (20A), partially saturated (60A) and fully saturated (100A) are shown in Fig. 4.11. For different current excitations (0-100A), flux linkages at partially aligned (4°) and completely unaligned (6°) pole positions are shown in Fig. 4.12 and Fig. 4.13 respectively. The no-load back EMF for different rotor positions is shown in Fig. 4.14. Induced EMF in the coil conductor under different loading condition (40A and 60A) are shown in Fig. 4.15 and Fig. 4.16 respectively. Fig. 4.17 and Fig. 4.18 show the stator winding inductance under unsaturated (no-load) and saturated condition for half of the electrical cycle. The results show some discrepancy at higher currents due to magnetic saturation, however, the error is below 10%.
Figure 4.10: Flux linkage at 0A.

Figure 4.11: Flux linkage at 20A, 60A, and 100A (from bottom to top).

Figure 4.12: Flux linkage at 4° rotor position for currents 0-100 A.

Figure 4.13: Flux linkage at 6° rotor position for currents 0-100 A.

Figure 4.14: No-load EMF.

Figure 4.15: Induced EMF at 40A.
4.4.3 Electromagnetic torque characteristics

Torque was predicted using the virtual work method. For the Z-Core TFM, the cogging torque from MEC and FEA simulations are shown in Fig. 4.19 with the results showing a good agreement. In Fig. 4.20, the torque produced at different rotor positions for both the unsaturated (20 A, 60 A) and saturated (100 A) cases are illustrated. Fig. 4.21 and 4.22 show the torque produced at different current excitations (electrical loading) in partially aligned and completely unaligned pole positions respectively. At high currents when the machine saturates, there are some discrepancies in the average torque values. The transient simulation was conducted with the MEC-Simulink circuit simulator and compared with 3D FEA-Simulink coupled simulation. In the FEA-Simulink coupled simulation, the Simulink
circuit was used to regulate the sinusoidal current supplied to the 3D machine in Flux3D that provided the FEA torque. The regulated phase current and the generated torque from both simulations are shown in Fig. 4.23 and 4.24 respectively.

A 3D FEA-Simulink coupled simulation was carried out for the Z- Core TFM which takes around 12 hours for one design iteration, whereas the MEC-Simulink coupled simulation takes just 10 minutes. The difference in average torque values is less than 10%.

Figure 4.19: Cogging torque at no load condition.
Figure 4.20: The torque produced (from top to bottom) at 100A, 60A, and 20A.
Figure 4.21: The torque produced at 4° rotor position for currents 0-100A.
Figure 4.22: The torque produced at 6° rotor position for currents 0-100A.
4.4.4 Changes in machine geometry

The developed MEC model for the Quasi-U Core TFM is capable of handling geometric parameter variation to predict the changes in average torque. The magnetic loading of the TFM is changed by altering the dimensions of the TFM rotor thickness and magnet length, and their effect on torque production is shown in Fig. 4.25 and Fig. 4.26 respectively.

4.5 Conclusion

A single phase, 1 kW, 400 rpm transverse flux machine (TFM) for use in wind power applications has been analytically modeled. The 3D model of the machine was developed using Flux 3D software. A dynamic analytical model for the machine was developed using
2D MEC modeling that allowed dimensioning and performance analysis of the designed machine with reduced computational resources. The MEC modeling includes a permanent magnet, rotor, stator, stator winding and air gap leakages and variable air gap reluctances. The 2D MEC model consists of a 3D reluctance network and flux paths lumped together into a reduced order 2D network.

The 2D MEC model is nonlinear and includes the effects of magnetic saturation which includes an iterative algorithm to calculate the reluctance of iron core elements as increasing current drove them into saturation. The flux linkage and torque generated by the machine were obtained analytically and verified with results obtained from Finite Element Analysis (FEA).

The MEC model holds its merit under dimension and current variations. The transient simulation was also conducted with the MEC model where current regulation was done to obtain torque and regulated phase currents. A single phase 3D-FEA coupled simulation was also carried out to verify the results. Compared with FEA results, the analytical MEC model provided estimates with less than 10% error, while significantly reducing the computation time.
CHAPTER V
DESIGN OPTIMIZATION OF TFM

5.1 Introduction

The design optimization (synthesis) of electrical machines is a nonlinear multi-objective problem. Typical objectives, such as highest efficiency, lowest cost, and a minimum weight of active materials, have to be simultaneously met by a process in which the electromagnetic problem is solved with consideration of the mechanical, thermal, and material aspects.

The optimization aspects for electric machines can be grouped into:

- Surrogate models that satisfactorily capture the relationship between design objectives and inputs.
- Optimal Design Search algorithms, which lead to a global optimum with minimum computational effort.

Surrogate models drastically reduce the repetitive analysis efforts by approximating the relationship between machine design input parameters and output characteristics. The result is a simplified model, which can be then optimized faster. Surrogate models include

- Design Space Reduction: A reduction of the design space can be achieved by identifying the major design parameters, which have the most significant effect on machine performance. One approach commonly adopted for this purpose consists of a statistical
screening, which determines the sensitivity of main performance indices to the decision, i.e., independent, variables. A stepwise multi-regression-based screening technique was proposed in [75]. The technique was validated through a comparison of the Pareto front calculated by using the full set of decision variables and the front determined only by employing the significant variables identified by the screening technique.

- Response Surface Methodology (RSM): The RS methodology (RSM or, in short, RS) is a mathematical and statistical technique that finds the functional relationship between a response of interest $y$ and a set of input variables. In the interior PM (IPM) machine design RS optimization example described in [76], three geometrical parameters of the PM are selected as decision (independent) variables, and the objective was to optimize the constant power speed range.

- Space Mapping (SM): Space mapping (SM) is a mathematical technique for establishing relationships between fine and coarse models during an optimization process [97].

Optimal search algorithms can be divided into two categories: deterministic methods that find the optima algorithmically and stochastic methods which explore the solution space randomly

- Deterministic Methods: There is a large variety of deterministic methods applied to electric machine design optimization problems, among them the sequential unconstrained minimization technique (SUMT) is one of the most popular ones [77]. Other deterministic methods include an error-based optimization search [78], the Hooke–Jeeves method [79], the interval branch and bound method [80], the interior-reflective Newton method [81], a
combination of the sub-problem approximation method and the first-order method [82], and the inverse problem method [83].

- Stochastic Methods: Stochastic methods tend to need more evaluations of design candidates but are gradient free and, in principle, not trapped by local minima. Furthermore, population-based searches can take advantage of parallel computing [85]. Some of the notable stochastic methods include the Genetic Algorithm (GA) and Particle Swarm Optimization (PSO). GA and PSO have been widely applied to electric machine design optimization [86-94].

5.2 Selecting Design Variables for Optimization

The proposed TFM has a complex geometry with a dual stator single rotor setup. During the design consideration for this TFM, certain design and mechanical limits were set for meeting the application specifications and prototyping feasibility, which were discussed in detail in Chapter 3. Those design considerations also dictated the optimization variables and limits, which will be analyzed in this chapter.

Design variables that are most sensitive to torque production need to be identified and selected for the optimization to reduce the design search space. For the proposed TFM, the magnet length (LM) and thickness (HRY), is extremely sensitive to the torque production and machine active weight. The pole width (WR and WS) and length (LP), control the area through which the electromagnetic energy in the axial air gap is being exchanged between the rotor and stator. The slot height (CH) in the stator affects the current density and winding fill factor. The stator back iron (SBI) thickness needs to be optimized to ensure that magnetic saturation does not occur at the rated condition in the stator. Another important design consideration is the flux-focusing factor $K_f$ of the rotor, since flux-
concentrating magnets are to be used in the rotor. $K_f$ sets the magnetic loading of the machine which dictates the torque density and power factor performance and is defined by (5.1). Thus, making the variables HRY and LP, critical optimization variables. The rotor and stator parameters that are to be optimized are listed in Table 5.1.

$$K_f = \frac{A_m}{A_p} = \frac{HRY}{LP} \tag{5.1}$$

**Table 5.1: Design variables selected for optimization in the rotor and stator.**

<table>
<thead>
<tr>
<th>Geometric Parameter</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>LP</td>
<td>Pole Length</td>
</tr>
<tr>
<td>LM</td>
<td>Magnet Length</td>
</tr>
<tr>
<td>HRY</td>
<td>Rotor/Magnet Thickness</td>
</tr>
<tr>
<td>WR</td>
<td>Rotor Pole Width</td>
</tr>
<tr>
<td>$\tau$</td>
<td>Pole Pitch</td>
</tr>
<tr>
<td>WS</td>
<td>Stator Pole Width</td>
</tr>
<tr>
<td>CH</td>
<td>Slot Height</td>
</tr>
<tr>
<td>SBI</td>
<td>Stator Back Iron Thickness</td>
</tr>
</tbody>
</table>

The pole lengths (LP) and stator back iron (SBI) thickness were all selected to equal each other to ensure that pole face and back iron area were kept constant to prevent magnetic saturation. The slot opening (SO) of the stator was equivalent to LM. The optimum values for pole pitch and pole width were already determined in Chapter 3, using 3D-FEA based optimization which was computationally intensive and time-consuming. It required a lot of simulations to be manually designed and post-processed for data. To optimize the remaining four geometric parameters of Table 5.1, along with current loading
(number of turns and peak RMS current), a different approach was taken that was computationally less intensive and allowed faster analysis.

5.3 Optimization using MEC and PSO

An improved analytical model for the proposed TFM based on the MEC method was developed in Chapter IV. The unique geometry of the proposed TFM structure required a new MEC model to be developed that takes into consideration magnetic saturation and the changes in the air gap flux pattern due to alternate magnet polarity in the rotor along with the fringing flux paths from the stator cores. The developed MEC is computationally faster and less complex to develop than conventional 3D MEC models for TFM and AFM and does not compromise on accuracy for predicting machine performance.

In order to optimize the remaining geometric parameters from Table 5.1, the developed MEC model for the TFM was used in tandem with the Particle Swarm Optimization (PSO) algorithm, to perform a multi-variable multi-objective optimization. The output of this proposed analytical tool would directly present optimized dimensions for the remaining design parameters much faster and would require no post-processing in FEA domain.

5.3.1 Particle swarm optimization

Particle swarm optimization (PSO) is an evolutionary population based stochastic optimization technique inspired by the behavioral pattern of birds flocking and fish schooling [95-96]. It is a computational efficient search algorithm irrespective of the size and nonlinearity of the problem. PSO can converge to the optimal solution in many problems where other algorithms may fail to converge [97]. The optimal search is achieved through a combination of self and swarm knowledge. Some of the advantages of PSO over
other optimization algorithm are: 1) PSO is robust to the values of its running coefficients; 2) computationally more efficient; 3) no derivatives required in the algorithm; 4) can be coupled with other optimization techniques; 5) no initial solution required for starting the iteration process.

PSO has been successfully applied in the field of electric machine design optimization. In [98], an optimal design of a TFM for reducing the cogging torque was done using PSO. Swarm intelligence was used to optimize induction machines [99], synchronous reluctance motors for traction applications [100], surface-mounted PM motors with segmented poles [101], linear machines [102-103] and for an IPM generator [104].

In this section, the PSO algorithm has been used along with the MEC model to optimize the single-phase TFM for maximum torque density at a rated power of 1 kW [105]. The stator pole length, which is equal to the rotor pole length (LP), magnet length (LM) and rotor thickness (HRY) are the variables used to define the search space for the optimization problem. Torque production in the TFM is extremely sensitive to these three aforementioned geometric parameters as they affect both the electrical and magnetic loading of the machine. The MEC model creates objective functions that analytically provide the machine weight and torque for the required set of machine parameters, without doing any Finite Element Analysis (FEA), thus saving computational time. Then a cost function is used to identify the optimum set of design parameters required to achieve maximum torque density while satisfying the design constraints. The FEA-based model developed in Flux3D is used to verify the results obtained from the optimized analytical model.
PSO emulates the behavior of a swarm of bees trying to locate places with the highest density of flowers. It is a stochastic search algorithm which explores the solution space randomly. They are gradient free and not trapped by local minima. In addition to that, they can also take advantage of parallel computing seen in other population-based searches. Fig. 5.1 shows the general flowchart of the PSO algorithm. If the optimization problem has \( N \) variables, a swarm of \( P \) particles is initialized in which each particle gets assigned a random position in the \( N \)-dimensional search space such that each particle’s position corresponds to a possible solution.

![PSO Flowchart](image)

**Figure 5.1:** General PSO flowchart.

PSO uses only two Equations, (5.2) and (5.3) in its algorithm, where each particle \( P \) is associated with a position \( X_i \) and velocity vector \( V_i \), which explores through the solution space for optimal solution. In each iteration step, \( gbest \) is the best particle position based on overall swarm’s experience and \( pbest \) is the best particle position achieved based on its own experience. The swarm is manipulated according to the following two Equations given in (5.2) and (5.3) where \( d=1,2,...,D \), and \( i=1,2,...,N \). \( N \) is the size of the swarm, \( c_1 \) and \( c_2 \) are two positive acceleration constants, namely social and cognitive parameters, \( r1 \) and \( r2 \) two random numbers distributed within the range \([0,1]\), \( t \) is the iteration number, \( \Delta \tau = 1 \),
and $\omega$ is inertia weight. Using (5.2), the particle updates its velocity according to its previous velocity and the distances to its current position from both its own best historical position and the best positions of the neighbors in every iteration step, and then it flies towards a new position given by (5.3). The PSO output is evaluated by the cost function given in Equation (5.4) where $K_1, K_2$ and $K_3$ are the coefficients setting precedence on the output of interest at the required power levels.

$$V_{i,d}(t) = \omega V_{i,d}(t-1) + \frac{c_1 r_1 (p_{best_{i,D}} - X_{i,D}(t-1))}{\Delta \tau} + \frac{c_2 r_2 (g_{best_{i,D}} - X_{i,D}(t-1))}{\Delta \tau}$$

\hspace{1cm} (5.2)

$$X_{i,d}(t) = (X_{i,d}(t-1) + V_{i,d}(t)\Delta \tau)$$

\hspace{1cm} (5.3)

cost function

$$= K_1(Torque\ density) + \frac{K_2}{Magnet\ weight}$$

\hspace{1cm} (5.4)

$$+ K_3(TorquePerAmpere)$$

5.3.2 Optimization procedure

For optimizing the TFM, it is important to select the optimizing variables that are sensitive to the performance parameter that has to be optimized. Optimizing for maximum torque density means the machine weight and torque are the output of interest. Magnet, rotor and stator poles lengths are the most sensitive geometric parameters that directly affect the torque production and active weight of the TFM. Selecting more optimization variables can give more freedom, but it would be difficult to balance the relationship between these variables and it would take more computation time. There exists a trade-off between the electromagnetic torque and the machine weight. The PSO algorithm is tasked
to seek the optimum combinations of geometric parameter size to obtain the maximum torque-weight ratio. The swarm changes the length of the optimizing variables and the MEC provides objective functions to predict the torque and weight of the TFM. The overall design strategy is shown in the flowchart of Fig. 5.2. There are geometric design constraints like machine outer diameter (MOD) and machine axial length (MXL) as well as constraints associated with power converters (inverter) such as current density, DC Bus voltage, and output power. The geometric constraints arise from the machine housing and packaging restriction and are set at 225 mm and 105 mm for MOD and MXL respectively. The rating for the available DC Bus voltage (48V) as well as the current handling capability of the converter switches (200A_{peak}), along with the insulating material and cooling limitations for stator windings (current density <5A/mm²), impose further design constraints. This takes into consideration the number of turns and slot height (CH) for the coils which are determined at the initial stage of the optimization process, even before the machine electromagnetics are solved analytically. Designs that do not satisfy the constraints give a high negative value at the output and are discarded. As a result, only optimizing variables that meet the design specifications are evaluated through the cost function at the output.
The constraints imposed by laboratory testing facilities, covering both physical size as well as torque handling capabilities, limit the TFM to be between 1 to 1.2 kW, at 400 rpm. The cost function output is associated with the following machine output of interest: torque density, magnet weight (active weight) and torque per ampere. It is desirable to have a machine design that has a high torque density and high torque per ampere while using less magnet material. The weights are assigned to the coefficients ($K_1$, $K_2$ and $K_3$) in the cost function depending on the sensitivity of the desired output. By altering the weight in the cost function, an optimum design based on the highest torque density for the given design parameters was obtained.

5.3.3 MEC-PSO optimization results

Analytically obtained results using MEC was verified using 3D-FEA results from Flux3D for different rotor positions as shown in Chapter 4. The PSO algorithm was configured using the parameters given in Table. 5.2, and the convergence of the optimizing
variables HRY, LM, and LP are shown in Figs. 5.3-5.5 respectively. Table 5.4 verifies the optimized machine performance with FEA using optimum design values from Table. 5.3. The error in the result estimation is less than 10%.

The computational time for carrying out an FEA based optimization for around 400 combined iterations for the three variable search space would take around 40 hours whereas the same optimization using the MEC-PSO tool would require around 18 hours. This corresponds to around 50% reduction in computation time associated with the optimization process.

Table 5.2: PSO characteristics.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of particles</td>
<td>30</td>
</tr>
<tr>
<td>Initial velocity of agent</td>
<td>1</td>
</tr>
<tr>
<td>Acceleration constants</td>
<td>0.5</td>
</tr>
<tr>
<td>Number of iterations</td>
<td>50</td>
</tr>
<tr>
<td>Number of variables</td>
<td>3</td>
</tr>
</tbody>
</table>

Table 5.3: Optimized variables and levels.

<table>
<thead>
<tr>
<th>Design variables</th>
<th>Range of variables</th>
<th>Optimal value</th>
</tr>
</thead>
<tbody>
<tr>
<td>HRY</td>
<td>20-28 mm</td>
<td>26 mm</td>
</tr>
<tr>
<td>LM</td>
<td>8-12 mm</td>
<td>10 mm</td>
</tr>
<tr>
<td>LP1</td>
<td>7-14 mm</td>
<td>8 mm</td>
</tr>
</tbody>
</table>
Table 5.4: Verification of the optimized model with FEA.

<table>
<thead>
<tr>
<th></th>
<th>FEA</th>
<th>MEC-PSO</th>
</tr>
</thead>
<tbody>
<tr>
<td>Average Torque (Nm)</td>
<td>25.6</td>
<td>24.28</td>
</tr>
<tr>
<td>Machine weight (kg)</td>
<td>6.2</td>
<td>6.2</td>
</tr>
<tr>
<td>Torque density (Nm/kg)</td>
<td>4.13</td>
<td>3.91</td>
</tr>
</tbody>
</table>

A single phase, 1 kW, 400 rpm TFM has been designed and optimized for use in wind power applications using the MEC-PSO method. MEC modeling was used to perform dimensioning and performance analysis of the initially designed machine. The flux linkage and torque generated by the machine were obtained analytically and verified with results obtained from 3D-FEA. The PSO algorithm was used to optimize performance-sensitive parameters of the TFM for attaining maximum torque density and verified using 3D-FEA. The dynamic MEC-PSO model developed in this section is an attractive alternative to 3D-FEA based TFM design and optimization, by being computationally less intensive and time efficient while giving accurate estimates for the designed machine figures of merit.

5.3.4 3D-FEA based performance analysis of finalized Z-Core TFM design

Based on these trends and design considerations, a 30-pole, 400-rpm, 1 kW machine has been designed. Finalized key design parameters are shown in Table 5.5. The electromagnetic analysis was performed in Flux 3D. The model meshed with second-order elements. The machine was designed to have a maximum flux density of 2 T in the cores, as demonstrated in Fig. 5.6. The back-EMF, cogging torque and no-load air gap flux density distribution are shown in Figs. 5.7-5.9 respectively.
Figure 5.6: 3D-FEA electromagnetic analysis for Z-Core TFM.

Figure 5.7: Back-EMF waveform for Z-Core TFM.

Figure 5.8: Cogging torque waveform for Z-Core TFM.

Figure 5.9: Axial air gap flux distribution due to PM for Z-Core TFM.

Table 5.5: Final design and performance characteristics.

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output power</td>
<td>1 kW</td>
<td>Rotor Height (HRY)</td>
<td>26 mm</td>
</tr>
<tr>
<td>Rated speed</td>
<td>400 rpm</td>
<td>Number of turns (N)</td>
<td>13</td>
</tr>
<tr>
<td>Current density</td>
<td>5 A/mm²</td>
<td>Slot height (CH)</td>
<td>16 mm</td>
</tr>
<tr>
<td>Air gap length</td>
<td>1 mm</td>
<td>Pole number (p)</td>
<td>30</td>
</tr>
<tr>
<td>Axial length (MXL)</td>
<td>102 mm</td>
<td>DC bus</td>
<td>48 V</td>
</tr>
<tr>
<td>Outer diameter (MOD)</td>
<td>225 mm</td>
<td>Frequency</td>
<td>100 Hz</td>
</tr>
<tr>
<td>Rotor core length (LP)</td>
<td>8 mm</td>
<td>RMS current</td>
<td>107 A</td>
</tr>
<tr>
<td>Magnet length (LM)</td>
<td>10 mm</td>
<td>Average Torque</td>
<td>25.6 Nm</td>
</tr>
<tr>
<td>Shaft length</td>
<td>68.5 mm</td>
<td>Power factor</td>
<td>0.34</td>
</tr>
<tr>
<td>Stator pole width (Wₛ)</td>
<td>8.25 mm</td>
<td>Weight</td>
<td>6.20 kg</td>
</tr>
<tr>
<td>Rotor pole width (Wᵣ)</td>
<td>8.25 mm</td>
<td>Torque density</td>
<td>4.13 Nm/Kg</td>
</tr>
</tbody>
</table>
5.4 Conclusion

The complete optimization of the critical rotor and stator geometry parameters were completed in this chapter using the MEC-PSO analytical tool. This allowed for a significant reduction in computation time compared to 3D-FEA, to get the final optimized Z-core TFM design.
CHAPTER VI
MAGNET-FREE TFM WITH ROTARY TRANSFORMER

6.1 Introduction

The proposed TFM was designed using non-rare-earth based PM to address the concerns associated with the supply and price volatility of rare-earth based PM. The logical step forward was to come up with a design for a magnet-free TFM.

The concept of the rotary transformer has been used to eliminate brushes and slip rings in doubly fed induction machines. The rotary transformer allowed access to the rotor circuit without any mechanical contact, improving the reliability of the machine. In [106], rotary transformers were used for contactless transfer of energy to establish field current in the rotor, which completely eliminated rare-earth magnets in synchronous motors used for hybrid vehicle applications. However, this concept was never extended to a complex machine like TFM.

In this chapter, a novel concept of integrating the rotary transformer within the proposed TFM design is explored to completely eliminate the use of magnets from the TFM [109].

6.2 Proposed TFM with Rotary Transformer

A double-sided TFM with the embedded rotary transformer is proposed that is modular in structure and free of PM materials. The 3D model of the TFM having Z-core stators, a
rotary transformer and rotor with field coils are shown in Fig. 6.1. The primary side of the rotary transformer is integrated with the stator and the secondary is embedded on both sides of the rotor. The rotor field windings are excited by using the contactless rotary transformer. The secondary voltage of the rotating transformer is rectified using a bridge rectifier to maintain a constant DC voltage across the field winding. This results in a unidirectional flux being set up in the rotor which replicates the behavior of PM embedded in the rotor.

Figure 6.1: 3D model of the proposed rotary transformer based TFM.

The proposed TFM has a modular structure with ring windings. Modularity is a very important advantage of the proposed TFM topology. Each phase is an independent module which makes the machine very attractive for mass production since a wide range of power can be covered by placing an adequate number of modules together. Fig. 6.2 shows a 2D cross-sectional view of a single pole of the proposed TFM where the flux paths are transverse (perpendicular) to the rotating motion of the rotor.
6.2.1 Rotor structure with transformer secondary and field coils

Each phase module consists of a rotor in the middle with stator cores located on both sides of the rotor as shown in Fig. 6.2. The structure of the rotor integrates the transformer secondary winding slots as well as the slots for the field coils as shown in Fig. 6.3. The rotor windings produce a unidirectional field that aids in the buildup of flux in the middle core of the rotor. The flux induced in the field coils are oriented such that they are either pointing to or away from each other. The adjacent rotor cores are arranged to align with the stator cores to the left, middle and right, respectively, as shown in Fig. 6.2 to close the magnetic circuit. The rotor pole has the same width to match that of the stator.
6.2.2 Stator structure with transformer primary

The stator consists of two alternate Z-cores with double active sides facing the rotor via two axial air gaps and two ring windings in the upper and lower stator slots. The primary winding slots of the rotary transformer are integrated with the stator cores as shown in Fig. 6.4. The stator orientation minimizes the exposed windings to reduce the leakage fluxes. The double-sided windings maximize the stator core utilization and make the machine highly fault tolerant. The ‘ring’ winding couples each stator core to the entire armature ampere-turns, and thus avoids the limitation of the “BIL” principle of force production. As a result, high torque can be achieved by increasing the pole number without sacrificing the electric loading.
In this section, one module unit is designed and optimized to be used for direct drive wind turbine applications. The motor design parameters along with the design specifications and restrictions are summarized in Table 6.1.

Table 6.1: Rotary transformer based TFM design parameters.

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Design and Performance Specifications</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of poles</td>
<td>30</td>
<td>Peak Torque</td>
<td>30 Nm</td>
</tr>
<tr>
<td>Output power</td>
<td>0.6 kW</td>
<td>DC bus voltage</td>
<td>400 V</td>
</tr>
<tr>
<td>Rated speed</td>
<td>400 rpm</td>
<td>Excitation frequency</td>
<td>100 Hz</td>
</tr>
<tr>
<td>Current density</td>
<td>5 A/mm²</td>
<td>Peak current</td>
<td>105 A</td>
</tr>
<tr>
<td>Air gap length</td>
<td>1 mm</td>
<td>Air gap flux density</td>
<td>1.2 T</td>
</tr>
<tr>
<td>Stator outer diameter</td>
<td>255 mm</td>
<td>Axial length</td>
<td>82 mm</td>
</tr>
<tr>
<td>Rotor inner diameter</td>
<td>142 mm</td>
<td>Rotor outer diameter</td>
<td>225 mm</td>
</tr>
<tr>
<td><strong>Stator winding</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of phases</td>
<td>1</td>
<td>Winding type</td>
<td>Ring</td>
</tr>
<tr>
<td>Number of coils</td>
<td>2</td>
<td>Turns per coil</td>
<td>11</td>
</tr>
<tr>
<td>Coil length</td>
<td>585 mm</td>
<td>Coil area</td>
<td>270 mm²</td>
</tr>
<tr>
<td>Fill factor</td>
<td>0.6</td>
<td>Coil material</td>
<td>Copper</td>
</tr>
<tr>
<td><strong>Field winding</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of coils</td>
<td>60</td>
<td>Winding type</td>
<td>Ring</td>
</tr>
<tr>
<td>Coil length</td>
<td>3 mm</td>
<td>Coil Area</td>
<td>40 mm²</td>
</tr>
<tr>
<td>Fill factor</td>
<td>1</td>
<td>Coil material</td>
<td>Copper</td>
</tr>
</tbody>
</table>
6.3 Rotary Transformer and Field Power Converter Design

Rotary transformers have been used before [106-108] to establish a field in the rotor via contactless transfer of energy. However, this concept has never been extended to TFM where it could eliminate the use of PM completely from the machine design. In this TFM design, four transformer pairs have been used. The secondary transformer cores are embedded in the rotor, while the primary transformer cores are integrated with the stator cores. The secondary winding of the rotary transformer is physically coupled to the field windings in the rotor through a field power converter.

6.3.1 Rotary transformer design

Rotary transformer designs are different from conventional transformers due to the presence of an air gap that allows movement between primary (stator) and secondary (rotor) windings as shown in Fig. 6.5.

Figure 6.5: Transformer primary and secondary core with ring windings.

An axial gap rotary transformer with C-shaped ferrite cores are used in this design with a nominal air gap of 1mm. The transformer primary cores are designed in such a way that it fits into the proposed TFM stator core structure used in Chapter 3. The transformer secondary cores are embedded in the rotor by replacing the magnets buried between the
rotor cores. Removing the magnets also provides the slots for the field coils located between the secondary of the transformer cores.

![Ideal Transformer Diagram](image)

Figure 6.6: The equivalent circuit of an ideal transformer.

Fig. 6.6 shows an equivalent circuit of a transformer where $L_m$ is the magnetizing inductance and $R_{fe}$ is the resistance to represent the core loss. Since the rotary transformer has an air gap between the primary and secondary core, the $L_m$ tends to be high which would require higher current for magnetization. In order to reduce the amount of current going through the magnetizing branch, the magnetizing reactance $X_m$, which depends on the dimensions of the transformer, the number of turns in the primary coil and the material properties, should be increased.

The expression for $X_m$ can be described as:

$$X_m = \omega L_m = \frac{\omega N^2 \mu_0 A}{g} \quad (6.1)$$

where $\omega$ is the applied frequency, $N$ is the number of turns, $\mu_0$ is the air permeability, $A$ is the area of the transformer faces, $g$ is the airgap between the transformer primary and secondary. ‘Ring’ windings are used in the transformers, where the primary side is excited at $\omega = 100$ kHz. Ferrite has been chosen as the material for the transformer cores to be able operate at higher frequencies without having significant core losses.
The primary and secondary winding resistances, magnetizing inductance, and the leakage inductance of the rotary transformer is calculated through FEA. The parameters of the equivalent circuit shown in Fig 6.6 are determined and their values summarized in Table 6.2.

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Core Material</td>
<td>Ferrite</td>
</tr>
<tr>
<td>Primary Number of Turns</td>
<td>4</td>
</tr>
<tr>
<td>Primary Resistance $R_p$</td>
<td>0.014</td>
</tr>
<tr>
<td>Primary Leakage inductance $L_p$</td>
<td>61.1uH</td>
</tr>
<tr>
<td>Secondary number of turns</td>
<td>40</td>
</tr>
<tr>
<td>Secondary resistance $R_s$</td>
<td>1.435 Ω</td>
</tr>
<tr>
<td>Secondary Leakage inductance $L_s$</td>
<td>611 mH</td>
</tr>
<tr>
<td>Magnetizing inductance</td>
<td>379.74H</td>
</tr>
</tbody>
</table>

6.3.2 Field power converter circuit

An H-bridge inverter is connected to the stationary primary side of the rotary transformer primary with a DC link voltage of 400V as shown in Fig. 6.7. The rotating secondary side of the transformer is electrically coupled to the field coil through a bridge rectifier. The LC filter ensures constant DC current is maintained in the field coil.

Figure 6.7: Field power converter circuit with rotary transformer and field coil.
6.4 FEA Results

The 3D FEA model was designed for two machines: a PM based TFM and a rotary transformer based TFM. Both machines have an identical outer diameter and axial height. The PM in the rotor of the TFM are replaced with field coils (FC) and rotary transformers are integrated into the existing rotor and stator structure of the 3D model without affecting its critical dimensions such as airgap length, pole number, pole length, and pole face area.

The no-load flux linkage and EMF were determined for different rotor positions for both designs. At higher speeds, it is important to have the capability to reduce the EMF. For rated operating speed the proposed TFM achieves the same EMF with the TFM using PM. This result suggests that the proposed machine can achieve wider speed operation as the EMF can be reduced as a function of the field current produced by the rotary transformer.

Torque was also determined for different rotor positions while exciting the coil conductor. The field coil current is controlled such that both of the designs produce similar peak torque and power output at the same operating speeds.

6.4.1 Flux linkage distribution and no-load EMF

Flux linkage under a no-load condition at different rotor positions for both of the designs is shown in Fig. 6.8. Flux linkage in the rotary transformer based TFM is comparable to that of the PM based TFM. No-load EMF for different rotor positions is shown in Fig. 6.9. The EMF waveform is almost sinusoidal and hence applying a sinusoidal current to the main stator winding would ensure a smooth torque. The induced voltage in the transformer secondary winding is rectified to maintain a constant DC field on the rotor as shown in Fig. 6.10.
6.4.2 Torque characteristics

The main windings in the stator are driven by a single phase 105A peak sinusoidal current. There are two components in the torque generated: a) the electromagnetic torque which comes from the non-sinusoidal no-load EMF, b) the cogging torque which comes from the interaction between the rotor field and stator slots. The torque produced at different rotor positions is obtained for both no-load (cogging torque) and loaded conditions as shown in Figs. 6.11 and 6.12 respectively. The back-EMF of the rotary transformer based TFM is a function of the field current as shown in Fig. 6.13, which can be controlled to improve the torque-speed curve for wide speed operation as shown in Fig. 6.14.
6.5 Conclusion

A single phase, 0.6 kW, 400 rpm TFM for use in wind power applications has been designed using a rotary transformer. The proposed TFM is free of PM materials and does not sacrifice peak power and torque ratings when compared to an equivalent PM based TFM. The designed rotary transformer can be retrofitted into an existing PM based TFM geometry without affecting its critical dimensions such as airgap length, pole number, pole length, and pole face area.
CHAPTER VII
TORQUE RIPPLE AND VIBRATION PERFORMANCE OF TFM

7.1 Introduction

Since TFMs suffer from the issue of cogging torque as discussed in Chapter 2, it is important to address this issue electromagnetically at the design phase. The effect of cogging torque will get carried to the torque ripple content of the three-phase TFM and will affect the vibration performance of the machine as well. In this chapter, the torque ripple and vibrational performance for the proposed TFM are analyzed and a unique design based solution for addressing the issue of torque ripple is presented.

7.2 Three-phase Analysis of the Proposed TFM

The TFM proposed in this research has already been introduced in detail in Chapter III. One of the key advantages of the TFM is its modularity, which means that each phase is an independent module. This feature is not present in conventional RFMs. Additional unit modules can be stacked in the axial direction to get more power out of TFMs. In order to obtain a three-phase version of the machine, three individual TFM modules would have to be stacked axially, with the rotor PMs having an angular displacement of 120 electrical degrees with each other as shown in Fig. 7.1. For the 30-pole TFM module designed during
this research, $8^0$ mechanical is equivalent to $120^0$ electrical, thus each rotor has a mechanical $8^0$ shift among each other.

![Figure 7.1](image.png)

Figure 7.1: (a) Rotor PMs with 120 electrical degrees displacement, (b) Three-phase TFM with three axially stacked modules.

3D-FEA has been used to analyze the three-phase TFM performance under no-load and load conditions at rated speed. The electromagnetic performance results are listed in Table 7.1.

<table>
<thead>
<tr>
<th>Cogging Torque (Peak to peak) (Nm)</th>
<th>Back-EMF constant $K_E$ (vs/rad)</th>
<th>Average Torque (Nm)</th>
<th>% Torque Ripple</th>
<th>Rated Power (kW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>15</td>
<td>0.28</td>
<td>79.5</td>
<td>8.24</td>
<td>3.3</td>
</tr>
</tbody>
</table>

Cogging torque and high amplitude back-EMF harmonics associated with the proposed TFM design are critical issues that affect the torque ripple and vibration performance of the TFM. Cogging torque is the zero average pulsating torque caused by the tendency of the PM to align with the stator iron. This appears whenever magnetic flux travels through a varying reluctance, which is during the interaction between the PM and the stator slots. A simplified expression of cogging torque $T_{cogg}$ is given by Equation (7.1), where $\phi_g$ is
the magnet flux crossing the air gap and $R_t$ is the total reluctance through which the flux passes.

$$T_{co,gg} = -\frac{1}{2} \Phi_g \frac{2 dR_t}{d\theta}$$  \hspace{1cm} (7.1)

Torque ripple consists of two components: electromagnetic torque fluctuation and cogging torque. Electromagnetic torque ripple is caused by the harmonic interaction between the back-EMF and the phase currents associated with the motor electrical dynamics. The instantaneous electromagnetic torque for a 3-phase PMSM can be given by Equation (7.2) [110]

$$T_e = \sum_{x=a,b,c} \frac{e_x i_x}{\omega} - \frac{1}{2} \Phi_g \frac{2 dR_t}{d\theta} + \frac{1}{2} i^2 \frac{dL}{d\theta}$$  \hspace{1cm} (7.2)

where $e_x$ and $i_x$ represent the phase back-EMF and phase current, respectively, the second term in the expression represents the cogging torque and the third term represents the reluctance torque, where $L$ is the inductance in the coil. The first term in (7.2) is known as the electromagnetic torque or the mutual torque. The motor can produce constant electromagnetic torque only if the part of the flux through the stator windings due to the rotor field, known as mutual flux, is purely sinusoidal. This also requires sinusoidal spatial distribution of either the stator windings, or of the field due to rotor magnets. In practice, the perfect sinusoidal distribution is not achievable and the mutual flux contains higher harmonics and causes a ripple in the steady state torque in response to a purely sinusoidal current excitation. The torque ripple cannot be separated from the cogging torque; most of the researchers mentioned these two issues together [111-113]. Therefore, the design
methods developed for reducing cogging torque can also be considered as methods for reducing torque ripple.

7.3 Stator-pole shaping to Improve Torque Ripple Performance

The back-EMF harmonics of the designed TFM expressed as a percentage of the fundamental is shown in Fig. 7.2. The harmonic orders which are multiples of 3 get canceled out since the phase windings have a Y-connection. The torque ripple harmonic content is shown in Fig. 7.3. The TFM is excited with a pure sinusoidal current, and the fundamental of the current reacts with the back-EMF harmonics to generate the torque ripple components. The fundamental component of the current reacts with the back-EMF harmonic orders, 5\textsuperscript{th} and 7\textsuperscript{th}, 11\textsuperscript{th} and 13\textsuperscript{th}, to generate the torque ripple harmonic orders, 6\textsuperscript{th} and 12\textsuperscript{th} respectively. The complex dual stator geometry results with non-sinusoidal spatial field distribution, resulting in high amplitude low order back-EMF harmonics and consequently contributes to a high torque ripple in the proposed TFM as listed in Table 7.1.

![Figure 7.2: Back-EMF harmonic contents expressed as a percentage of the fundamental.](image)

![Figure 7.3: Torque ripple harmonic content.](image)
In order to compensate for the cogging torque and torque ripple issue, two design-based solutions are explored in the form of stator pole-shaping as shown in Fig. 7.4. The motivation behind the pole-shaping is to alter the spatial air gap field distribution from a design point of view, thereby reducing the cogging torque and the amplitude of the low order back-EMF harmonics. However, introducing such features in the active air gap interface will also affect the back-EMF constant $K_E$, and will lead to a penalty in terms of average torque output. Two new geometric parameters, $h_{sx}$, which is used to control the slant pole height, and $r_{sx}$, which controls the dent of the inverted-U in the stator pole, are introduced. The optimum dimensions for the newly introduced variables that lead to an acceptable trade-off between loss in average torque and decrease in cogging torque and torque ripple will be determined during this analysis. The limit for the dimensions for $h_{sx}$ and $r_{sx}$ (0.75mm to 1.5mm) is dictated by the manufacturing limitations and pole-tip height for the center pole.

Figure 7.4: (a) Baseline stator core (b) Stator core with slant-pole (c) Stator core with inverted-U pole.

The effect of introducing the pole shaping techniques on the dominant back-EMF harmonics, for different dimensional lengths of $h_{sx}$ and $r_{sx}$ is shown in Fig. 7.5 and the corresponding effect in torque ripple harmonic content are shown in Fig. 7.6. Both the pole shaping features show a reduction in dominant back-EMF harmonics and subsequent
decrease in the dominant 6\textsuperscript{th} and 12\textsuperscript{th} harmonic order of the torque ripple by more than 50%.

Figure 7.5: Comparison of back-EMF harmonic content with varying pole-shaping parameter (in mm) (a) \( hsx \) (b) \( rsx \).

Figure 7.6: Comparison of torque ripple harmonic content with varying pole-shaping parameter (in mm) (a) \( hsx \) (b) \( rsx \).

The impact of the pole-shaping geometric parameters \( hsx \) and \( rsx \), on back-EMF and torque ripple harmonics also suggests its negative impact on the back-EMF constant \( K_E \), and consequently the average torque output. In order to quantify the trade-off between average torque loss and torque ripple improvement, a new performance index is defined. The ratio of the loss of average torque to the reduction in torque ripple, when compared to
the baseline model, serves as an indicator to identify a potential design candidate. The lower the ratio means the rate of torque loss is less compared to the rate of torque ripple improvement.

Table 7.2: Comparative analysis of no-load and load performance under rated speed for the three-phase TFM with pole-shaping.

<table>
<thead>
<tr>
<th>Model</th>
<th>Ke (vs/rad)</th>
<th>Cogging Torque (Peak to Peak) (Nm)</th>
<th>Average Torque (Nm)</th>
<th>% Torque Ripple</th>
<th>DTavg / Tripple</th>
</tr>
</thead>
<tbody>
<tr>
<td>Baseline</td>
<td>0.28</td>
<td>15.4</td>
<td>79.4</td>
<td>8.24</td>
<td></td>
</tr>
<tr>
<td>hsx=0.75mm</td>
<td>0.24</td>
<td>8.84</td>
<td>70.5</td>
<td>3.77</td>
<td>0.20</td>
</tr>
<tr>
<td>hsx=1mm</td>
<td>0.23</td>
<td>7.7</td>
<td>67.3</td>
<td>3.69</td>
<td>0.27</td>
</tr>
<tr>
<td>hsx=1.25mm</td>
<td>0.22</td>
<td>6.7</td>
<td>64.8</td>
<td>3.04</td>
<td>0.29</td>
</tr>
<tr>
<td>hsx=1.5mm</td>
<td>0.21</td>
<td>6.4</td>
<td>62.7</td>
<td>2.6</td>
<td>0.30</td>
</tr>
<tr>
<td>rsx=0.75mm</td>
<td>0.23</td>
<td>8.35</td>
<td>67.7</td>
<td>3.7</td>
<td>0.26</td>
</tr>
<tr>
<td>rsx=1mm</td>
<td>0.22</td>
<td>6.87</td>
<td>64.8</td>
<td>3.4</td>
<td>0.31</td>
</tr>
<tr>
<td>rsx=1.25mm</td>
<td>0.21</td>
<td>5.68</td>
<td>62.8</td>
<td>3.4</td>
<td>0.35</td>
</tr>
<tr>
<td>rsx=1.5mm</td>
<td>0.20</td>
<td>5.29</td>
<td>59.5</td>
<td>3.5</td>
<td>0.43</td>
</tr>
</tbody>
</table>

From Table 7.2, it could be seen that stator core with slant pole, with pole shaping dimension (hsx) of 0.75mm, is the most promising candidate to reduce torque ripple while preserving the average torque output.

7.4 Structural and Vibrational Analysis of TFM

7.4.1 Vibration in PM machines

It is very important to consider the vibration performance of the electrical machine during the design process. The machine vibration is primarily due to the eccentric position of the rotor with respect to the stator bore [119]. The rotor eccentricity can be caused as a
result of imperfection in rotor assembly which leads to shaft misalignment. Also, the unbalanced magnetic pull, if present in a motor even with the perfectly aligned shaft, can create the rotor eccentricity [119]. The three categories of noise and vibration in electric machines are discussed in [114-116]. Electromagnetic (EM) vibration and noise are associated with parasitic effects due to higher space and time harmonics, eccentricity, phase unbalance, slot openings, magnetic saturation, and magnetostrictive expansion of the core laminations [119]. The vibration of the electric machine is a function of two parameters, namely the mode number and the frequency. The condition of “resonance” arises where the frequency of the harmonic content of the exciting force is similar to the natural frequencies of the machine. Electromagnetic vibration and noise are caused by the generation of electromagnetic fields.

Both the stator and the rotor excite magnetic flux density waves in the air gap. If the stator produces a $B_{ms} \cos(\omega_1 t + k\alpha + \varphi_1)$ magnetic flux density wave and the rotor produces a $B_{mr} \cos(\omega_2 t + l\alpha + \varphi_2)$ magnetic flux density wave, then the magnetic stress wave in the airgap is proportional to the product of the flux densities given, as [117]

$$\frac{1}{2} \left[ B_{ms} B_{mr} \cos[(\omega_1 + \omega_2)t + (k + l)\alpha + (\varphi_1 + \varphi_2)] + B_{ms} B_{mr} \cos[(\omega_1 - \omega_2)t + (k - l)\alpha + (\varphi_1 - \varphi_2)] \right]$$

(7.3)

where $B_{ms}$ and $B_{mr}$ are the amplitudes of the stator and rotor magnetic flux density waves, $\omega_1$ and $\omega_2$ are the angular frequencies of the stator and rotor magnetic fields, $\varphi_1$ and $\varphi_2$ are the phases of the stator and rotor magnetic flux density waves, and $k=1,2,3\ldots$ and $l=1,2,3\ldots$
The magnetic stress wave in the air gap along with the slots, distribution of windings in the slots, input current waveform distortion, air gap permeance fluctuations, and phase unbalance give rise to mechanical deformations and vibrations [119].

For the novel TFM topology proposed in this dissertation, the electromechanical energy conversion due to the interaction between the magnetic fields of PMs and armature conductors takes place in the axial airgap on both sides of the rotor. The tangential component of the electromagnetic force is responsible for the generated torque. The axial or normal component of the electromagnetic force between the rotor magnets and the stator teeth causes the vibration of the stator structure that is propagated as vibration in TFMs. In this section, the vibration characteristics (natural frequencies and corresponding mode shapes) of the designed prototype structure are analyzed [118]. The steady-state response of the TFM stator teeth to a sinusoidally (harmonically) varying load and the consequent TFM performance under resonance, and fatigue are also investigated.

7.4.2 Modal and structural harmonic analysis of baseline TFM

The natural frequencies and mode shapes are important parameters in the design of a structure for dynamic loading conditions. The modal analysis in ANSYS Mechanical was used to determine the critical resonant frequencies of the proposed TFM stator core with housing. The critical mode frequencies and the mode shapes are shown in Table. 7.3.
Table 7.3: Critical resonance frequency for the proposed baseline TFM.

<table>
<thead>
<tr>
<th>Natural frequency for stator with housing</th>
</tr>
</thead>
<tbody>
<tr>
<td>1845 Hz</td>
</tr>
</tbody>
</table>

When running the TFM at rated speed under load condition, it is quintessential that the harmonically varying load acting on the stator does not excite the critical natural frequencies. For the 30-pole machine designed during this research, the fundamental excitation frequency stands at 100 Hz, which is much lower than the critical frequency spectrum. However, there are harmonic contents of the electromagnetic force acting on the stator that will excite the natural frequency under certain loading conditions. In order to check the vibrational response of the TFM under resonance, Structural Harmonic Analysis is carried out in ANSYS Workbench using the block diagram shown in Fig. 7.7. 3D-FEA was used to generate the harmonic forces acting on the stator tooth, which was directly coupled to the mechanical domain, in ANSYS Workbench. The frequency response for the directional deformation and acceleration on the stator housing due to the harmonic load acting on the stator tooth was analyzed. The peak deformation and acceleration at the resonant frequency condition was recorded and shown in Fig. 7.8.
At rated speed and load condition, the designed TFM exhibits good vibrational performance under resonance, with the peak deformations being less than 0.08 micron, and the acceleration less than 10m/s². This shows that a sudden increase in vibration during resonance, will not significantly impact the structural integrity of the TFM assembly.
7.4.3 Modal and structural harmonic analysis of slant-pole TFM

For the optimized slant-pole shaped stator core proposed in Section 7.2, it is important to check how shaping the pole of the stator cores is affecting the vibrational performance of the TFM under resonant condition. The slant-pole shaping of the stator core causes a slight reduction in the stator pole lamination area and will affect the overall stiffness of the stator core. The Modal analysis was used to analyze and compare the critical resonant frequency for the Slant-pole TFM with the baseline TFM design as shown in Table. 7.4. The results confirm that slightly changing the stator geometry through pole shaping, does not lower the resonant frequency values towards the fundamental excitation frequency.

Table 7.4: Critical resonance frequency for the slant-pole and baseline TFM.

<table>
<thead>
<tr>
<th>Natural frequency for stator with housing</th>
</tr>
</thead>
<tbody>
<tr>
<td>Baseline</td>
</tr>
<tr>
<td>Slant-pole</td>
</tr>
</tbody>
</table>

At rated speed and rated load condition, the structural harmonic analysis was carried out with the slant-pole stator, and the comparative results for the axial deformation and acceleration at resonant frequencies with the baseline model are shown in Fig. 7.9.
Both graphs show an increase in deformation and acceleration for the slant-pole TFM compared to the baseline model. However, the deformation magnitude is only 0.14 micron and the acceleration is 19 m/s² for the slant-pole TFM at 1800 Hz. This indicates that while carrying out pole-shaping to tone down the torque ripple, the TFM will not suffer from any issues related to structural mode frequencies and resonant vibrations.

7.5 Conclusion

The issue of torque ripple in a 3-phase TFM was addressed in this Chapter. The torque ripple harmonics were successfully toned down using stator pole shaping. The vibrational analysis also showed the proposed TFM design to have good performance under resonant conditions.
8.1 Introduction

After performing comprehensive 3D-FEA and MEC based electromagnetic optimization and vibrational performance analyses, the Z-Core TFM was prototyped. The prototype TFM was used for experimental validation of the 3D-FEA and MEC model, and also acts a testament to the working principle of a double-sided Z-TFM with ferrite based FC rotor.

8.2 Structural Prototyping of the TFM

The TFMs have a double-stator, single-rotor structure with FC placement of ferrite magnets in the rotor. Ferrite magnets were embedded with the rotor cores, and FC techniques were used to amplify the air gap magnetic flux density. The rotor and stator cores were made of lamination steel stacked together as shown in Figs. 8.1 and 8.2. Individual lamination sheets of M270_35A at 29 gauge (0.35 mm) were cut according to the required dimensions and welded to the required width.
Figure 8.1: Rotor cores with stacked laminations.

Figure 8.2: Z-Core stator with stacked laminations.

Y30 grade sintered ferrite (ceramic) magnets were used in the rotor for flux concentration and leakage reduction. The magnets for the prototype are shown in Fig. 8.3. The rotor cores and the magnets are attached to a non-magnetic disk-shaped rotor housing. Aluminum 6061 plates were machined into a circular ring to hold the rotor cores and the magnets and to increase the rigidity of the rotor. Carbon steel 14L was used for the shaft. The machined rotor housing with the shaft is shown in Fig. 8.4. Radial ball bearings are to be used in this prototype.

Figure 8.3: Ferrite magnets used in the rotor.

Figure 8.4: Rotor housing with shaft and bearing.

The assembly of the rotor is shown in Fig. 8.5. A plastic fixture from Delrin material is developed to hold the rotor cores and radial flux-focusing magnets in place. Delrin does not react the epoxy. The rotor cores and the magnets are glued together using Permabond Acrylic Adhesive (2-part Epoxy). Once the rotor parts are cured, they are detached from the plastic fixture and mounted on a Lathe machine to rotate the rotor at rated and maximum speeds to check for the structural rigidity. The assembly of the stator is shown in Fig. 8.6.
The stator cores are attached to a supporting end-plate at the two ends. The end plates were also machined out from Aluminum 6061. The stator cores were connected to the endplates using Permabond Acrylic Adhesive. Once the stator cores have bonded with the end-plate, the cores were wrapped with insulation tape. Then the ring windings made using AWG 15 gauge magnet wire was inserted into the cores.

Figure 8.5: Rotor cores and magnets attached to the rotor housing.

Figure 8.6: Stator assembly with windings.

The final assembly of Z-TFM prototype is shown in Fig. 8.7. The double-sided stators are held in place using rods bolted to the end plate that allowed for varying the axial airgap on both sides of the rotor. The distance between the two stators in the final assembly could be adjusted through the adjustment nuts holding the prototype together as shown in Fig. 8.7. The axial location of the bearing could also be adjusted through the bearing adjustment nuts. Having the flexibility of changing the axial air gap length, allowing the TFM prototype to be tested at different operating points. It also aided in fine-tuning and validating the FEA models at different air gaps. Fig. 8.8, shows the test bench developed
for testing the prototype Z-TFM. A C-Channel is used as the supporting base. Three rectangular aluminum plates with three supporting threaded rods are used to form the frames for mounting the TFM and the Parker servo PMSM. Nuts and washers are used to ensure a tight fit between the supporting structures and the machines. Two couplings and a torque sensor were used to couple the TFM with the servo PMSM.

Figure 8.7: Final assembled TFM prototype.  
Figure 8.8: Test bench for experimental verification.

8.3 Discrepancies in the Prototype Z-TFM

During the prototype development, there were some discrepancies associated with the mechanical assembly and the lamination and magnet materials used. The lamination M270_35A (0.35 mm) that was received had a 0.34 mm lamination thickness which was used to create a pole width (WR=Ws) of 8.15 mm instead of 8.25 mm (used in 3D-FEA design) with a 90% stacking factor. The ferrite magnets used in the prototypes had lower remanence flux density (0.3 T instead of 0.4 T used in the 3D-FEA analysis). This will drive down the power factor of the machine as well as the torque density. The slot fill factor in the prototype was very poor since it became very difficult to create the ring winding inside the stator core after gluing them to the end-plate. The process would have been much simpler if the winding inside the stator were done before the cores were glued to the end-
Due to a low slot fill factor, the copper resistance in the experiment was higher which would negatively impact the efficiency of the TFM.

8.4 TFM Drive and Motor Controller used for Experimental Testing

A commercial drive was used for experimental verification of the TFM. The drive consists of two APS IAP300T (300A, 850V) inverters connected in a back to back configuration to attain the desired 6-Phase Bridge inverter. In each case, only 4 bridges will be used. The other two bridges would not be utilized. The drive can operate with switching frequencies of up to 10 kHz and support 600 A peak currents. An interface board for communication between the drive and the controller has been developed. The interface board provides the buffers and necessary voltage shifts needed for the PWM gate signals. The interface boards also condition the current for input to the ADC of the controller. A DSpace microautobox is to be used as the controller. The drive, interface board, and the DSpace controller are shown in Fig. 8.9.

Figure 8.9: TFM drive setup (a) drive for testing (b) interface board for testing (c) DSpace controller.

The Parker servomotor coupled to the Z-TFM will be used both as a motor and as a generator for no-load and load testing respectively. The motor controllers used for controlling the Parker Servo and the Z-TFM are shown in Figs. 8.10 and 8.11. A vector controller was used to regulate the current and a PI controller to regulate the speed of the Parker servo PMSM. It was used for spinning the Z-TFM at the unloaded condition to get
the cogging torque and back-EMF data. The Z-TFM was then tested as a motor with current regulation in the phases. In this case, the TFM was speed controlled and PMSM was connected to a programmable DC load via a passive rectifier to act as the load. A dual-band hysteresis current control was used in the controller as shown in Fig. 8.11.

Figure 8.10: Motor controller for driving the Parker Servo PMSM.

Figure 8.11: Motor controller for driving the Z-TFM.

8.5 Experimental Results

Since the single-phase Z-TFM is not self-starting and has a high cogging torque. Keeping an axial air gap length of 1 mm made it extremely difficult to manually provide a starting torque to the TFM during the load operation. Due to the presence of a high cogging torque in the TFM and limitations with the Parker servo PMSM to overcome that, it could not spin the TFM at the unloaded condition. As a result, an air gap length of 1.25 mm was selected to complete the experimental study for the Z-TFM for both unloaded and loaded conditions.

8.5.1 Unloaded condition: Z-TFM operating as a generator

The Parker servo PMSM was used to spin the rotor of the TFM at the rated speed of 400 rpm, and the peak cogging torque and the open circuit voltage (back-EMF) across the terminals of the TFM windings were measured. Since the prototype TFM has the added
feature of changing the air gap length, the back-EMF and cogging torque data were taken at different axial air gap lengths, the results are shown in Table 8.1. With a decreasing air gap, the effective magnetic flux linking with the stator winding increases due to a decrease in fringing and leakage flux. As the result the RMS back-EMF voltage increase with decreasing air gap. The increase in cogging torque is due to an increase in the amount of air gap flux crossing during pole overlap. The difference between 3D-FEA and experimentally measured values for RMS back-EMF voltage is around 20% at a 2.25 mm air gap. But the difference is gradually decreasing with the lower air gap, with only 6% difference at 1.25 mm. There are discrepancies in the 3D-FEA simulations, which do not take into account the low slot fill factor and rectangular magnet pieces with orthogonal magnetization, which are modeled as trapezoidal magnet pieces in the 3D-FEA simulation, due to the complication with meshing. Due to these modeling differences, at higher air gap where the amount of leakage flux is more significant in the prototype machine, it is not completely captured in 3D-FEA. The difference in estimating the cogging torque with 3D-FEA is due to the use of the coarse mesh setup in the 3D-FEA solver to save computation time. Nevertheless, the 3D-FEA simulation accurately predicts the trend for both the performance parameters due to decreasing air gap length.

Table 8.1: Back-EMF and cogging torque characteristics at different air gap length.

<table>
<thead>
<tr>
<th>Air Gap Length (mm)</th>
<th>RMS back-EMF voltage (V)</th>
<th>Peak Cogging Torque (Nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Experimental</td>
<td>3D-FEA</td>
</tr>
<tr>
<td>2.25</td>
<td>2.76</td>
<td>3.33</td>
</tr>
<tr>
<td>1.75</td>
<td>3.48</td>
<td>3.95</td>
</tr>
<tr>
<td>1.25</td>
<td>4.61</td>
<td>4.89</td>
</tr>
</tbody>
</table>

The back-EMF waveform at a rated speed obtained experimentally at an air gap of 1.25 mm is shown in Fig. 8.12. The experimental back-EMF waveform at rated speed was
also compared with 3D-FEA and the improved MEC model results as shown in Fig. 8.13. The MEC and 3D-FEA results show good correlation with the experimental result.

![Image](image1.png)

**Figure 8.12:** Experimental back-EMF waveform at 1.25mm air gap.

![Image](image2.png)

**Figure 8.13:** Experimental back-EMF waveform compared with 3D-FEA and MEC model results.

### 8.5.2 Loaded condition: Z-TFM operating as a motor

The Z-TFM was operated as a motor and the Parker servo PMSM was connected to a programmable DC load via a passive rectifier to act as the load. The Z-TFM, with an air gap length of 1.25 mm, was tested at different speeds of 400, 600, 800 and 1000rpm, where 400 rpm is the base speed and 1000 rpm is the maximum speed. Regulated phase currents in the TFM at a different speed and loading conditions are shown in Fig. 8.14. The results show good current regulation but at 1000 rpm, it could be further improved with better noise immunity.
The performance results for the prototype TFM at the aforementioned speed range for different RMS current loading were taken using a WT 3000 power analyzer. The results for output power factor and output power for different current loading at different speeds are given in Fig. 8.15 and 8.16 respectively. During the experiments, at low currents values, the presence of switching harmonics in the phase current is significant. With increased loading, the switching harmonics become less significant. The power factor and output power all show an increasing trend with increasing RMS current loading. The Z-TFM was successful to generate at peak power of 880 W at 1000 rpm, and also reached the power factor of more than 0.4 at the same speed. This is quite impressive for a machine that is built with low-grade ferrites and has a poor slot fill factor in the stator.
Figure 8.15: Power factor at different current loading.

Figure 8.16: Power output at different current loading.

The efficiency and the torque per ampere performance at different speed and current loading are shown in Figs. 8.17 and 8.18 respectively. As discussed in Section 8.3, the slot fill factor was low while winding the stator and resulted in higher copper resistance in the winding. This is abundantly visible in Fig. 8.17 where the efficiency tends to drop at higher loading conditions at different speeds. The torque per ampere also shows a similar trend as that of efficiency, which is a linearly increasing trend at lower current loading, but the gradually declining trend at higher current loading condition due to magnetic saturation and higher copper loss.

Figure 8.17: Efficiency at different current loading.

Figure 8.18: Torque per ampere at different current loading.

A comparison between the experimental and 3D-FEA results for output torque against current loading is shown in Fig. 8.19. The difference between the 3D-FEA and
Experimental results is due to the presence of switching harmonics which is significant in the phase current at low RMS values. However, the switching harmonics become less dominant at higher current loading, which is evident in Fig. 8.19, where the 3D-FEA and experimental values show good correlation at higher RMS current loading. The torque-speed for the TFM at a constant power output of around 270 watts is shown in Fig. 8.20.

![Graph of Average Torque at Different Current Loading](image1)

**Figure 8.19:** Average torque output at different current loading.

![Graph of Torque Speed Curve](image2)

**Figure 8.20:** Torque speed curve for the TFM at a constant power of 270W.

The vibration performance of the Z-TFM was also experimentally obtained using the setup shown in Fig. 8.21, while running the machine as a motor. The accelerometer mounted on the housing of the TFM measured the acceleration in the z-axis (axial direction) of the TFM.

![Test Setup for Measuring Vibration on Z-TFM](image3)

**Figure 8.21:** Test setup for measuring vibration on the Z-TFM.
The axial pressure acting on the stator tooth is propagated to the stator housing. The harmonic response for the peak acceleration on the stator for both the machine under different loading conditions is shown in Fig. 8.22. The accelerometer was mounted on the end plate of the stator housing, and the vibration along the axial direction (z-axis) was measured. The vibration of the Z-TFM was characterized at different speeds while varying the RMS current loading of the machine. As shown in Chapter VII, the natural frequency of the Z-TFM setup started from 1800 Hz. The fundamental excitation frequency for the 30-pole Z-TFM is 100Hz, 150Hz, 200Hz, and 250Hz at 400rpm, 600rpm, 800rpm and 1000rpm respectively. Since these fundamental frequencies are not exiting the natural frequency of the motor, the condition of resonance is avoided. At the rated speed of 400 rpm, the Z-TFM has a good vibration performance at low order frequencies. However, at higher speeds, the low order components of the axial harmonic forces become more dominant and cause an increase in vibration at lower frequencies but not severe enough to cause any structural damage. The experimental results verify the structural FEA results from Section 7.4, which also indicated that the vibration performance for the proposed TFM had low magnitude acceleration.
8.6 Conclusion

In this chapter, the mechanical development of the proposed Z-TFM was presented in detail. The Z-TFM prototype was tested, both as a motor and as a generator. The open circuit back-EMF voltage and cogging torque were measured and compared with 3D-FEA at different air gap levels. While running as a motor, the Z-TFM was tested at different speeds at an air gap of 1.25 mm. The prototype was developed using low-grade ferrite magnets and a low slot fill factor, which contributed to its low power factor and efficiency.

The experiments verify the feasibility and performance potential of the proposed machine. Improved prototyping techniques, better winding with a high slot fill factor and use of high-performance ferrite magnets would result in a better performance of the machine.
CHAPTER IX
SUMMARY AND FUTURE WORK

9.1 Introduction

This chapter recaps the contribution made through the research and prototype development of a novel double-sided Z-Core TFM with ferrite based FC rotor. The proposed TFM topology had never been built before and the advances made through its development during this dissertation are expected to have a positive impact on the future development of TFMs. Section 9.2 highlights the contribution of the research presented in this dissertation and Section 9.3 discusses the ideas that can be explored in the future for better TFM design and development.

9.2 Research Contribution

The research in this dissertation contributes toward the MEC modeling, design optimization, torque ripple minimization and vibration analysis of a novel Z-Core TFM. The important contributions can be summarized as:

- Complete characterization of TFM topologies, including their advantages and disadvantages.
- The proposed novel double-sided topology using Z-core TFM and ferrite based FC rotor held in a cantilever arrangement.
An improved Magnetic Equivalent Circuit model for faster TFM electromagnetic analysis.

MEC based PSO analytical tool for faster TFM optimization of critical design parameters.

A novel magnet-free TFM concept with an integrated rotary transformer, designed for a wide speed operating range.

Cogging torque and torque ripple reduction in the proposed TFM through slant-pole shaping in the stator.

Complete mechanical prototype and test-bench development at the Alternative Energy Laboratory at the University of Akron.

Experimental analysis for characterizing the baseline TFM prototype performance.

9.3 Future Work

There is always a scope for improving the existing research and investigating new topologies with a different take on the modeling, design and optimization approach. The developed Z-Core TFM holds the promise of delivering better performance with improved prototyping materials, manufacturing, and assembling techniques. The following research works could be carried out in the future for this TFM design:

- A prototype implementation for the magnet-free TFM with a rotary transformer.
- A 3-phase prototype implementation of the TFM with a slant-pole Z-core stator for validating torque ripple improvement.
- Investigate different slot-pole combinations for torque density and power factor improvement.
- Investigate the use of skewed magnets in the rotor for cogging torque reduction.
REFERENCES


