ABSTRACT

ADEEB, AHMED. Design and Optimization of a Novel Permanent Magnet Synchronous Machine Based on Transverse Flux Topology. (Under the direction of Dr. Iqbal Husain).

Rotating electric machines are the primary mean for world energy generation and a major candidate for energy consumption as well. A minute change in the performance of these machines inflicts a large impact in the world energy expenditure. Among different types of electric machines, permanent magnet machines are gaining attention due to their capability of high-efficiency operation and compact size. Moreover, recent electrification strategy in the automotive industry, increasing popularity of wind energy harvest and overall desirability of device compactness are creating higher demand for machines with smaller physical footprints.

Electric machines are primarily torque producing entity and the compactness of these machines is synonymous with their torque yield capability within a physical constraint or more commonly referred as torque density. When torque density is of concern, a special variant of electric machines- known as transverse flux machines (TFM) deserve special consideration. These machines are reputed for high torque density operation but have never been popularized due to some limiting factors such as low power factor and consequent increase in drive circuit cost, structural complexities and associated manufacturing cost. The primary target of this research is to perform a thorough investigation in identifying the potentials of this topology over conventional radial flux machines (RFM), recognizing the root sources responsible for causing the limitations and finally addressing the concerns with a systematic approach.

The outcome of the in-depth analysis provided the concept of a novel transverse flux topology with substantially improved performance over existing state of the art machines. The topology was built upon strong logical reasoning with the highest emphasis on reducing structural complexities and exclusion of unconventional construction materials otherwise common in existing TFMs. An innovative approach was undertaken to mathematically represent the factors primarily responsible for performance degradation and an objective-function based method was followed for design optimization. Moreover, mathematical analysis of torque ripple harmonic contributions was carried out and design optimization was performed to ensure reasonable performance. As subsidiary works, a mathematical model representing TFM performance was developed which can be an effective tool for thorough investigation of these machines.
Finally, a prototype was built which validated the topology benefit and the systematic approach undertaken during the design process. Detailed experiments were performed with the prototype and the performance proved to be in close correspondence with finite element analysis (FEA) results. The machine exhibited high torque density with power factor comparable with conventional permanent magnet machines. During the test procedure, a novel approach was undertaken for accurately separating the loss components and consequent loss comparison with FEA prediction.

In conclusion, the research work provided a novel TFM capable of exhibiting high torque density with good power factor performance. The topology relied on conventional motor construction materials such as silicon steel and structural complexity—especially for the moving part was reduced to a negligible extent. Subsequently, the primary concerns of this topology were effectively addressed and the proposed novel TFM brings these machines a step closer to large-scale industry adoption.
Design and Optimization of a Novel Permanent Magnet Synchronous Machine Based on Transverse Flux Topology

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DEDICATION

Words cannot do justice to describe her efforts and contributions. To my beloved mother Humaira Ahmed.
BIOGRAPHY

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Chapter 1: Transverse Flux Machines- A Closer Look

1.1 Transverse Flux Machine Background

Transverse flux machines are electric machines which rely on a magnetic flux path not confined in a radial plane. The concept of transverse flux paths in electric machines have been proposed back in 1895 by W.M. Morday[1] but never been popularized. The fundamental idea behind transverse flux topology lies upon the use of a simple ring-shaped winding which is one of the main attractions of TFMs. A number of stator pole pieces can be wrapped around this simple ring-shaped coil to guide the magnetic flux lines around the winding. This eventually makes the manufacturing of the stator structure rather complicated and can be attributed as one the major causes of TFMs being the unconventional topology despite offering some definite advantages.

Transverse flux machines are best known for their high torque density capability. Though it fell behind in competition due to structural complexities, the advent of new materials and improved manufacturing technologies brought attention in TFM research primarily after some works done by Weh and May in 1986 [2]. In recent years, several research groups have been involved with TFM related research focusing on new ideas and developments. Though transverse flux topology itself does not necessitate the use of permanent magnets, almost all researchers primarily targeted permanent magnet (PM) machines. Limited works have been done on transverse flux reluctance motors [3] and there have been significant works on transverse flux topology in linear actuator designs [4][5][6][7]. Among the permanent motor TFM designs, notable researchers worked on the very basic structure with simple ‘C’ shaped stator cores [8][9][10][11]. These designs are simple to construct but come at the expense of some sacrifice in torque density and power factor. Designs with added structural complexities have been recently developed to make better utilization of material and to improve performance[12][13][14][15][16][17]. Some of these structures have such 3D flux paths that makes it necessary to use isotropic material like Soft Magnetic Composite (SMC) which suffers from higher reluctance compared to laminated steel[12][13]. Even more ambitious designs have been developed with double-sided stator further increasing the torque density[18][19]. Primary objectives of most of these researchers have been to increase the torque producing capability of these machines. Though TFMs are primarily known as low-speed high-torque machines, analytically they are identical to regular radial flux machines and there is no such restriction in high-speed operation. Such attempts can be seen by few researchers[20], [21] where TFM was used rather as low-torque high-speed drive.
1.2 Basic Flux Path and Torque Production

There are different variants of TFMs each addressing certain issues. Despite dissimilarities in structure, the basic concept of creating the flux linkage and torque production is comparable for most TFMs. Essentially, principle of torque generation from the interaction of stator and rotor magnetomotive force (MMF) is similar for all electromagnetic machines. The only factors that create the difference between TFMs and conventional radial flux machines are based on the geometric planes where these flux lines traverse with respect to the rotational axis. The concept can be best realized from the simplest TFM structure shown in Figure 1-1. One-quarter of a single phase 16-pole TFM is shown in the figure. Magnets with opposite magnetization polarities are shown in red and blue. The ring winding goes through all the stator poles and is shown in orange. The direction of rotation of the rotor is around the z-axis. The expected flux paths through one stator pole are shown with the green arrows. The 2D flux paths along the cross-section of the stator pole at this aligned position is shown in Figure 1-2(a). Figure 1-2(b) shows the flux direction at a different rotor position when the stator poles are aligned with the red magnets at the top. Thus, rotation of the rotor will create flux linkage with alternate polarities. This constitutes a single phase permanent magnet motor. It should be noted that, in this simple TFM, half of the magnets are being unutilized at a certain time. Essentially, this is not the most preferred TFM structure and further details will be explained in later sections.

One-quarter of a similar 32-pole outer rotor single phase TFM is shown in Figure 1-3. Fundamental benefit of transverse flux can be realized by simple assessment of the 16-pole and 32-pole variants of the machine shown in Figure 1-1 and Figure 1-3 respectively. It is obvious from the two figures that both the machines have same air-gap area and total effective pole face area facing the air-gap region. In the 16-pole machine, the air-gap pole face region can be calculated by summing up the eight stator pole face areas facing the air-gap region. For the 32-pole variant, it is the sum of the sixteen stator pole frontal areas. Only, in this case, each pole is thinner in dimension essentially forming the same overall area. This will result in the total flux circulating around the winding to be similar for both machines. In other words, both machines will have similar flux linkage. Moreover, since adding number of poles does not change the shape of the 2D cross-section of these stator poles, same winding can be used for both structures. This enables injection of the same current MMF for both machines. Since the electromagnetic torque production depends on the spatial derivative of the linking flux and current MMF, having a larger spatial derivative gives the
larger pole machine the advantage of higher torque producing capability. In other words, for certain mechanical rotation of the rotor, the high pole machine will experience a larger time rate change of linking flux and based on Faraday’s law of induction, will generate a larger electromotive force (EMF).

![Figure 1-1: One-quarter of a 16 pole TFM.](image)

Since the number of poles can be increased indefinitely without making any change in the stator coil area and hence stator MMF, this feature gives the transverse flux topology its key advantage over the radial flux counterpart. Conclusively, the flux path in TFM eliminates the competition between magnetic loading and electric loading and this lays the foundation for a high torque machine. In an ideal situation, this enables the TFM to produce virtually infinite torque given that the number of poles can be increased indefinitely. However, in a practical scenario, increasing the number of poles creates higher leakage causing the benefit to diminish and that limits the torque producing capability to a finite quantity.
Three-phase machines in TFM topology can be constructed by simply stacking three single-phase machines axially with necessary phase shifts. The phase shift can be introduced by shifting the magnets in the rotor as shown in Figure 1-4(a) or by giving a similar shift in the stator poles as well. Another variant of the three-phase machine is shown in Figure 1-4(b) and an inner rotor structure is shown in this case. In this case, the stator poles are grouped into three separate clusters with necessary phase shifts and this limits the axial length of the motor. But it comes at the cost of a less efficient space utilization—evident from the required spacing between the phases. Moreover, it also results in a significant increase in the end-winding length and winding resistance.
The basic ‘C’-core variant of TFM is the simplest form the machine can have but suffers from lower torque density since 50% of the total magnets have no contribution at certain times. To make it even worse, these idle magnets are not truly idle and rather creates a negative flux linkage further reducing the flux linkage and therefore torque density. To address the issue, several design modifications have been proposed and new topologies have been introduced.
The simplest way to mitigate the issue of negative flux linkage is shown in Figure 1-5(a) where blocks of magnetically permeable bars—often referred as I-cores, are placed in between two ‘C’ shaped cores. This creates an alternative flux path and reduces the negative flux linkage created by these idle magnets. It also helps in reducing the flux pulsation inside the magnets and thus reducing the magnet Eddy current loss [19]. Nevertheless, this does not necessarily convert these flux into any useful quantity that can contribute in torque generation and doing so requires introduction to a more complex core structure. Such structure is shown in Figure 1-5(b) where the magnets are magnetized in the tangential direction to create flux concentration in the SMC material wedged between the magnets. Stator teeth are separated by a full electrical cycle to receive the inward flux from the alternate poles at the upper segment. The flux is carried back and driven towards the bottom layer where it completes the path through the stator teeth at the bottom segment which are 180 degrees electrically shifted from the top layer teeth. Therefore, all the magnets are contributing to creating the flux linkage and this significantly increases the torque density of these machines. While this structure ensures additive flux linkage from all the magnets, this is only achievable with the employment of flux concentration in the rotor. However, from the complex flux paths, it is evident that the material used in the stator core must be capable of carrying flux both in circumferential and axial directions. This requires usage of SMC materials in these machines adding some level of complexities. The $BH$ characteristics of these SMC materials are not as good as that of silicon steel and that diminishes the gain achieved from the complex structure[22]. Besides, not being the most conventional material used for the construction of electric machines, SMC increases the manufacturing complexity and often lacks the mechanical strength needed for a rotating machine structure[19]. Some alternatives have been proposed in some literatures where a hybrid structure is used to incorporate the 3D flux paths. In these machines, stator back iron is constructed using SMC while the tooth area are made with conventional silicon steel [22][15]. Silicon steel stacked in different directions and then combined to form the 3D flux paths have been proposed as well which eventually resembles a structure shown in Figure 1-5(b) [23]. The only difference between this variant with the one with SMC is in the construction of the stator core but both result in similar 3D flux paths.
Though there have been attempts to incorporate flux produced by all magnets into single coil using only laminations, the concept is suitable only for linear actuator [12]. It requires bending the lamination core to get a defined form that ensures desired flux paths to create additive flux linkage from all the magnets. However, this structure is not practical for a rotary machine. It is evident from the discussions so far that TFMs made with regular ‘C’ shaped cores and single ring coil fail to link the flux from all the magnets in an additive fashion. Doing so requires either SMC or some hybrid structure made from both lamination and SMC or at least lamination stacks of different stacking directions. If the simple conventional lamination-based structure is desired, it is, however, possible to use all the magnet flux and achieve a higher torque density by simply adding another coil to the structure. Such designs are proposed in [16][24][18] where the fundamental idea is to harvest the flux linkage created by the adjacent magnet pairs with two separate coils. Since the flux path will be in opposing direction for adjacent magnet pairs, a suitable connection of the two coils will be required to form a phase winding. The design shown in Figure 1-6 (a) exhibits such topology where the cores can be referred as modified ‘C’ core that allows passing of one coil through it while allowing the other coil to skip the flux associated with it.
Similar stator pole is flipped to capture the flux linkage from the adjacent magnet pairs making the full utilization of magnets and space. A similar concept is shown in Figure 1-6 (b) where flux linkage from magnets are captured with two coils placed at the alternative sides of the rotor giving it a double stator structure.

![Figure 1-6: Transverse flux machine with double coil structure](image)

Further variations can be introduced in the topologies described so far by introduction or omission of flux concentration in the rotor. Using flux concentration in the rotor can be a way to further increase the air-gap flux density and mostly effective when using low energy density magnets. However, in most cases, the incorporation of flux concentration in the rotor requires SMC material in the rotor to incorporate the bending of flux inside the rotor. Irrespective of the use of flux concentration, fundamentally TFMs can be categorized into three classes, namely-

i) ‘C’-shaped core with single coil and unutilized magnets

ii) Core incorporating 3D flux path with full magnet utilization and single coil (requires SMC or hybrid lamination and flux concentration at rotor)

iii) Modified ‘C’ core with double coil with full magnet utilization

There have been proposals of TMF based on flux switching topology where magnets are placed in the stator and suitable rotor assembly is used to direct the flux to alternate the flux linkage. Despite the variations in the means to alternate the flux direction, the fundamental flux path in these machines closely resembles the topology shown in Figure 1-5 (b)
1.4 Transverse Flux Topology- Advantages

As described earlier, number of poles in TFMs can be increased indefinitely without any change in the coil geometry and flux linkage and this eliminates the competition between electric and magnetic loading. Consequently, it is a common practice for most TFM based designs to have a high pole count and targeted for high-torque low-speed applications. Overall benefits of the transverse flux topology can be summarized as follows—

1) The absence of competition between electric and magnetic loading enables the design of high pole machine. This essentially results in a high torque throughput as well. High torque density motors are especially suitable for direct drive applications and can greatly increase system efficiency and reliability.

2) As seen from the structures, the phases are decoupled making the design of multiphase machines straightforward. One single phase structure can be used to create either a two-phase, three phase or poly-phase machine by stacking the stator with proper phase-shift. Another benefit of the decoupled phase is that it simplifies the control of these motors[25].

3) Decoupled phases also permit a modular design. In many applications modularity in the design can be an effective way for cost minimization.

4) The transverse flux path greatly simplifies the winding configuration. A simple ring-shaped coil is simple to wound and can yield high fill factor and thus reduce copper loss. A ring winding yields shorter end winding and consequently further reduces copper volume and copper loss.

1.5 Drawbacks and Limiting Factors

All the advantages offered by TFMs described so far seem to lead these motors as the best candidate for high torque applications. Unfortunately, if the non-idealities in magnetic flux paths are considered, the performance of TFMs deviates significantly from the anticipated course considered otherwise. The potential of linear increase in torque generation with the increasing number of poles merely becomes a delusion considering the non-idealities. Nevertheless, if the torque production capability is taken as the only measure of performance, TFMs can often defeat the radial flux topologies. Despite offering significant benefits over radial flux counterparts, there have been limited examples of commercialized TFMs. The shortcomings deterring the TFM adaptation in most applications can be attributed on several factors as follows—
1) Conceivably the leading reason that prevents TFM to take over the radial flux equivalent is that it suffers from high leakage flux and subsequently offers a low power factor. As cost associated with the machine drive circuitry increases significantly for a high volt-ampere requirement, having a low power factor often surpasses the gain accomplished by having high torque density. Moreover, as torque production increases with the increase in pole number, high pole count also reduces the tolerance in maintaining angular displacement in the mechanical structure. The effective stator slot and rotor pole number combinations in TFM also result in high cogging torque and torque ripple.

2) Unconventional structure often results in a significant increase in the manufacturing cost practically restricting the commercial viability. While modular design can be beneficial for some applications, it also requires a complicated mechanical structure to support the active materials.

3) The most common TMF structures require the use of SMC. SMC shows promising results in many applications but yet to defeat silicon steel, especially for rotary machines. Further improvement of SMC in terms of $BH$ characteristics and mechanical rigidity can help in turning the table in favor of TFM.

1.6 Research Goals

Transverse flux topology shows promising performance in terms of torque density and can be an attractive design option provided that the issues have been addressed. The primary objective of the thesis is to devise a novel design that addresses the issues while maintaining reasonable manufacturing complexity. The major objectives can be set as follows—

1) Achieve high torque density with a reasonable mechanical structure
2) Reduce leakage flux and hence improve power factor
3) Improve the machine performance in terms of cogging torque and torque ripple
4) Use conventional silicon steel and keep the manufacturing process as simple as possible.
2.1 Introduction

In this chapter, a novel transverse flux topology will be described using simple 3D and 2D diagrams. As discussed earlier, the primary objective in the development stage was to construct a high torque density TFM using conventional material and simple manufacturing process. The first structure presented here can be considered as the base structure to describe the design basic and gradual tuning of this structure will be done in the succeeding sections to address specific issues.

2.2 Topology Description

3D view of three poles of the proposed outer rotor TFM is shown in Figure 2-1(a) with the desired flux paths shown with green arrows. It is evident from the previously discussed TMF topologies that the simple ‘C’ core based TFM fails to capture flux linkage from all the magnets and the only way to do so using standard lamination core is to use two coils. The bottom coil or coil- in the figure is linking all the flux coming from the central magnet in Figure 2-1(a) and the 2D flux path is shown in Figure 2-1(b). Similarly, the top coil or coil-2 is linking flux coming out of the alternate poles and the 2D view is shown in Figure 2-1(c). This eliminates the need for any circumferential motion of flux and thus use of special material that allows 3D flux path. The flux path in the rotor is ideally axial but also possible in the circumferential direction. The rotor can be made of solid magnetic steel that significantly simplifies the construction.
Figure 2-1: (a) 3D view of three poles of the proposed topology with desired flux path shown with green arrows. (b) Flux path in a 2D plane for the stator poles aligned with red magnets at the top layer (c) Flux path in a 2D plane for the stator poles aligned with blue magnets at the top layer

2.3 Space utilization and overlapping concept

3D views of stator poles for two different cases of the described topology are shown in Figure 2-2 (a) and (b). Since steel laminations are stacked along the circumferential direction, these stator poles must have the same thickness at any radial location. As evident from the 3D view, the thickness of these poles is readily calculated by the number of poles and inner radius $R_{in}$ of the stator. Smaller $R_{in}$ results in a thinner stator pole thickness and this reduces the effective pole face
area at the air-gap. This causes inefficient utilization of space which becomes even more challenging with smaller diameter designs. Even with larger diameter machines, increasing $R_{in}$ to reduce this effect require keeping a large hollow space at the center. The issue is mitigated by creating a unique stator pole shape that allows sharing of angular space between the adjacent poles or more appropriately an overlap between the adjacent poles. To create such structure, a vertical gap is maintained at the inner section of the poles denoted by $L_z$ in Figure 2-1(b) and (c). This vertical space allows each pole to extend to the angular space designated for the adjacent poles. In this special case, the thickness of each stator pole can be calculated as

$$S_w = 2 \times R_{in} \sin(OSF \times \frac{\pi}{P})$$

where $P$ is the number of poles, $R_{in}$ is the inner radius of the stator and $OSF$ can be called the over span factor. This number is determining how much each pole extends to the neighboring region and impacts largely in the flux linkage profile.

![Figure 2-2: (a) 3D view of the stator poles with $OSF = 1$ (b) 3D view of the stator poles with $OSF = 1.5$](image)

There have been misleading concepts about ring winding of having no end turns which is not true particularly for lamination based TFM. As seen from Figure 2-2 (a) and (b), only a fraction of the coil is creating flux linkage from alternating poles and the rest of the segment is virtually not contributing for any constructive electromagnetic performance rather than creating copper loss and leakage inductances. This coil region is analogous to end-winding in radial flux machine and must be minimized. The over spanning of stator poles can be also credited as an effective mean of end
winding reduction. Figure 2-2 (a) illustrates the case where each stator pole is creating an angle of \( \frac{2\pi}{P} \) at the center and hence no over spanning. Stator pole with \( OSF = 1.5 \) is shown in Figure 2-2 (b) which gives around 50% increase in pole face area and a plays a direct role in the reduction in end-winding length. The orange dotted line in Figure 2-2 (b) demonstrate the extra area achievable with this unique feature.

2.4 Inner Vs Outer Rotor Design

Though inner or outer rotor structures are often required by certain applications, in cases where both are equally acceptable, a justification is needed to come up with the desired variant. It is evident from the previous discussion that the lamination-based stator pole pieces become thinner for inner rotor structure and the unique feature had to be incorporated to mitigate the issue. This simply questions the justification behind outer rotor structure for such topology. In this section, analytical and physical explanation will be provided to justify the design and an unbiased comparison based on FEA results will be provided.

As total flux linkage by the coil directly represents the torque production capability of a motor, an idealized model can be compared for both cases for initial comparison. The simplified diagram for the stator and rotor in a 2D plane is shown in Figure 2-3 for both outer and inner rotor variants. For both structures, stator pole axial length facing the magnet at air gap is denoted by \( L_{MZ} \). To ensure nearly equal flux density inside the core throughout the flux path, stator tooth width of \( W_T \) is assumed both at the pole face and at the back of the pole. For both cases, total rotor thickness is denoted by \( L_R \), where \( L_R \) comprises rotor back iron, magnet and air-gap length. Only difference in the stator pole design for the outer rotor and the inner rotor structure comes from the fact that the outer stator structure does not need any vertical gap at the outers section as adjacent poles are well separated along the circumferential direction at this end. This gives the inner rotor structure the advantage of larger coil area. Though the vertical gap \( L_Z \) reduces the available coil space, it also enables the stator to extend to the neighboring region making this a critical design parameter. Larger \( L_Z \) results in further reduction in coil area while providing additional over spanning capability. For the inner rotor design, however, the stator tooth design shown in Figure 2-3(b) has the horizontal section symmetric across the central line and this represents the maximum space utilization.
Assuming similar wire gauge for both designs, number of turns placed in the winding area for outer rotor structure can be simply calculated as

\[ N_O = \frac{K_f L_{CX} L_{CZO}}{A_W} \]  

here, \( A_W = \) wire cross sectional area, \( K_f = \) Fill factor. For both inner and outer rotor structure, \( L_{CX} = R_{out} - R_{in} - 2W_T - L_R \). So, number of turns for outer rotor structure can also be expressed as

\[ N_O = \frac{K_f L_{CZO}}{A_{coil}} (R_{out} - R_{in} - 2W_T - L_R) \]  

Similarly, number of turns for the inner rotor structure would be

\[ N_I = \frac{K_f L_{CZI}}{A_{coil}} (R_{out} - R_{in} - 2W_T - L_R) \]  

For the outer rotor design, thickness of each stator pole is given by Equation 2-1. Thus, the total area of the stator pole facing the air-gap magnet can be given by

\[ A_{StatorPoleO} = 2P L_{MZ} R_{in} \sin(OSF \times \frac{\pi}{P}) \]  

Effective magnet area facing the air-gap can be calculated as

\[ A_{magnetO} = 2L_{MZ} (R_{out} - L_R) \sin(PAC_M \times \frac{\pi}{P}) \]  

Here, \( PAC_M \) is the magnet pole arc coefficient. The effective air-gap area can be readily available by taking the average of these two areas.

Considering a constant flux density of \( B_M \) at air-gap, total flux entering the stator can be readily available from this area and the flux density. Incorporating the number of turns the flux linkage available for the outer rotor structure is thus

\[ \lambda_O = PB_M \frac{(K_f L_{MZ})}{A_{coil}} (R_{out} - R_{in} - 2W_T - L_R) L_{CZO} \left[ R_{in} \sin(OSF \times \frac{\pi}{P}) + (R_{out} - L_R) \sin(PAC_M \times \frac{\pi}{P}) \right] \]
Figure 2-3: Simple 2D dimension for stator pole and coil area and the axis of rotation (a) Outer rotor structure (b) Inner rotor structure
Stator pole thickness for the inner rotor structure is difficult in a sense that it needs careful tuning to maximize flux linkage as there is a possibility of large rotor leakage flux as shown in Figure 2-4. Since the poles are facing the air-gap region where the distance between the adjacent poles are shortest, the inter-stator distance near air-gap must be significantly larger than the effective air-gap length to create meaningful flux linkage. The ratio between the stator pole arc to the maximum physical limit of this arc at the region is often defined as pole-arc-coefficient and thus

\[ PAC_s = \frac{\text{Arc created by Stator Pole}}{2\pi(R_{in} + L_R)} \]  \hspace{1cm} 2-8

Thus, stator pole thickness can be given by

\[ S_{Wi} = 2(R_{in} + L_R) \sin(PAC_s \times \frac{\pi}{P}) \]  \hspace{1cm} 2-9

here, \( PAC_s \) is the pole arc coefficient for stator pole and must be less than unity to avoid flux leakage. Magnets facing the air-gap will have an area of

\[ A_{magnet} = 2L_{MZ}(R_{in} + L_R) \sin(PAC_M \times \frac{\pi}{P}) \]  \hspace{1cm} 2-10

Going through similar calculations, total flux linkage for inner rotor design can be expressed as

\[ \lambda_i = PB_M \frac{K_f L_{MZO}}{A_{coil}} (R_{out} - R_{in} - 2W_T - L_R) L_{CZI} (R_{in} + L_R) \left[ \sin \left( PAC_s \times \frac{\pi}{P} \right) \right. \\
\left. + (PAC_M \times \frac{\pi}{P}) \right] \]  \hspace{1cm} 2-11

Comparison of Equation 2-7 with 2-11 gives us the analytical interpretation of some well-known feature of overall TFM topology and can be summarized as follows—
i) The flux linkage term comprises of the term $\frac{\pi}{P}$ inside the Sine function which indicates thinner stator pole for larger pole number but also has the multiplication factor $P$ which essentially cancels out the effect keeping the flux linkage constant for any pole number. This is namely the unique feature that attracts most TFM researchers as the torque production capability of these machines increase almost linearly with pole number given that the ideal design is achieved.

ii) There are several factors controlling the difference between outer and inner rotor flux linkage. First, the effective coil area which is larger for inner rotor design. The second contributor is the stator pole thickness and that depends on $PAC_s$ for inner rotor structure and $OSF$ for the outer rotor one. The third factor is coming from the fact that the outer rotor structure has a larger air-gap area and hence can have larger magnet MMF due to space availability in larger air-gap area.

From the main leakage path in inner rotor design, it is obvious that $PAC_s$ in these structure plays a critical role and must be optimized through 3D finite element analysis (FEA). To complete the comparison, both structures where simulated in Ansys Electronic Desktop for with varying $PAC_s$ and $OSF$. The comparison processes were repeated for both 16-pole and 32-pole designs and for two different magnet thicknesses. Major geometric dimensions and parameters for the analysis are given in Table 2-1.

Results of the FEA are shown in Figure 2-5(a) and (b) for magnet thickness of 2.5mm and 3.5mm respectively. Flux linkages for the outer rotor design are plotted in blue line whereas that for the inner rotor variant are plotted in red. For the inner rotor structure, $PAC_s$ maximizing the flux linkage can be considered the optimum solution and for 16-pole design, maximum flux linkage is achieved near $PAC_s = 0.8$ resulting in a flux linkage of 0.21 Wb. For outer rotor design, the flux linkage is lower than the inner rotor structure without the over spanning but even with $OSF = 1.15$, it surpasses maximum flux achievable from inner rotor structure. However, for this structure, the limit of over spanning is far more and maximum flux linkage of 0.26 can be achieved with $OSF = 1.45$. The difference is more pronounced for the 32-pole design. As evident from the analytical expression, the flux linkage should not change with the change in pole number as the effective pole face area remain constant, this is clearly visible for the outer rotor structure as both 16 and 32-pole designs are giving near identical flux linkage. This however fails to hold for the
inner rotor structure because of large leakage flux which undoubtedly the foremost drawback for TFM s. As a matter of fact, close correspondence to the flux linkage performance can be considered as a design closer to the ideal situation where leakage flux path does not exist.

Table 2-1: Simulation Parameters for Outer Vs Inner Rotor Topology comparison

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Outer Rotor</th>
<th>Inner Rotor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Remnant flux density $B_r$ (T)</td>
<td>1.28</td>
<td>1.28</td>
</tr>
<tr>
<td>Coil cross-sectional Area $A_w$ (mm$^2$)</td>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td>Fill Factor</td>
<td>0.6</td>
<td>0.6</td>
</tr>
<tr>
<td>Air-gap length (mm)</td>
<td>1.2</td>
<td>1.2</td>
</tr>
<tr>
<td>Structure inner radius $R_{in}$ (mm)</td>
<td>25</td>
<td>25</td>
</tr>
<tr>
<td>Structure outer radius $R_{out}$ (mm)</td>
<td>67.5</td>
<td>67.5</td>
</tr>
<tr>
<td>Rotor back iron thickness (mm)</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Magnet pole arc coefficient</td>
<td>0.8</td>
<td>0.8</td>
</tr>
<tr>
<td>Single pole axial length (mm)</td>
<td>35</td>
<td>35</td>
</tr>
<tr>
<td>Tooth width $W_T$ (mm)</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>Magnet axial length $L_{MZ}$ (mm)</td>
<td>14</td>
<td>14</td>
</tr>
<tr>
<td>Magnet thickness (mm)</td>
<td>[2.5 3.5]</td>
<td>[2.5 3.5]</td>
</tr>
<tr>
<td>Number of Poles $P$</td>
<td>[16 32]</td>
<td>[16 32]</td>
</tr>
<tr>
<td>Stator pole end vertical gap $L_Z$ (mm)</td>
<td>3</td>
<td>NA</td>
</tr>
<tr>
<td>Over span factor OSF</td>
<td>1-1.45</td>
<td>NA</td>
</tr>
<tr>
<td>Stator pole arc coefficient $PAC_s$</td>
<td>NA</td>
<td>0.55-0.99</td>
</tr>
</tbody>
</table>
Results for a magnet thickness of 3.5 mm are shown in Figure 2-5(b) where the performance of the inner rotor design is found to improve slightly. But still with proper over spanning, outer rotor structure always surpasses the inner rotor structure and most importantly yields the same results for both 16 and 32-pole design. The better performance in outer rotor structure is also caused by the extra amount of magnet that can be placed at the larger diameter rotor when the outer rotor topology is used.
Another benefit the outer rotor structure provided was overlooked from the flux linkage number but can contribute greatly to motor performance. As evident from the expression, larger coil area was used for the inner rotor topology and since the same wire gauge was considered for the comparison, the inner rotor resulted in a larger number of turns. This evidently significantly increases the coil resistance and copper loss. If a larger wire gauge is considered to keep an equal number of turns, the flux linkage numbers shown in Figure 2-5 will go more in favor of outer rotor topology but eventually larger allowable current for lower gauge wire will result in similar results as shown in Figure 2-5. In the diagram used in the simulation, inner rotor structure has approximately 60% larger copper area and that reflected in the number of turns as well. Another benefit the outer rotor structure provides is ubiquitous for any topology using ring winding. As winding resistance is directly proportional to the total copper length, a ring with a smaller radius is definitely a desirable feature and can be easily obtained from an inner stator design. From Figure 2-3, the average radius of the coil can be calculated and given by

\[ R_{\text{coilOuter}} = \frac{R_{\text{in}} + R_{\text{out}} - L_R}{2} \]

\[ R_{\text{coilInner}} = \frac{R_{\text{in}} + R_{\text{out}} + L_R}{2} \]

where \( R_{\text{coilOuter}} \) and \( R_{\text{coilInner}} \) are the radius for the coil for outer and inner rotor structures respectively. Both higher number of turns and larger coil resistance will eventually result in a much larger coil resistance for the inner rotor topology.

### 2.5 Effect of non-idealities

Though the outer rotor TFM seems to perform well compared to the inner rotor counterpart, it is not resistant to non-idealities like leakage flux and core saturation. The visualize the major leakage flux and the saturation condition inside the iron, two planes were chosen to observe flux density profile for the aligned position of the magnets.
Figure 2-6: (a) Simulation model of the outer rotor TFM base model with two 2D planes to observe leakage flux (b) Flux density plane along the red plane (c) Flux density plot along the blue plane (d) Flux vector for the blue plane showing the direction of leakage flux.

The red plane shown in Figure 2-6 (a) is placed through the center of the stator core and there should be a high flux density is expected through the ideal flux path inside the stator. The blue plane is placed in the middle of adjacent poles and in an ideal case, this plane should not contain any flux at all. Results for flux density and vector profile at no-load condition are given in Figure 2-6 (b) through Figure 2-6 (d). From the flux density plot along the central red plane shown in Figure 2-6 (b), it is obvious that there are some leakage flux in the air along the axial direction. Another fact that is not considered in an ideal motor model is the reluctance of the stator core due to saturation. The presence of core saturation and the associated reluctance drop is also obvious from the flux density. More concerning leakage flux can be observed in the blue plane from observing Figure 2-6 (c) and (d). There is a significant magnitude of leakage flux taking place in the circumferential direction and this results in a significant drop in the flux linkage. Thus, the ideal analytical model used in the previous section merely represents the practical system and an elaborate model considering all these non-idealities should be developed.
2.6 Analytical Model based on Non-linearity

The objective of a detailed analytical model of the topology is two-fold

i) As a 3D finite element modeling is time-consuming, an analytical model often comes handy in determining the initial sizing of different parameters. FEA based model can be used for further fine-tuning of the initial base design.

ii) A functioning model helps in gaining more insight of the system which eventually helps in fine tune through FEA. It is a good practice to learn the contribution of each factor rather than a brute force method to optimize through FEA.

An analytical approach with lumped parameter model is a useful approach in determining the initial motor parameters. The most convenient way is to use the magnetic circuit approach. The equivalent circuit model of the proposed topology is given in Figure 2-7. The reluctances along the desired flux paths are shown in green while the leakage path reluctances are shown in red. Reluctances in different flux paths are calculated assuming idealized flux paths. The following reluctances are assumed:

\[ R_{Magnet} = \text{Permanent magnet reluctance} \]
\[ R_{AG} = \text{Air gap reluctance} \]
\[ R_{Stator} = \text{Stator reluctance} \]
\[ R_{Rotor} = \text{Rotor reluctance} \]
\[ R_{LA} = \text{Axial leakage reluctance} \]
\[ R_{LC} = \text{Circumferential leakage reluctance} \]

The voltage sources in the magnetic circuit can be replaced by MMF magnitude

\[ F_m = H_C L_{magnet} \]

where \( H_C = \text{Coercivity of permanent magnet} \)

\( L_{magnet} = \text{Magnet thickness} \)
Figure 2-7: Lumped parameter model of the flux path in TFM

A similar approach is presented in [16] in a simplified manner where reluctance associated with laminated steel have been neglected in the solution process. The flux that effectively contributes in building the flux linkage is denoted by \( \phi_\delta \) and only this part circulates the full stator core. After going through KVL, KCL and some algebraic manipulations, \( \phi_\delta \) can be given by

\[
\phi_M = \frac{2F_m}{2R_{Magnet} + R_{Rotor} + (2R_{AG} + R_{Stator})R_p}
\]

where \( R_p = 1 + R_{Rotor}/(R_{AL}||R_{CL}) + \frac{2R_{Magnet}}{R_{AL}+R_{CL}} \)

Flux linkage amplitude can be readily computed as

\[
\lambda = 2PN\phi_M
\]

where \( N \) = Number of turns. Anticipated flux path with zero leakage is shown using green dotted lines in Figure 2-7. Flux path along this desired path is well defined and reluctances associated with this path can be computed considering idealized flux path. Flux lines along the leakage paths on the other hand exhibit more uncertainty and reasonable approximation should be adopted. Some basic diagram exhibiting assumed flux paths are given in Figure 2-8. Since the magnet and the stator pole width can vary, a trapezoidal air-gap cross section can be assumed for simple analysis as shown in Figure 2-8(a). Effective air-gap reluctance can be computed as
$$\mathbb{R}_{AG} = \frac{L_{AG}}{\mu_0 L_{MW} + \frac{S_W}{2}}$$

Figure 2-8: Detailed sketch for air-gap reluctance modeling (b) Leakage flux path along axial direction (c) Leakage flux along the circumferential direction.

Reluctance associated with the leakage flux paths are more complex and comprises of both air and steel. Some tuning factor is needed to account for these uncertainties. Optimization of these tuning factors based of few FEA results can ensure good correspondence over a wide range of design variables. Thus

$$\mathbb{R}_{LA} = K_{LA} \frac{L_{LA}}{\mu_0 A_{LA}}$$

$$\mathbb{R}_{LC} = K_{LC} \frac{L_{LC}}{\mu_0 A_{LC}}$$

where $L_{LA}$ and $L_{LC}$ are the effective length of the associated leakage flux path and $A_{LA}$ and $A_{LC}$ are the corresponding cross-sectional area. $K_{LA}$ and $K_{LC}$ are two tuning factors needed to be optimized to account for the uncertainty in the leakage flux.

Along the desired flux path, laminated steel holds a non-linear $BH$ characteristics as shown in Figure 2-9 making it difficult to calculate the reluctance. Due to the higher reluctance of air and magnet compared to laminated steel, steel reluctance is often neglected in initial calculations. But high torque density machines are often pushed close to the limit reaching the saturated zone of the $BH$ profile making it necessary to include the steel reluctance in the magnetic circuit solution. To assess the modeling performance, FEA based simulation for TFM was carried out with varying
different geometric parameters. The stator pole shape depends on in a wide variety of parameters but some dominating contributors such as $OSF$, number of poles, tooth width $W_T$ and stator inner radius $R_{in}$ was varied. A typical example of the effect of saturation can be realized by comparing the FEA and analytical flux linkage shown in Figure 2-10. In this analytical design, MMF drop in stator steel was ignored resulting in a substantial overestimation in flux linkage. This large estimation error necessitates the inclusion of the non-linearity in the analytical expression. In this section, an analytical approach will be presented to incorporate material non-linearity in the solution process to get more accurate flux linkage characteristics.

Figure 2-9: (a) B-H characteristics of M-19 steel (b) M-19 steel permeability as a function of flux density.

As electric steel carries more flux, it reaches saturation and hinders additional flux. This justifies plotting the permeability of the material as a function of flux density as presented in Figure 2-9(b).
The shape of this permeability profile can be approximated by a quadratic approximation as shown with a red dotted line in the figure. Thus, electric steel permeability can be approximated as

\[
\mu_{steel} = \sum_{n=0}^{2} P_n \left( \frac{\phi_M}{A_{steel}} \right)^n
\]

where \( P_n \) = Quadratic approximation coefficients, \( A_{steel} \) = Average cross-sectional area in steel flux path. By going through algebraic simplifications in the circuit model, stator flux can be computed by solving the following cubic equation

\[
\sum_{n=0}^{3} K_n (\phi_M)^n = 0
\]

where \( K_0 = -P_0 F_m A_{steel} \), \( K_1 = P_0 R_{ALL} A_{steel} - P_1 F_m + L_{steel} \), \( K_2 = P_1 R_{ALL} - \frac{P_2 F_m}{A_{steel}} \), \( K_3 = \frac{P_2 R_{ALL}}{A_{steel}} \), \( R_{ALL} = R_{AG} + R_{Magnet} + \frac{2R_{AG} R_{Magnet}}{R_{LA} + R_{LC}} \), \( L_{steel} \) = Effective length of flux path in steel.

Figure 2-10: Comparison between FEA and analytical flux linkage without considering reluctance drop in the stator core.

Flux linkage variation as a function of \( R_{in} \) for varying stator control parameters are shown in Figure 2-11. Figure 2-11(a) shows flux linkage variation in a 24-pole machine as a function of \( R_{in} \) for two different OSF values. For both cases, the analytical solution exhibit close correspondence with the FEA results proving the robustness of the model. Flux linkage variation for a 32-pole machine are plotted in Figure 2-11(b) this time for different tooth width \( W_T \). As flux
linkage is the product of flux entering the stator poles and total number of turns, a thinner tooth can result in a larger flux linkage as it increases the coil slot area.

Another interesting fact evident from the observation of Figure 2-11 is that there exists an optimum $R_{in}$ that maximizes flux linkage. For Figure 2-11(a), the maxima can be found at the same inner radius whereas for Figure 2-11(b), slightly larger $R_{in}$ maximized flux linkage for $W_T = 5 \text{ mm}$.
compared to $W_T = 6\, mm$. This relationship can be readily derived from the analytical expression of the flux linkage for outer rotor structure given in 2-7. Since lower $R_{in}$ results in smaller pole face area but a larger number of turns, by taking the derivative of flux linkage expression with respect to $R_{in}$, the optimum $R_{in}$ can be derived. Thus

$$R_{inOpt} = \frac{R_{out} - 2W_T - L_R}{2}$$

2-21

This explains a smaller $R_{inOpt}$ for 6mm tooth thickness. Putting the geometric dimensions in 2-21, optimum $R_{inOpt}$ for 5 and 6 mm tooth thickness are found to be 26 mm and 25 mm respectively which perfectly matches with the FEA and non-linear analytical expression of flux linkage. For TFM of such shape, the inner radius of the stator can be considered as one of the most dominating design component if volume torque density is to be maximized. Since there are some space requirement for the inner shaft and wiring to pass through, too small of an $R_{in}$ is not feasible. However, in the prototype motor design, the inner radius was set fixed as 25 mm and in next chapters, the detailed optimization procedure will be explained.
Chapter 3: Design Optimization-Torque Maximization

3.1 Objectives

Though 3D FEA is a strong tool in predicting motor performance, it cannot substitute the necessity of a physical prototype to prove that can be used for the ultimate comparison. Moreover, as the structural complexities of these machines are to blame for the unpopularity of TFM topology, construction of a proof of principle machine can give some in site about the complexities associated with the proposed machine.

As the topology is well suited for high torque production, primarily TFMs are considered for high-torque low-speed systems which often includes direct drive applications. Thus, the prototype is also targeted for a high torque machine. To make the testing and prototyping cost reasonable, a reasonably sized motor was designed which can be easily tested with the available test bench. The design of the machine will be primarily optimized considering high energy density rare earth magnets. However, in the final prototype, testing will be performed using two different rotors – one constructed using rare –earth magnet and the second one using non-rare earth based Ferrite magnets. Based on the external geometric limit, step by step design procedure for the prototype will be described in the following sections. Though the primary objective of the process is set as achieving a high torque density motor, reasonable performance in other factors such as efficiency, torque ripple, power factor will be addressed in the process.

Table 3.1: Design constraints

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<table>
<thead>
<tr>
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<tbody>
<tr>
<td>Outer Radius (R_{out})</td>
<td>67.5 mm</td>
<td>Current Density</td>
<td>6 (A/mm^2)</td>
</tr>
<tr>
<td>Axial Length (A_L)</td>
<td>125 mm</td>
<td>Air Gap Length</td>
<td>1.5 mm</td>
</tr>
<tr>
<td>Cooling system</td>
<td>Air cooled</td>
<td>Magnet Type</td>
<td>NdFeB, Ferrite</td>
</tr>
<tr>
<td>Number of phases</td>
<td>3</td>
<td>Maximum Speed</td>
<td>600 RPM</td>
</tr>
<tr>
<td>DC Bus Voltage</td>
<td>120 V</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

3.2 Effect of Number of Poles

As seen from the analytical expressions and some FEA presented in Chapter 2, TFMs are supposed to produce indefinitely large amount of torque with the increasing number of poles given that the non-idealities are absent. Though it is not possible to completely remove leakage flux in
a practical machine, a well-designed TFM can exhibit the linear relationship between the number of poles and the output torque until reaching a point where leakage flux overwhelms the incremental benefit achieved from the additional pole. Moreover, with increasing the number of poles, construction of the machine gets more complicated as larger pole count reduces the tolerance in the mechanical construction. The control of the machines also requires a high precision position sensing as any error in mechanical position sensing is scaled by the number of pole pairs. Therefore, deciding on the number of poles in these machines requires careful consideration.

3.2.1 Pole Pair Contribution for NdFeB Rotor

Since the effect of number of poles is realized by observing the leakage flux, a 3D finite element model is the only reliable mean for performance evaluation in such case. As the 2D flux path in stator pole and winding area does not change with changing number of poles, a straightforward comparison is possible for TFM topology. Results of such analysis with NdFeB magnets in the rotor are presented in Figure 3-1. Figure 3-1(a) shows the relationship between number of poles and mean output torque for two different current excitation levels. As stated earlier, the non-idealities in the machine are inevitable and the effect is well noticeable in the mean torque profile. The dotted lines in the figures represent the expected torque in case the ideal machine could have been designed with zero flux leakage. The effect of current excitation can be seen to have some impact in this non-linearity in torque vs pole profile. This can be comprehended by observing the difference between the actual torque and the ideal line for two different excitation levels.

The decision involving the number of poles gets more complicated if the terminal voltage in loaded condition shown in Figure 3-1(b) is considered. As torque production loses the linear relationship with increasing pole number, the terminal voltage does not necessarily lose that linearity making the larger pole more undesirable. Moreover, at higher excitation level, the voltage seems to be larger than the expected voltage rise calculated from pole number. Given a standard inverter with a fixed voltage and current levels, the phase voltage in loaded condition determines the corner speed and maximum power and hence power factor of these machines. The corresponding corner speeds of the machine assuming 120V DC bus voltage are shown in Figure 3-1(c) which maintains the near ideal inverse relationship with pole number.

As seen from the dotted lines, the terminal voltage in ideal condition would have the linear relationship with pole pair and that alone is acceptable and should not affect in maximum power and power factor given that the torque had maintained the linear relationship. The total output
power for the scenario is shown in Figure 3-1(d) which shows a larger drop in output power despite increasing torque. The drop in available output power can be attributed to the following causes

i) With increasing number of poles, torque output fails to increase linearly

ii) The terminal voltage, on the other hand, keeps increasing with a nearly linear relation.

   For larger current excitation it is even larger than the anticipated voltage ideally would have found.

This makes the choice of pole count a critical factor in the design. If maximizing output power is the primary objective, the 16-pole design capable of 325W output power seems to be the best choice but this will yield only 11 Nm torque with 4A current excitation. The power factor for this machine will be 0.74 at this excitation level as seen from Figure 3-2(a). The 40-pole design, on the other hand, produces 25.1 Nm torque with the same excitation which is more than 80% larger than the 20-pole design. But this will come at a price with a large penalty in power factor and hence maximum power. With the 40-pole design, the output power with similar inverter constraint is 213W with a power factor of 0.46 at the same excitation level. The 40-pole design also demands a more sophisticated construction due to reduced tolerance. Beyond the 40-pole mark, the gain in torque diminishes and thus going for that high pole number seems unreasonable. If a high torque density machine is required, the low pole machine with higher power factor does not seem to be a reasonable choice. Moreover, if such low torque performance is acceptable, the selection of transverse flux topology becomes questionable as the most attractive feature transverse flux topology capability of providing high torque density.

Since the primary objective is to maximize the torque output, 32-pole design can be considered a reasonable choice which gives 20.5 Nm torque at 4A excitation which is 88% larger than the 20-pole design torque. Output power at this current excitation was found to be 258W with a power factor of 0.53 both calculated with same inverter constraint.

The effect of this nearly linear voltage can be explained if the flux linkage of the machine at different conditions are observed. In Figure 3-2(b), flux linkages are plotted for no load, 2A peak current, and 4A peak current conditions. As stated earlier, the no-load flux linkages in TFM are supposed to be constant for any pole number, the no-load flux in Figure 3-2(b) is seen to have a small deviation from the ideal case. However, close correspondence of the no-load flux with the ideal condition is the result of lower rotor leakage achieved by the outer rotor structure.
Figure 3-1: Effect of number of pole variation with two different current excitation levels for NdFeB magnet based rotor. (a) Mean torque (b) Phase voltage at 100 RPM (c) Corner speed considering 120V DC bus voltage (d) Output power at corner speed.

Though there is a small drop in no-load flux with increasing pole number, this small drop alone does not explain the drastic deviation in mean torque from the linear line shown in Figure 3-1(a). The adverse effect of larger pole number can be better realized by observing the flux linkage with current excitation. Despite a small drop in no-load flux with an increase in pole number, the flux in the loaded condition with 4A peak current excitation seems to be increasing with higher pole count indicating larger armature reaction due to stator leakage. Thus, the stator leakage is increasing with increasing number of poles and this large contribution of armature reaction flux is mostly responsible for the larger voltage drop for high pole machine. Moreover, the larger circulating flux in the stator is also responsible for core saturation preventing the machine from effective torque output despite having a reasonable flux linkage and no load back EMF.
3.2.1 Pole Pair Contribution for Ferrite Rotor

To further analyze the effect, the same stator structure was simulated for a non-rare earth based TFM. Keeping everything else identical also allows isolating the effect of magnet MMF strength for motor performance.

Corresponding FEA results for pole number variation are shown in Figure 3-3 and Figure 3-4. A visual comparison between the plots presented in Figure 3-3 with that shown in Figure 3-1 initially shows no definite change in the pattern. The only major difference is in the torque, voltage and corresponding corner speed and power magnitude which is the direct outcome of the lower magnet MMF contribution. As expected, the weaker non-rare earth magnet can produce much less torque compared to the stronger NdFeB counterpart. This also causes these motors to exhibit lower power factor as evident from Figure 3-4 (a). The relationship between these performance numbers and number of poles is however similar if visual inspection is performed. A closer look into the RMS voltage pattern for these two types of machines seems to have some difference where Ferrite based structure is seen to have a steeper exponential rise compared to the NdFeB based motor. This sharper voltage rise can be explained if the loaded flux linkage profile shown in Figure 3-4(b) is compared with the same profile for NdFeB based motor shown in Figure 3-2(b). Though both motors exhibit near constant no-load flux linkage, the Ferrite magnet based motor flux at 4A current excitation seems to increase with larger number of pole while that for NdFeB magnet shows a flat profile. This flat profile for NdFeB based design can be attributed to the saturation in

Figure 3-2: (a) Effect of number of pole variation on power factor with two different current excitation levels for NdFeB based motor (b) Flux linkage at no load and 2A and 4A peak current excitation for different pole numbers.
the stator core due to high stator leakage. Once the flux density reaches the saturation limit of the silicon steel, the flux linkage profile loses the sinusoidal shape and results in a trapezoidal profile. Detail analysis of this stator flux linkage in motor performance will be explained in next section.

![Graphs showing the effect of number of pole variation with two different current excitation levels for Ferrite magnet based rotor.](image)

Figure 3-3: Effect of number of pole variation with two different current excitation levels for Ferrite magnet based rotor. (a) Mean torque (b) Phase voltage at 100 RPM (c) Corner speed considering 120V DC bus voltage (d) Output power at corner speed.

As most of the performance degradation has been attributed over increasing leakage with an increase in pole number, something unusual can be observed if the no-load flux linkage for low pole count is observed in Figure 3-2 (b) and Figure 3-4(b) both for NdFeB and Ferrite based designs. At this very low pole count, the no-load flux can be seen increasing until reaching the peak value both for NdFeB and Ferrite rotors, but the effect is more pronounced for the stronger magnet. This can be explained if the total flux path in the machine is considered which include the rotor back iron. In this topology, the flux in the rotor back iron can travel axially keeping the flux path confined in the same plane as the stator poles. But there is an alternative way in the
circumferential direction just like a radial flux machine. The effective reluctance of the rotor back iron should be calculated considering these parallel paths and this path is not essentially independent of pole number. The topology explained here can be considered more of a hybrid structure with a transverse flux stator and a pseudo-radial flux rotor. Similar to the radial flux designs, lower pole number increases the flux path length in the back iron requiring a thicker back iron and vice versa. Since the rotor back iron was kept identical in the analysis, the very low pole number resulted in a drop in performance due to core saturation in the rotor which accentuated for the stronger magnet.

![Graphs showing](image)

Figure 3-4: (a) Effect of number of pole variation on power factor with two different current excitation levels for Ferrite based motor (b) RMS flux linkage at no load and 2A and 4A peak current excitation for different pole numbers.

### 3.2.2 Stator MMF and Power Factor

From previous results, it is obvious that the exponential rise in voltage with increasing pole number resulted in a smaller corner speed and hence output power, the cause and effect can be better realized if the phenomenon is analyzed using motor power factor. In this section, a simple description of power factor and how different magnetic phenomenon adversely affect the power factor will be explained.
Figure 3-5: Phasor diagram showing the effect of inductance and current excitation in permanent magnet motors (a) Small inductance and small current (b) Small inductance and large current excitation (c) Large inductance and small excitation.

In Figure 3-5, the typical phase diagram of a permanent magnet motor is shown for three different cases. The green lines in the diagrams represent flux linkage while the blue lines are used for voltage phasors. The black arrow is used to show the stator excitation phasor. In a surface permanent magnet motor, the absence of saliency suggests the optimum current to be in phase with the back EMF $E_m$ and such current phasors are maintained for all three diagrams. As seen in the figure, the back EMF is the result of magnet flux linkage $\lambda_m$ and is constant for a constant speed. The stator flux represented by $\lambda_s$ is the direct outcome of stator current and depends on stator leakage and given by

$$\lambda_s = L_s I_s \quad 3-1$$

where $L_s$ is the motor inductance. For constant speed operation and sinusoidal operation, the terminal voltage can be expressed as

$$V_T = \omega \lambda_m + j \omega L_s I_s \quad 3-2$$

Since power factor is defined as the cosine of the angle between voltage and current, the operating power factor can be simply calculated from the phasor diagram and given by

$$PF = \frac{\lambda_m}{\sqrt{\lambda_m^2 + L_s^2 I_s^2}} \quad 3-3$$

Thus, the main contributing factors for power factor can be summarized as

i) Larger magnet flux linkage $\lambda_m$ will result in larger power factor. This can be seen by the difference in power factor between NdFeB and Ferrite based design. Reduction of rotor leakage also increases this flux linkage helping in improvement in power factor.
ii) Larger stator excitation will lower the power factor. This is also an inevitable case and must be accepted.

iii) Larger stator inductance will cause lower power factor. This stator inductance is basically the vital contributor and depends on the design of the motor.

The basic difference in motor performance in terms of power factor can be attributed to this stator inductance and this can be the deciding factor in choosing a motor topology in many applications. It should be noted that, even with the idealized flux path with zero leakage results in non-zero stator inductance and the associated inductance often denoted as $L_{ms}$ or the linking inductance is a completely acceptable feature in a motor and cannot be avoided. It becomes an issue when there is a circulating flux inside stator creating a leakage inductance increasing the total inductance in a significant manner. Thus, total stator inductance can be given by

$$L_s = L_{ls} + L_{ms} \tag{3-4}$$

The first part denoted by $L_{ls}$ in the equation is called the leakage inductance and causes a large voltage-drop reducing the power factor. Though, in an electric machine operating in the linear region of the $BH$-curve, this leakage does not necessarily affect the torque generation. But in practical high torque density machines, the effects of $L_{ls}$ can be even severe and the severity of such leakage flux will be explained in next section.

### 3.2.3 Flux Fringing, Leakage, and Saturation

As seen from Figure 3-2 (b) and Figure 3-4(b), despite having nearly constant no-load flux linkage, there is a drastic drop in torque production for motors with different pole number. In an ideal situation, the no-load back EMF or flux linkage constant if often considered as a measure of motor performance. In this section, the effect of pole number on stator inductance and corresponding performance degradation will be explained using time domain flux of the motors.

Total stator flux for with different stator excitation levels have been shown in Figure 3-6 (a). For zero current excitation, the total flux linkage and the magnet flux linkage $\lambda_m$ coincides and this no-load flux linkage is shown in orange line. The interesting fact about this no-load flux is for both 20-pole and 40-pole designs, the no-load flux remains very similar and this indicates near non-existent rotor leakage. Or a more accurate statement would be that the rotor leakage is independent of number of poles which creates a possibility of a machine with linear torque-pole number relationship.
Figure 3-6: Total flux linkage $\lambda_T$ for no-load and 2A and 4A stator excitation for 20-pole and 40 pole designs. (b) Difference between the $\lambda_T$ with the two excitation levels with the magnet flux $\lambda_m$ (c) Excitation currents.

As the motor is excited with a current orthogonal to the magnet flux as shown in the phasor diagram in Figure 3-5, the total flux $\lambda_T$ is expected to change and by that can be seen by the blue and red lines in the figure. But comparing the total flux for a 20-pole motor with that of the 40-pole motor clarifies the difference in performance of these motors in loaded condition. Despite having same current MMF, the 40-pole motor exhibits much higher flux linkage compared to the 20-pole counterpart. This explains the radical drop in the power factor of the motor with large pole number. To explain the drop in the torque production, a more detailed analysis is needed. In Figure 3-6 (b) the difference between total flux $\lambda_T$ and the no-load magnet flux $\lambda_m$ are shown for two different excitation levels. In an ideal condition, this flux difference $\lambda_T - \lambda_m$ can be considered a sole contribution of stator MMF and that would result in linear relationship between this flux with the current excitation as $\lambda_s = L_sI_s$. But despite sinusoidal excitation as shown in Figure 3-6 (c) , the armature reaction flux is seen to have large harmonics for the 40 pole design which indicates considerable saturation in the magnetic circuit. Thus, the saturation in flux linkage due to magnetic
saturation causes a substantial drop in torque generation. As seen from the figure, the armature reaction increases with increase in pole number making the high number of pole an unattractive choice at a certain point.

![Image](a)

**Figure 3-7**: Flux density vectors with stator excitation for different pole numbers (a) 20 pole (b) 32 pole (c) 40 pole.

So far, the effects of high stator leakage in the motor performance have been discussed. To explain the cause of leakage, closer look at the finite element analysis is required. Since same coil slot and stator pole design was used for all the motors, such an intense change in only the motor pole number cannot be explained if idealized flux path is considered. To evaluate the accurate stator leakage flux effect, the simulation was performed with nullifying the magnet MMF and only pushing stator current. The vectors representing the flux in the central stator pole are shown for 20-pole, 32-pole and 40-pole designs in Figure 3-7.
Figure 3-8: Effect of flux fringing with increasing pole count. (a) 2D plane placed inside the stator pole face (b) Flux density for 20-pole TFM (c) Flux density for 32-pole (d) Flux density for 40-pole.

The FEA results exhibit the actual representation of flux lines and the effect of number of poles can be truly understood. The flux lines tend to diverge near the airgap and this results in a larger effective air-gap area and the phenomena is referred as flux fringing. Though flux fringing effect is often ignored in analytical machine model, this is obviously not an option for transverse flux topology as evident from Figure 3-7. Despite having similar air-gap, the higher pole count resulted in significantly larger flux density in the core which resulted in larger inductance. Since fringing takes place near the edges of the pole, larger pole count increases the number of edges in the system and thus results in a large increase in the effective air-gap cross-sectional area. To get a relative comparison of the numerical values of flux density inside the core, a 2D plane was placed inside the stator pole face as shown in Figure 3-8(a). For identical current MMF and absence of rotor MMF, flux density plots for a 20-pole, 32-pole and 40-pole machines are given in Figure 3-8(b), (c) and (d) respectively. Despite having same current MMF and same geometric air-gap length,
larger pole counted resulted in a substantial higher flux density with higher crowding of flux near the edges. This contributes to the definitive decline in the performance of these machines with larger pole count.

3.3 Leakage Reduction and Power Factor Improvement

It is evident from the previous discussion that the upsurge fringing flux with the increase in pole count will introduce adverse effect in high pole machines. This causes the main attraction of TFM of keeping magnet loading independent of pole count to stray from ideal condition. Though flux fringing is inevitable in any electric machine, the building technique based on assembling large number of pieces worsen the situation for TFMs.

The main contribution of fringing is to scale up the effective cross-sectional area of the leakage flux path with an increase in pole count. The ratio between the effective and the ideal cross-sectional area of these leakage paths can be referred as fringing factor and this factor usually increases with the decrease in surface area of the face. As this scaling cannot be avoided, at least some attempt can be given to decrease the leakage path cross-sectional area without considering fringing. Thus, an optimization technique based on 2D flux path in the stator will be presented to decrease stator leakage flux and hence improve power factor.

3.3.1 Ideal Flux Path and Deviations

Leakage flux paths responsible for low power factor and desired flux paths for torque production are shown in Figure 3-9 (a). The green solid line in Figure 3-9 (a) shows the desired flux lines both for magnet MMF or coil MMF. Considering field-oriented operation for SPM, stator coil would have zero current at the time of maximum flux linkage ($d$-axis rotor position). Desired rotor flux would follow the green solid line with some small leakage shown in red dotted lines. Due to outer rotor structure, large separation between adjacent stator poles in the air-gap region would reduce the circumferential rotor leakage to a negligible extent. There will be some additional leakage of rotor flux (shown in red dashed lines in Figure 3-9 (a)) causing fraction of the rotor flux not to link with the coil. Since most MMF drops occur in the air-gap and magnets, flux carried to the stator will mostly follow the ideal flux path making these leakage paths insignificant. Figure 3-9 (b) shows such condition when the current MMF on the stator was made zero to represent the $d$ – axis rotor location. Most rotor flux reaching the stator is seen to circulate the coil and no visible leakage flux can be seen between the top and bottom segments of the stator.
The leakage flux becomes dominant at the \( q \)-axis location when the current MMF reaches its peak. Large airgap caused by the magnet causes a reluctance comparable with that of the leakage flux path. Thus, at the stator pole face region, a significant fraction of the flux takes the shortest leakage path (shown in orange dashed lines in Figure 3-9 (a)) instead of circulating the rotor. This causes a large circulating flux in the stator significantly increasing the stator inductance. Moreover, large circulating flux also causes the stator to saturate early which in turns adds harmonics in the back-EMF under loaded conditions. Figure 3-9 (c) shows such leakage flux where most of the stator flux is seen to circulate inside the stator. The small gap between the top and bottom segment of the stator pole seems to be mostly responsible for this large leakage flux. Though fringing is not taken into consideration in this 2D simulation, as 3D flux path is considered, the reluctance associated with these leakage paths will decrease with the increase in pole count due to fringing.

![Image](image_url)

Figure 3-9: (a) Ideal and leakage flux paths (b) Flux density plot for magnet MMF (c) Flux vector plot for current MMF (d) Flux density plot for current MMF.

The geometric adjustment needed to reduce the stator flux is obvious from Figure 3-9 (c) and (d). Flux density associated with stator MMF for some design modifications are presented in Figure 3-10 (a) through (c). It is obvious that significant improvement can be made in the design by this visual tuning. The associated stator and rotor flux through this design progression are plotted in
Figure 3-10 (d). This shows a significant reduction in stator flux $\lambda_s$ with some small sacrifice in rotor flux $\lambda_m$. Though the visual inspection approach is a decent measure to come up with some base design, in the next stage, a more systematic approach of optimization based on some objective function maximization will be presented.

![Diagram](image)

Figure 3-10: Reduction in stator leakage flux with design progress. (a) Highest $\lambda_s$ (b) Small reduction in $\lambda_s$ due to the increased gap between the top and bottom stator segments (c) Further reduction due to increase in the vertical gap between the top and bottom face (d) Variation in $\lambda_s$ and $\lambda_m$ with progression in design.

### 3.3.2 Trade-offs

The unique shape of the TFM stator provides a large number of design parameters to optimize. In this section, only two factors affecting the power factor will be discussed. The stator pole shaped presented in this section in Figure 3-11 are the result of the preliminary tuning of the shape through visual observation of leakage flux as described in the previous section.

From previous discussions, leakage flux paths causing poor power factor is obvious. Addressing the responsible region will impact the power factor with some definite trade-offs. A straightforward
approach to reduce stator inductance would be to increase the gap between top and bottom pole face as shown in Figure 3-11(a). Decreasing both $K_x$ and $K_y$ shown in Figure 3-11 would cause large separation between the top and bottom pole face region reducing the stator leakage, and hence, the inductance.

![Diagram](image)

Figure 3-11: (a) Design suitable for high power factor, and low no-load flux linkage (b) Design suitable for higher no-load flux linkage, and lower power factor.

But this would also reduce effective pole face area reducing the flux linkage and back-EMF. Thinner pole front region (smaller $K_x$) will also result in earlier pole saturation further reducing the flux linkage. Figure 3-11(b) on the other hand shows the design with large no-load flux linkage and back-EMF (large pole face area, and thicker pole front region). But, smaller separation will result in a large armature reaction making poor full load performance. Only increasing the no-load flux linkage with high stator circulating flux will cause the motor to saturate with small stator current causing the back-EMF to have large harmonics even in lightly loaded condition. An optimum design maintaining both no-load flux and lower circulating stator flux must be adopted for satisfactory motor performance. Moreover, changing design parameters that effect the copper region will change the allowable coil turns also affecting the flux linkage and available current MMF at certain current density.

### 3.3.3 Objective Function

To perform the optimization of stator pole shape for power factor improvement, a numerical assessment of leakage flux is required. One straightforward approach is to use a dummy coil in the rotor such that ideal flux path circulates through both the main (shown in orange in Figure 3-11) and dummy coils (shown in cyan in Figure 3-11) resulting in a unity coupling coefficient between the two coils. Since high torque density with decent power factor is desired, an objective function containing all performance factors should be formulated.
The magnetostatics simulation results for varying $K_x$ and $K_y$ are shown in Figure 3-12. As anticipated, air gap flux is found to have maxima towards higher $K_x$ and $K_y$ (Figure 3-12 (a)). At the same time, the coupling coefficient between the main and dummy coils follow the opposite trend. Due to different copper regions with change in pole shapes, available current MMF for certain current density will vary as well, and will be directly proportional to the copper area shown in Figure 3-12 (c). Figure 3-12 (d) shows the objective function, which can be expressed as

$$J_o = \phi_{NL} K_c A_c$$

where, $J_o$ is the objective function to be maximized, $\phi_{NL}$ is the no-load airgap flux, $K_c$ is the coupling coefficient and $A_c$ is the coil area. It should be noted that higher coupling coefficients
correspond to a closer to ideal flux path and lower inductance. If a single stator pole is wound with a coil, the flux path comprises of large airgap resulting in a small inductance.

![Figure 3-13](image)

**Figure 3-13:** Comparison between optimized and base structure of stator pole as a function of pole count (a) Mean torque with NdFeB magnet (b) Mean torque with Ferrite magnet (c) RMS voltage with NdFeB magnet (d) RMS voltage with Ferrite magnet.

### 3.3.4 Performance Comparison with the base design

The lowering of armature reaction should bring an overall decrease in stator leakage even with the inclusion of flux fringing in the 3D domain. An apple to apple comparison can be made if the same FEA based experiment presented in section 3.2 is repeated with the optimized stator pole structure. Torque and RMS voltage for different armature excitation as a function of pole count were plotted for both the optimized and base structure.

The large saturation in torque output for base structure can be seen in dotted line whereas for the optimized design, a significant boost in torque output is observed as shown in Figure 3-13. In case
of RMS voltage, the solid lines representing optimized stator can be seen to follow a closer to ideal linear relationship with pole count. The increase in torque and reduction in terminal voltage are favorable conditions for any design and hence the optimization process seems to have a significant impact on the performance.

3.3.5 Optimized Design and Simulation Results

After going through all the necessary optimizations, transient simulations of the designed TFMs were performed in Ansys Electromagnetic Suites 17.0 with conventional maximum torque per ampere (MTPA) with flux weakening control scheme. Simulating practical control scheme like MTPA for all torque speed conditions requires simulating the motor with all possible current vectors with reasonable resolution (both for vector magnitude and angle) and performing such 3D simulations for the complete motor would be impractical. Thus, single-phase of the motors was simulated with the use of appropriate periodicity. This will neglect the effect of phase coupling but since more than 6 mm axial gap is aimed for the prototype, this effect would be negligible. To compare the effect of different design variables, five different designs were chosen for testing. Four of them (design 1, 2, 4, 5) were taken from the four corners of the plot shown in Figure 3-12 with the extreme values of $K_x$ and $K_y$. The third design with $K_x = 4.55$ and $K_y = 15.3$ represents the design with maximum objective function. Only this variant was constructed, and experimental results are available for this optimized design only. It is obvious from Figure 3-12 (c) that the five designs will have three different copper areas (for three different values of $K_x$), and hence, different current MMFs. Design 1 and design 4, having the largest copper area will allow larger current MMF with same current density. Design 2 and design 5 on the other hand have the smallest copper area. The optimized design (design 3) was wound with 110 turns of 20 AWG wire. The number of turns for the other four designs were approximated by their respective copper areas. Putting a different number of turns with the same AWG wire will ensure the same current rating requirement on the inverter side which is an essential condition for a reasonable comparison of torque-speed envelopes. The number of turns and some performance parameters for the five machines are given in Table 3.2.

The no-load back EMF at 50 RPM for the five designs are shown in Figure 3-14 (a). It is obvious from Figure 3-14 (a) that design 3, 4, and 5 have larger back EMFs having large pole face area compared to that of design 1 and design 2. Design 5 has comparable back EMF compared to that
of design 3 and design 4 despite having a significantly lower number of turns. This is due to very high air-gap flux for that design as seen in Figure 3-12 (a). As the machines are loaded with current MMF, the effect of armature reaction can be evaluated. All machines were operated with 6.5A RMS current resulting in different current MMF due to different number of turns. Despite having similar no-load back EMFs, maximum attainable torque varies for designs 3, 4 and 5. Design 5, having lower coil turns, and hence, smaller current MMF, gives 8% lower torque compared to design 3. Design 4, on the other hand, allows larger current MMF due to larger copper area, but fails to outperform design 3 in terms of torque production.

Table 3.2: Performance Comparison to Validate Optimization Process

<table>
<thead>
<tr>
<th>Design</th>
<th>Number of Turns</th>
<th>$V_{dc} = 120$ V</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$I_{RMS} = 6.5$A</td>
<td>$I_{RMS} = 3.9$A</td>
</tr>
<tr>
<td>Torque (Nm)</td>
<td>Power (W)</td>
<td>Torque (Nm)</td>
</tr>
<tr>
<td>1</td>
<td>116</td>
<td>34</td>
</tr>
<tr>
<td>2</td>
<td>104</td>
<td>33</td>
</tr>
<tr>
<td>3</td>
<td>110</td>
<td>37</td>
</tr>
<tr>
<td>4</td>
<td>116</td>
<td>36</td>
</tr>
<tr>
<td>5</td>
<td>104</td>
<td>34</td>
</tr>
</tbody>
</table>

The torque-speed envelopes in Figure 3-14 (b) are produced considering 120-volt DC bus voltage, 6.5A RMS current, and 80-degree Celsius coil temperature. Conventional space vector modulation PWM scheme [26] were considered to compute the voltage limit. A motor with lower leakage flux will have a smaller inductive drop, and hence, larger corner speed. As discussed earlier, higher coupling coefficient should result in lower leakage flux and hence larger corner speed. The concept can be verified from the torque-speed plot in Figure 3-14(b). Despite having lower maximum torque, design 1 and design 2 have the largest corner speeds of all designs. Primary reasons behind this are having lower flux linkages and back EMFs for these two designs. The effect of exhibiting poor performance due to lower coupling coefficient can be evaluated by comparing the corner speeds of design 5 with that of design 3. Despite having similar no-load back-EMF compared to design 3, design 5 has the lowest corner speed of all the designs due to very high circulating flux which gives rise to a very high leakage inductance. Large voltage drop in this leakage inductance forces the machine to go to flux weakening region at a lower speed compared to other designs. Having a higher torque production capacity makes the optimized design (design 3) outperform all
designs producing the maximum envelope up to 150 RPM. Only for a brief operating region ranging from 150 RPM to 225 RPM, design 2 gives larger torque compared to all other designs. It should be noted that operating power factor for all 5 designs ranges between 0.4-0.5 despite all optimization. This is due to large current loading and high torque density operation. Power factor is a strong function of current MMF and drops rapidly for high torque density operation [13]. Torque-speed envelopes for a lower current operation with 3.9 A RMS current and same bus voltage are presented in Figure 3-14(d). Since the machine operates at a lower current level than the limiting range in the flux weakening region, the same power level is achieved at high speed with some sacrifice of torque at the low-speed region. Thus, if the high peak torque can be compromised, an inverter with lower VA rating can be used resulting in a better power factor operation (in the range of 0.75). Even at this lower current value, the optimized design covers the largest envelope in the torque-speed plot as seen from Figure 3-14(d). The machine also exhibits higher power factor than some machines in the literature [27][28] when operated at comparable torque density region. Some basic machine design and performance parameters based on simulation results of the optimized design (design 3) are given in Table 3.3 and Table 3.4 respectively.

Table 3.3: Preliminary Design Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outer Diameter (mm)</td>
<td>135</td>
</tr>
<tr>
<td>Total Axial Length (mm)</td>
<td>124</td>
</tr>
<tr>
<td>Air gap length (mm)</td>
<td>1.2</td>
</tr>
<tr>
<td>Number of Turns</td>
<td>110</td>
</tr>
<tr>
<td>DC Bus Voltage (Volt)</td>
<td>120</td>
</tr>
<tr>
<td>RMS Current (A) (Peak Torque)</td>
<td>6.5</td>
</tr>
<tr>
<td>RMS Current Density (A/mm²)</td>
<td>10.6</td>
</tr>
<tr>
<td>Magnet Type</td>
<td>N42H</td>
</tr>
<tr>
<td>Magnet Mass (kg)</td>
<td>0.605</td>
</tr>
<tr>
<td>Lamination Mass (kg)</td>
<td>2.96</td>
</tr>
<tr>
<td>Active Mass (kg)</td>
<td>5.35</td>
</tr>
</tbody>
</table>
Table 3.4: Optimized TFM Performance (FEA)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Torque (Nm)</td>
<td>37</td>
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<tr>
<td>Corner Speed (RPM)</td>
<td>70</td>
</tr>
<tr>
<td>Max Speed (RPM)</td>
<td>250</td>
</tr>
<tr>
<td>Volume Torque Density (Nm/L)</td>
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</tr>
<tr>
<td>Torque per mass (Total) (Nm/kg)</td>
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</tr>
<tr>
<td>Torque per mass (Active) (Nm/kg)</td>
<td>8</td>
</tr>
</tbody>
</table>

Figure 3-14: (a) No load back EMF at 50 RPM (b) Torque-speed envelope considering 120 V DC bus and 80 degree Celsius coil temperature and 6.5A RMS current (c) Power envelope considering same constraints as (b) (d) Torque speed envelope considering 120 V DC bus and 80 degree Celsius coil temperature and 3.9A RMS current
3.3.6 Phase-to-Phase Coupling

Phase inductance comprises of both self-inductance and mutual inductance between the phases. Increasing the coupling coefficient between the main and dummy coils would decrease the self-inductance of each coil, but the three phases of the machine are still susceptible to high mutual inductance. This can be taken care off by simply increasing the axial separation between phases at the expense of reduced volumetric torque density. Considering phase 2 to be the one at the middle, coupling coefficients between phases 1 and 2 and between phases 3 and 2 have larger mutual inductance for smaller gap between the phases (shown in Figure 3-15(b)). The top and bottom phases (phase 1 and phase 3) have large separation resulting in a negligible coupling coefficient.

Another effect of smaller phase-to-phase gap is the imbalance in self-inductances as shown in Figure 3-15 (b). Since the central phase is sandwiched between two other phases, this phase is expected to have larger inductance compared to the other two. This imbalance in phase inductance would also complicate the control of the machine as the usual \(dq\) transformation will not hold for these machines.

![Coupling Coefficient vs Axial Gap](image1)

<table>
<thead>
<tr>
<th>Axial Gap (mm)</th>
<th>Coupling Coefficient</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>0.07</td>
</tr>
<tr>
<td>10</td>
<td>0.05</td>
</tr>
<tr>
<td>15</td>
<td>0.03</td>
</tr>
<tr>
<td>20</td>
<td>0.01</td>
</tr>
</tbody>
</table>

![Inductance (pu) vs Axial Gap](image2)

<table>
<thead>
<tr>
<th>Axial Gap (mm)</th>
<th>Inductance (pu)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>1</td>
</tr>
<tr>
<td>10</td>
<td>0.98</td>
</tr>
<tr>
<td>15</td>
<td>0.96</td>
</tr>
<tr>
<td>20</td>
<td>0.94</td>
</tr>
</tbody>
</table>

Figure 3-15: (a) Coupling coefficient among phases as a function of phase-to-phase axial-gap (b) Per-unit value of three-phase self-inductance as a function of phase-to-phase axial-gap.

One half of a single-phase stator is shown in Figure 3-16(a); the machine consists of six similar pieces. Each phase contains two coils connected in series. The full 3-phase stator is shown in Figure 3-16(b). Six ring coils are named as \(a_1, a_2, b_1, b_2, c_1\) and \(c_2\), respectively. Since phase-to-phase coupling can introduce imbalance in stator inductances making the central phase (phase – B) having different inductance compared to others, attention was given to prevent such event.
Individual inductances of six coils are given in Table 3.5. It is seen that all six coils have approximately similar inductances making the individual phases independent of each other. Total phase inductances of the three phases also shows minimal differences. To check the mutual inductances between phases, different phases were connected in series and the measured inductances are given in Table 3.5. Different mutual coupling between phases would have resulted in $L_{ca}$ to be significantly different compared to the other two. Though there are small differences in the numbers, they fall within the measurement accuracy of the measuring device.

![Figure 3-16: (a) Half of a single-phase stator with winding (b) 3-phase stator](image)

### Table 3.5: Open Rotor Inductances

<table>
<thead>
<tr>
<th>$L_{a1}$ (mH)</th>
<th>$L_{a2}$ (mH)</th>
<th>$L_{b1}$ (mH)</th>
<th>$L_{b2}$ (mH)</th>
<th>$L_{c1}$ (mH)</th>
<th>$L_{c2}$ (mH)</th>
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<tbody>
<tr>
<td>19.72</td>
<td>19.72</td>
<td>19.75</td>
<td>19.59</td>
<td>19.78</td>
<td>19.79</td>
</tr>
<tr>
<td>$L_{an}$ (mH)</td>
<td>$L_{bn}$ (mH)</td>
<td>$L_{cn}$ (mH)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>43.29</td>
<td>42.7</td>
<td>43.31</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$L_{ab}$ (mH)</td>
<td>$L_{bc}$ (mH)</td>
<td>$L_{ca}$ (mH)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>85.61</td>
<td>85.35</td>
<td>84.3</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

### 3.4 Magnet Thickness Optimization

Magnet thickness is one of the major design parameters that can play a critical role in machine performance. The optimization process in this TFM has some complex relationship since the outer dimension of the motor was fixed at the beginning of the design procedure. Overall, the thickness of a surface magnet permanent magnet machines can have following trade-offs for varying magnet thickness—
i) Very thin magnet will result in smaller flux linkage and hence smaller torque. Thinner magnets also have the risk of easier demagnetization.

ii) Thicker magnets produce larger flux-linkage but the effect saturates after certain thickness is reached. Once the flux linkage goes towards saturation, adding magnets becomes unnecessary and can contribute towards cost increase. Though more immune to demagnetization, thicker magnets also make the high-speed flux weakening operation more difficult.

iii) In this design, the outer motor dimension was made fixed and thus thicker magnets reduced space for other materials such as coil slot area. This made the magnet thickness selection even trickier.

Figure 3-17: Motor flux linkage and torque performance with magnet thickness variation (a) Air-gap flux and flux linkage for NdFeB TFM (b) Torque performance for two different excitation levels for NdFeB TFM (c) Air-gap flux and flux linkage for Ferrite TFM (d) Torque performance for two different excitation levels for Ferrite TFM.

The primary target was to design a high torque machine with rare earth magnets and hence the optimization process was focused on the NdFeB based rotor design. Later, a similarly sized rotor
was used to test the motor with Ferrite magnets. As mentioned earlier, the air-gap flux and the flux circulating the coil increases with the increase in magnet thickness as seen in Figure 3-17(a) for NdFeB magnet and Figure 3-17(c) for Ferrite magnet. But there is hardly any increase in flux linkage once the magnet thickness reaches 2.5 mm for NdFeB magnet. For Ferrite though the saturation does not take place with this thinner due to the low energy density of these magnets. But as said earlier, the outer and inner dimension of the motor was fixed and increasing the magnet thickness yield smaller copper area. The flux linkage \( \lambda_m \) therefore peaked at a different magnet thickness (1.5 mm for NdFeB and 3 mm for Ferrite). With a constant current density, the amount of stator MMF that can be supplied also depend on this coil slot area and therefore must be optimized as well. If the motors are excited with similar current density, the torque yield for different magnet thicknesses are shown in Figure 3-17(b) and (d) for NdFeB and Ferrite structure respectively. For two different excitation levels, the maximum torque output can be found for 2.5 mm of magnet thickness for NdFeB magnet. Though the amount of torque increment is small compared to 2 mm magnet thickness, this thickness was used to ensure added robustness against demagnetization.

### 3.5 Demagnetization Analysis

Demagnetization of permanent magnets is one of the prime concerns in PM machine design and utmost care is required to prevent such occurrence. \( BH \) curve of N42SH grade magnets for different temperatures are shown in c. In the absence of any external magnetic field, the magnet operates somewhere along this line depending on the overall permeance of the magnetic flux path. With negligible stator core saturation, the operating point of the magnet ignoring the reluctance drop in the stator can be calculated using the magnet thickness and effective air-gap length and is shown by the black dotted line in the figure.

In the presence of an external magnetic field, the operating point of the magnet will shift in either direction depending on the direction of the external field. The most common cause of such excitation is the high-speed operation of the motor which requires exciting the stator coil which opposes the magnet flux and hence reduces total flux linkage. This process eventually reduces the terminal voltage and ensures high-speed motor operation beyond corner speed considering the inverter constraints. Since the excitation imposes a magnetic field that opposes the permanent magnet field, the operating point shifts left and hence reduces the overall average flux density in the magnet. Withdrawing such field will cause the magnet operating point to come back to the
previous steady state point given that the external field was not strong enough to permanently demagnetize the magnet. Permanent demagnetization occurs only with the application of an external field strong enough that the operating point moves beyond the knee point in the curve. The knee point is the point where the straight line takes the sharp bend and a sharp drop toward x-axis takes place. The magnetic field strength responsible for this permanent demagnetization is called coercive force or coercivity and denoted as $H_c$.

To better understand the demagnetization risk of permanent magnets, a closer look at the demagnetization curve shown in Figure 3-18 (a) or (b) is required. The knee points of permanent magnets are a strong function of magnet thermal grade and operating temperature as evident from the figure. For NdFeB magnets, higher temperature reduces the remnant flux density. At the same time, the magnet coercively reduces or shifts right making these magnets more susceptible to demagnetization at elevated temperature. The reduction in remnant flux density $B_r$ or magnetic field strength further escalade the dogmatization process since there is a weaker magnetic field to oppose the external demagnetizing field. In the demagnetization study, magnet temperature of 120°C will be considered to accommodate the worst-case scenario for NdFeB magnet structure.

The Ferrite magnets are also susceptible to demagnetization, but the temperature relationship is quite different from that of NdFeB as evident from Figure 3-18(b). At higher temperature, the remnant flux density reduces, unlike the NdFeB magnets. But the coercive force of these Ferrite magnets shows opposite trend and increases with the increase in temperature. This makes these magnets more susceptible to demagnetization at a lower temperature. To address the worst-case scenario, room temperature demagnetization curve was considered for the Ferrite rotor structure.

![Figure 3-18: Demagnetization curve and permeance lines for two different air-gaps (a) NdFeB (b) Ferrite.](image)
Some preliminary idea about the simulation process is shown in Figure 3-19. For both NdFeB and Ferrite magnet variants, a rectangular plane was taken 0.1 mm inside the magnet surface facing the air-gap. In the simulation tool, magnets are characterized as linear magnets ignoring the bending of the curve at knee point. There are some iterative ways to incorporate the non-linearity at the knee point. Since the primary objective was not to evaluate the machine performance beyond the knee point, the simple linear magnet was assumed for calculation. As evident from Figure 3-19(b), until the knee point is reached, both $BH$ curve would yield identical performance in all cases. The only objective is to see if the magnet reaches the knee point and thus any result resulting in field strength higher than the knee point field strength can be ignored and the magnet can be considered demagnetized.

![Simulation model and approximation for demagnetization analysis](image)

**Figure 3-19:** Simulation model and approximation for demagnetization analysis (a) $d$-axis magnet location (b) Linear assumption of magnet $BH$ curve.

The machine was optimized and robustness against demagnetization tested for initially intended air-gap length of 0.7 mm. Construction difficulty of the first prototype later forced the air-gap length to 1.6 mm and 1.5 mm for NdFeB and Ferrite structure respectively. Using FEA tool, a negative $d$ – axis current was injected which creates a magnetic field directly opposing the permanent magnet field. The effective flux linkage of the coil reduces with the increase in this negative $d$ – axis current as seen in Figure 3-20 (a). The current responsible for completely nullifying the flux linkage is called characteristics current and determines the high-speed operation capability of the machine. It is obvious that the flux linkage drops for the higher air-gap and so does the characteristics current. Negative $d$ – axis current beyond this point is not required for normal motor operation as it creates a negative flux linkage resulting in an increase in terminal voltage. Just to ensure robustness against any failure in control algorithm or inverter operation,
$d$-axis current was pushed till two times the rated current. The maximum magnetic field inside the magnet facing the air-gap region is shown in Figure 3-20(b). As seen from Figure 3-18(a), the load line shifted left due to the increase in air-gap length, the effect can be realized from the maximum magnetic field in Figure 3-20(b). The structure with larger air-gap exhibit higher magnetic field with no external excitation which eventually indicates being more susceptible to demagnetization. As the external demagnetization current increases, so does the magnetic field but it never reaches the demagnetization point of 750 kA/m even with the excitation level of two times the peak current. The magnetic field for two different air-gap eventually intersects as the larger air-gap also makes it harder for stator magnetic field to pass the large air-gap and demagnetize the magnet.

For the Ferrite magnet structure as well, increasing the demagnetizing current reduces the flux linkage and characteristics current can be found to be much smaller compared to the NdFeB variant as seen from Figure 3-21(a). The worst case coercively of the magnet grade used is 375kA/m and the 0.7 mm structure is found to surpass the limit at relatively small current compared to the NdFeB variant. The final prototype with 1.5 mm air-gap, however, is more reluctant to demagnetization and a peak current of 10A along $d$-axis is required to demagnetize the magnet. As stated earlier, due to a linear approximation of magnet $BH$ curve, the points beyond this point is not a true representation of the magnet status and these results should be ignored. These points in Figure 3-21 are marked with dark diamond tabs and should be ignored.

![Figure 3-20](image_url)

Figure 3-20: Effect of demagnetization current on NdFeB based design (a) Flux linkage (b) Magnetic field strength inside the magnet.
Figure 3-21: Effect of demagnetization current on Ferrite based design (a) Flux linkage (b) Magnetic field strength inside the magnet.
Chapter 4: Torque Ripple Minimization

Like all electric machines, TFMs would have torque ripple issues that must be addressed. To fully understand the torque ripple optimization process, sources of the ripple must be analyzed and modeled in this section. The proposed topology described here has broadly four major sources contributing to torque ripple [29]

i) The cogging torque generated by the interaction between rotor magnet and stator poles [30]

ii) Ripple due to the interaction of the flux linkage harmonics and current harmonics,

iii) Saturation of the magnetic circuit and

iv) Issues caused by motor control algorithm and practical limitations e.g. error in feedback signals e.g. rotor position, phase currents. In the article, only the first two sources and remedy will be discussed.

4.1 Cogging Torque Fundamental

Since cogging torque is inherent for all PM motors, it has been considered as a popular research area and different solutions have been provided like rotor skewing, stator tooth shaping, magnet shaping etc. But most of these techniques have been addressed for more common radial flux machines [31][32] and not all of them are applicable for the transverse flux topology. Works specifically targeting TFM torque ripple have been found in [33][34][35] but most of these works targeted TFM made out of soft magnetic core (SMC) material which allows stator pole shaping.

Before addressing the solution, a deeper look into the physics behind cogging torque generation is required.

A typical case of a surface magnet PM machine is shown in Figure 4-1. Figure 4-1.(a) shows the position where rotor magnets are perfectly aligned with the stator poles and maximum flux lines should be crossing through the stator poles at this state. In another word, this can be stated as the rotor location that sees the minimum reluctance in the flux path.

Figure 4-1.(b) on the other hand shows the rotor location where magnets are in between two stator poles, and this will cause in minimum flux crossing through the stator consequently making this the position for maximum reluctance. Inherently the rotor wants to stay at the minimum reluctance position and without any external force, the rotor will tend to stay at the minimum reluctance positions. At any of these locations, magnetic energy of the machine can be calculated from the flux density as
\[ W(\alpha) = \frac{1}{2\mu} \int B^2 dV \]  

where \( \alpha \) is the rotor position. The cogging torque at any rotor position can be calculated as

\[ T_{cogging}(\alpha) = -\frac{\partial W(\alpha)}{\partial \alpha} \]

Since cogging torque depends on the derivative of the co-energy, scenario represented in Figure 4-1 can be considered the worst case scenario where all the magnets are getting aligned concurrently and vice versa. This represents a single-phase machine and usually not common for radial flux machine structure.

![Figure 4-1: Linear representation of stator and rotor structure with \( N_s = N_R \) (a) Aligned position with maximum co-energy (b) Unaligned position with minimum co-energy.](image)

For a typical PM motor with identical PM poles, number of cogging torque periods during a slot pitch rotation is given by [36]

\[ N_p = \frac{N_R}{HCF(N_S, N_R)} \]

where, \( N_R = \) number of rotor poles, \( N_S = \) number of stator slots and HCF corresponds to highest common factor. Lower value of \( N_p \) typically causes high cogging torque and vice versa. In typical motor design, \( N_R \) and \( N_S \) are chosen carefully to avoid lower \( N_p \) and high cogging torque. Peak to peak cogging torque of a radial flux machine with different rotor pole and stator slot combinations are given in Table 4.1. Motor was simulated with same outer radius and stack length for a reasonable comparison. It is obvious from the table data that \( N_p \) plays dominating role in determining the cogging torque and hence torque ripple. Since, cogging torque is the result of reluctance variation, the method described above is applicable for TFM with 3D flux path as well. Unfortunately, the intended flux paths of TFMs require equal number of stator and rotor poles in a single phase (\( N_s = N_R \)) which is the worst case for cogging torque characteristics as seen from Table 4.1 Even if all three phases of the machine are considered as a lump, shifted stator poles of
three different phases is comparable with having three times the stator poles, making the design equivalent to \( N_S = 3N_R \). Still, this makes no change in \( N_P \) essentially creating no substantial benefit in cogging torque reduction. The designed TFM has \( N_R = 32 \) and an equivalent \( N_S = 96 \) which apparently gives high cogging torque and special attention must be given for ripple reduction.

<table>
<thead>
<tr>
<th>( N_R )</th>
<th>2</th>
<th>2</th>
<th>2</th>
<th>4</th>
<th>4</th>
<th>32</th>
<th>32</th>
</tr>
</thead>
<tbody>
<tr>
<td>( N_S )</td>
<td>3</td>
<td>6</td>
<td>9</td>
<td>12</td>
<td>6</td>
<td>9</td>
<td>32</td>
</tr>
<tr>
<td>( N_P )</td>
<td>2</td>
<td>1</td>
<td>2</td>
<td>1</td>
<td>2</td>
<td>4</td>
<td>1</td>
</tr>
<tr>
<td>( T_{cog} ) (Nm)</td>
<td>5.2</td>
<td>9.9</td>
<td>4.4</td>
<td>6.7</td>
<td>11.5</td>
<td>4.7</td>
<td>32.7</td>
</tr>
</tbody>
</table>

### 4.2 Flux Linkage Harmonics Contribution

In the surface PM, the only source of EM torque is called the reactance torque and caused by the interaction between the rotor and stator flux at the air-gap. In other words, it can be represented by a direct interaction between the back-EMF wave and the current waveform and this makes the analysis much simpler. Let’s assume a motor has following back EMF profile

\[
E_a = \sum_{n=1,3,5,...} E_n \cos(n\omega t) 
\]

\[
E_b = \sum_{n=1,3,5,...} E_n \cos\{n\left(\omega t - \frac{2\pi}{3}\right)\} 
\]

\[
E_c = \sum_{n=1,3,5,...} E_n \cos\{n\left(\omega t + \frac{2\pi}{3}\right)\} 
\]

It should be noted that back EMF usually only holds the odd harmonics and the elaborate reasoning behind will be explained in next section. Usually, the injected current is expected to be purely sinusoidal having only the first harmonics, but control issues can cause the current to have some harmonics and the back-EMF harmonics can be attributed as the source of these harmonics. Consequently, phase currents also exhibit the presence of harmonics.

\[
I_a = \sum I_n \cos\{n(\omega t - \beta)\} 
\]

\[
I_b = \sum I_n \cos\{n\left(\omega t - \beta - \frac{2\pi}{3}\right)\} 
\]
\[ I_c = \sum I_n \cos \left\{ n \left( \omega t - \beta + \frac{2\pi}{3} \right) \right\} \]  

If torque generation for only one phase is calculated, it can be applied for all three phases. At a certain instant, input power taken by phase A can be represented as

\[ P_a(t) = \left[ \sum_{n=1,3,5,...} E_{an} \cos(n\omega t) \right] \left[ \sum I_{an} \cos\left\{ n(\omega t - \beta) \right\} \right] \]  

\[ T_a(t) = \left( \frac{1}{\omega_m} \right) \left[ \sum_{n=1,3,5,...} E_{an} \cos(n\omega t) \right] \left[ \sum I_{an} \cos\left\{ n(\omega t - \beta) \right\} \right] \]

The interaction between all the harmonics in back-EMF and currents results in certain harmonics components and can be easily calculated using trivial trigonometric calculations.

Due to three-phase structure, all resulting orders which are even and not divisible by 3 will be separated by 120 electrical degrees and will have no effect on the three-phase system. All components that are divisible by 6 will add up in phase and will create major issues in torque performance. The following summary can be helpful in determining the responsible back-EMF components

i) First order back EMF and first-order current will create zeroth order (DC) component and second order ripple. Since second order components from three different phases will be separated electrically by 120 degrees, no contribution will be seen in the machine output.

ii) Third order harmonics will create 2nd and 4th order with sinusoidal current resulting in no visible ripple. In case there are higher order harmonics in phase currents, other unwanted harmonic components will be generated which might be visible in output torque.

iii) Fifth order of back EMF will create a 4th and 6th order ripple components. Though 4th order will have no effect, 6th order will be added in phase for three-phase systems and will be the major source for torque ripple. Since this component is created by interacting with first-order of current, the main torque producing current harmonic will be responsible for ripple production and this is the largest component as the first order current is the primary current order and the only order in case of a very good current control algorithm.
4.3 Understanding of Torque Ripple through frequency domain analysis

A mathematical model of the electromagnetic system is often the best option to explore the different options available to improve torque ripple performance. In this section, a simplistic model of the magnetic flux path and corresponding flux linkage and co-energy will be modeled. As the total flux going into the stator pole depends on the combined reluctance of the associated flux path at certain rotor location, flux entering towards the stator at any rotor location \( \theta \) can be expressed as

\[
\phi(\theta) = \frac{F_m(\theta)}{R(\theta)} \quad 4-12
\]

\[
\phi(\theta) = F_m(\theta)P(\theta) \quad 4-13
\]

\[
\lambda(\alpha) = \int_0^{2\pi} \phi(\theta) d\theta \quad 4-14
\]

where \( R(\theta) \) and \( P(\theta) \) are the reluctance and permeance of the flux path at location \( \theta \). Typical mapping of flux path permeance resulting from one stator pole for a TFM is shown in Figure 4-2. This figure shows the permeance to have non-zero values at a span of 90 degree which represents a pole arc to pole pitch ratio of 0.5 for stator pole. The ideal magnet MMF is drawn for a magnet pole arc to pole pitch ratio of 0.75 making the magnets to produce non-zero MMF for a span of 135 electrical degrees. Air-gap flux at this rotor location (\( \alpha = 0 \)) is shown on the third plot which is simply the product of path permeance and air-gap MMF. The total flux linkage at this rotor location is the integration of this air-gap flux and is shown in the fourth plot. It should be noted that the flux linkage is a function of rotor position and the fourth plot has a different x-axis than the other three plots. The whole procedure can be represented by a circular convolution of the air-gap MMF and flux path permeance and can be represented as

\[
\lambda(\alpha) = (F_m * P)(\alpha) \quad 4-15
\]

This straightforward representation of the process allows a simple evaluation of the process through frequency domain analysis since the circular convolution in space domain can be substituted by a direct multiplication in the frequency domain.

\[
F(\lambda) = F(F_m * P) = F(F_m).F(P) \quad 4-16
\]

where \( F \) represents the Fourier transform operator.

All these plots are drawn with no motor dimension and this makes the analysis simple and straightforward. Only variable in the figures are the pole pitch to pole arc ratio for both stator poles.
and rotor magnets. Since there only two stator poles are sufficient for calculation, flux linkage caused on two poles can be added to account for total flux linkage in the motor. In this TFM, stator pole arc to pole pitch ratio or pole arc coefficient at air-gap can be calculated as

\[ \frac{P_A}{C_s} = \frac{S_{W}}{2\pi R_{AG} P} \]  

where, \( S_{W} = 2 R_{in} \sin(OSF \times \frac{p_l}{P}) \).

Since the magnet MMF has no DC component, all DC components can be ignored in the calculation. The \( n^{th} \) harmonic component for the flux path permeance can be denoted as

\[ P_{an} = \left( \frac{2}{n\pi} \right) P_{m} \sin\left( \frac{n\pi P_{AC_s}}{2} \right) \quad [n = 1, 2, 3 \ldots] \]  

where \( P_{m} \) represents the maximum permeance magnitude at aligned position. All the sine components will be equal to zero due to the even symmetry of the waveform. Likewise, \( n^{th} \) harmonic component for air-gap MMF can be shown as

\[ F_{m,an} = \left( \frac{2}{n\pi} \right) F_{m} \sin\left( \frac{n\pi P_{AC_m}}{2} \right) \quad [n = 1, 3, 5 \ldots 2m - 1] \]  

where \( F_{m} \) is the magnitude of the rectangular rotor MMF. The air-gap MMF will have only the odd harmonics components and the sin terms will be zero for even symmetry. Thus, the flux linkage component will also have only odd harmonic components and so does the back EMF. Flux linkage created by each magnet can be calculated and added considering linear operation region in the lamination. Considering air gap MMFs (upper and lower pole face) and a series connection of the coils, the total harmonic magnitude will be four times the flux created by one pole and one air-gap MMF. So, the \( n^{th} \) harmonic component of the total flux linkage can be denoted as

\[ \lambda_{an} = \left( \frac{16}{n^2\pi^2} \right) P_{m} F_{m} \sin\left( \frac{n\pi P_{AC_s}}{2} \right) \sin\left( \frac{n\pi P_{AC_m}}{2} \right) \quad [n = 1, 3, 5 \ldots 2m - 1] \]  
\[ \lambda_{an} = \left( \frac{16}{n^2\pi^2} \right) P_{m} F_{m} \left[ \cos\left( \frac{n\pi}{2} (P_{AC_m} + P_{AC_s}) - \cos\left( \frac{n\pi}{2} (P_{AC_m} - P_{AC_s})) \right) \right] \quad [n = 1, 3, \ldots 2m - 1] \]  

Back EMF harmonic component can be obtained using Fourier transform theorem and denoted as

\[ E_{an} = \left( \frac{32}{n\pi} \right) P_{m} F_{m} \left[ \sin\left( \frac{n\pi}{2} (P_{AC_m} - P_{AC_s}) - \sin\left( \frac{n\pi}{2} (P_{AC_m} + P_{AC_s}) \right) \right] \quad [n = 1, 3, 5 \ldots 2m - 1] \]  

It is evident from the Fourier components that specific harmonic components can be canceled by selecting the appropriate magnitude of either \( P_{AC_s} \) or \( P_{AC_m} \). Since \( P_{AC_s} \) depends on the rotor
inner radius and over span factor, it to maximize torque production, it is more appropriate to choose proper magnet pole arc to have proper back EMF. To cancel out \(n^{th}\) order harmonic from flux linkage and hence back EMF, \(PAC_m\) can be chosen as

\[
\sin\left(\frac{n\pi PAC_m}{2}\right) = \sin(k\pi) \quad [k = 0, 1, 2, \ldots] \tag{4-23}
\]

\[
PAC_m = \frac{2k}{n} \tag{4-24}
\]

Thus both \(PAC_m = 0.4\) and 0.8 will cancel out the 5th order harmonics. Since higher \(PAC_m\) results in a larger first order component and thus larger torque constant, usually largest possible solution is preferred. Since it is impossible to come up with a \(PAC_m\) that cancels out all unwanted harmonics, minimizing the total harmonic distortion (THD) can be a straight forward objective function where THD can be defined as

\[
THD(E) = \sqrt{E_{a3}^2 + E_{a5}^2 + E_{a7}^2 + \cdots} \tag{4-25}
\]

\[
THD(E) = \sum_{n=3,5,7,\ldots} E_{an}^2 \tag{4-26}
\]

To understand the formation of co-energy, a similar diagram can be used as shown in Figure 4-3. In this case, the first attempt would be to compute the co-energy map for the entire region. Since co-energy depends on the square of the flux density and the system is represented only by bipolar binary signals, taking the absolute value of the rotor flux can be considered as co-energy distribution. In this case, flux path permeance created by both stator poles must be considered. In this case, the path permeance function will have only the even harmonics as evident from Figure 4-3. Thus \(n^{th}\) order Fourier coefficient for path permeance function would be

\[
P_{an} = \left(\frac{4}{n\pi}\right) P_m \sin\left(\frac{n\pi PAC_m}{2}\right) \quad [n = 2, 4, 6, \ldots 2m.] \tag{4-27}
\]

Likewise, \(n^{th}\) harmonic component of the absolute value of air-gap MMF can be shown as

\[
F_{man} = \left(\frac{4}{n\pi}\right) F_m \sin\left(\frac{n\pi PAC_m}{2}\right) \quad [n = 2, 4, 6, \ldots 2m] \tag{4-28}
\]

Considering unit area, air-gap flux can be represented as the flux density giving us the \(n^{th}\) order component of co-energy

\[
W_{an} = \left(\frac{16}{n^2\pi^2}\right) P_m F_m \sin\left(\frac{n\pi PAC_s}{2}\right) \sin\left(\frac{n\pi PAC_m}{2}\right) \sin\left(\frac{n\pi PAC_m}{2}\right) \quad [n = 2, 4, 6, \ldots 2m] \tag{4-29}
\]
\[ W_{an} = \left( \frac{16}{n^2 \pi^2} \right) P_m F_m \left[ \cos\left( \frac{n \pi}{2} (PAC_m + PAC_s) \right) - \cos\left( \frac{n \pi}{2} (PAC_m - PAC_s) \right) \right] \quad [n = 4, 2, 6 \ldots 2m] \]

Thus, the \( n^{th} \) harmonic component of the cogging torque would be

\[ T_{an} = \left( \frac{32}{n \pi} \right) P_m F_m \left[ \sin\left( \frac{n \pi}{2} (PAC_m - PAC_s) \right) - \sin\left( \frac{n \pi}{2} (PAC_m + PAC_s) \right) \right] \quad [n = 4, 2, 6 \ldots 2m] \]

Since only the even harmonics have non-zero coefficients, all the orders not a multiple of three will have a 120° phase shift resulting in no-contribution in the three-phase cogging torque. The only concerns are the orders that are multiple of 6 and mostly the 6\(^{th}\) order itself. Like the back-EMF optimization procedure, it can be found that a \( PAC_m = 2/3 \) can completely nullify this order.

![Graphs](image)

Figure 4-2: (a) Flux path permeance variation as a function of location \( \theta \) (b) Air-gap MMF created by the permanent magnet as a function of \( \theta \) drawn for rotor location \( \alpha = 0 \). (c) Resulting air-gap flux as a function of \( \theta \) drawn for rotor location \( \alpha = 0 \) (d) Flux linkage as a function of rotor position \( \alpha \).
4.4 Objective Function and Optimization

As seen from the harmonic components, there exist specific magnet pole arc coefficients that can completely cancel certain harmonic components from either back EMF or cogging torque. Unlike any optimization process, the optimization of back EMF harmonics and cogging torque is not free from trade-offs. Since a multi-objective optimization is desired, and the stator width or $PAC_s$ was set before to maximize output torque, the process only has one degree of freedom making it difficult to attain all requirements. The desired outcome of the optimization process can be summarized as follows,

1) Minimize cogging torque, primarily the $6^{th}$ order
2) Make the back EMF sinusoidal, hence minimize signal THD.
3) Give special attention to 5th order harmonics in back EMF as it directly interacts with first order current.

4) Last but not the least, maximize the fundamental back EMF or the first order component of the back EMF. This indeed is the torque producing component, and smaller first harmonic results in low torque constant.

A typical variation of different voltage harmonics and THD as a function of $PAC_m$ are shown in Figure 4-4. Three plots are shown for three different OSF options which necessarily corresponds to different $PAC_s$ values. It is obvious that larger OSF results in larger first order harmonic and hence larger torque and the value increases with increasing values of $PAC_m$. The voltage harmonics values also get canceled at the $PAC_m$ calculated from Equation 4-24. Though the first order harmonic is following a predictable pattern with larger OSF resulting in for larger back EMF, the higher harmonic components do not necessarily follow the order. This is due to the difference in effective $PAC_s$ values resulting from different OSF numbers. The values of $PAC_s$ for these OSF fall in a range between 0.45-0.65 making it a dominant contributor in determining the harmonic content and hence THD.

It is obvious from the analytical results that the optimization leads to different $PAC_m$ for different OSF values of stator designs. In most RFMs, $PAC_m$ is usually chosen based on the value of $N_S$ and $N_R$ more importantly, the value is calculated from the least common multiple of $N_S$ and $N_R$. The process is suitable for most RFMs due to the most common stator pole structures adopted by most RFMs which is not suitable for a TFM made with laminated steel. The difference in the stator structure between lamination steel TFM and RFM can be easily realized by comparing Figure 4-6 (a) and (b). A typical outer rotor RFM is shown in Figure 4-6 (a) which shows small slot opening as used in regular motor structure. A 2D equivalent representation of a lamination based TFM is shown in Figure 4-6 (b) displaying the obvious difference. Thus, optimization of THD results in different $PAC_m$ for different stator thicknesses. The term $Sin\left(\frac{n\pi PAC_s}{2}\right)$ for the $n^{th}$ harmonic component acts as a modulating factor and the modulating effect can be seen by comparing the values of the 3rd and 5th harmonic components. In regular radial flux machine, the slot opening is usually kept small making the $PAC_s$ close to unity and reducing the effect of stator thickness in the harmonic distribution. The optimized $PAC_m$ that minimizes the THD is found to be a function of stator thickness. The optimization gets more complicated as this structure represents one of the worst-case scenarios for co-energy variation and thus cogging torque. If the
most dominant 6\textsuperscript{th} order torque ripple is to be minimized, it is evident that the desired $PAC_m$ for minimizing cogging torque and THD does not coincide. In fact, the required value of $PAC_m$ for cogging torque variation falls in the region where the 5\textsuperscript{th} order of back EMF is close to its maxima. This eventually necessitates adding more design variables to give better control over different harmonic orders.

![Graphs showing magnitude of back EMF and THD](image)

Figure 4-4: (a) First order magnitude of back EMF (b) Third order magnitude of back EMF (c) Fift order magnitude of back EMF (d) Back EMF THD.
4.5 Rotor Skewing

To add one more dimension in the optimizing process, a simple and effective solution can be the introduction of a step skewing in the rotor magnets. The effective MMF in the air gap region created by top and bottom magnets are shown in Figure 4-7 and denoted by $F_mT$ and $F_mB$ respectively. The total MMF experienced by the flux path is the summation of these two MMFs. There is a fundamental difference between the effect of skewing taking place in TFM and RFM that can be better visualized by comparing Figure 4-8(a) and (b). In a radial flux machines, the flux path is confined in a 2D plane and the total magnet MMF created in a plane does not necessarily interact with that of another plane. In case of a twostep skewing as shown in Figure 4-8(a), any
segment will experience a magnet MMF similar to $\mathcal{F}_{mT}$ or $\mathcal{F}_{mB}$. The co-energy variation of flux linkage created by these individual MMFs will essentially will have a phase shift due to the spatial shift of the MMF in two different planes. For the TFM on the other hand, the flux path consists of both the top and bottom magnets giving rise to an effective MMF which more steps and by appropriate phase shift, this diagram can be made closer to a sinusoidal distribution. Apparently, it does not seem so different in an ideal condition, but when saturation is considered, the change in MMF profile creates a significant difference in the response as the system loses its linearity at this point. The skewing effect in the TFM can be compared with creating a more sinusoidal MMF in RFM through introducing thickness variation in the magnet. Mathematically following notation can be used for a TFM

$$\lambda_T(\alpha) = \mathcal{f}[(\mathcal{F}_{mT} + \mathcal{F}_{mB}), \mathcal{P}](\alpha) \tag{4-32}$$

where $\mathcal{f}$ represents the function taking care of the non-linearity. In case of a radial flux machine, it can be represented by

$$\lambda_R(\alpha) = \mathcal{f}[\mathcal{F}_{mT}, \mathcal{P}](\alpha) + \mathcal{f}[\mathcal{F}_{mB}, \mathcal{P}](\alpha) \tag{4-33}$$

Since the distributive law does not apply for the non-linear system, Equation 4-32 and 4-33 does not necessarily yield the same results.

For the harmonic analysis, the spatial shift in the magnet MMF can be represented by a phase shift in frequency domain. Since no manipulation is done with the stator, the harmonic components of the permeance will be unchanged. If the total shift is attributed to only the bottom rotor, $n^{th}$ harmonic component for air-gap MMF can be shown as

$$\mathcal{F}_{m_{Tn}} = \left(\frac{4}{n\pi}\right)\mathcal{F}_m\sin\left(\frac{n\pi P_{ACm}}{2}\right) \quad [n = 1, 3, 5 \ldots 2m - 1] \tag{4-34}$$

$$\mathcal{F}_{m_{Bn}} = \left(\frac{4}{n\pi}\right)\mathcal{F}_m\sin\left(\frac{n\pi P_{ACm}}{2}\right)\{\cos(2\pi n S_k) + j\sin(2\pi n S_k)\} \quad [n = 1, 3, 5 \ldots 2m - 1] \tag{4-35}$$

$$\mathcal{F}_{m_n} = \left(\frac{4}{n\pi}\right)\mathcal{F}_m\sin\left(\frac{n\pi P_{ACm}}{2}\right)\{1 + \cos(2\pi n S_k) + j\sin(2\pi n S_k)\} \quad [n = 1, 3, 5 \ldots 2m - 1] \tag{4-36}$$

Here, $S_k$ is the coefficient of skewing. $S_k=1$ represents a full electrical cycle.

When calculating the co-energy distribution, the absolute value of the air-gap MMF will hold the same expression of $n^{th}$ harmonic but will only have even harmonics. Thus, flux linkage will have following odd harmonic components
\[
\lambda_{an} = \left(\frac{16}{n^2\pi^2}\right)P_m F_m \sin\left(\frac{n\pi PAC_s}{2}\right) \left\{\sin\left(\frac{n\pi PAC_m}{2}\right) \left[1 + \cos(2\pi n S_k) + j \sin(2\pi n S_k)\right]\right\} \quad [n = 1, 3, 5 \ldots 2m - 1]
\]

Addition of another degree of freedom gives the opportunity to further manipulating with the harmonic components of both co-energy and flux linkage.

Figure 4-7: (a) Air-gap MMF created by the top row permanent magnet as a function of \(\theta\) drawn for rotor location \(\alpha = 0\) (c) Air-gap MMF created by the bottom row permanent magnet as a function of \(\theta\) drawn for rotor location \(\alpha = 0\) (c) Total effective air-gap MMF as a function of \(\theta\) drawn for rotor location \(\alpha = 0\) (d) Flux linkage as a function of rotor position \(\alpha\).

The effect of skewing can be visualized from Figure 4-9. In Figure 4-9.(a), first harmonic of the back-EMF component is shown which shows an insignificant drop in back EMF or torque constant. But there have been significant improvements in THD and for 20-degree skew, THD is minimum between \(PAC_m = 0.7\) and 0.8. Even greater improvement can be observed when the cogging torque harmonics are compared.
Figure 4-8: Two-step skewing for a radial flux machine (b) Two-step skewing for TFM.

Figure 4-9: (a) First order magnitude of back EMF (b) Back EMF THD (c) Sixth order cogging torque (d) 12th order cogging torque.

4.6 Stator Pole Asymmetry

The asymmetry of the stator by the axial plane adds some asymmetric fringing pattern in the air-gap flux giving rise to some additional shifting in the flux linkage. According to Figure 4-7
irrespective of the magnet skewing, the two stator poles are always 180-degree phase shifted and will experience the identical but inverted air-gap MMF and hence produce identical flux linkage. But as seen in Figure 4-10 (a), when the top part of the adjacent stators are aligned with the magnet, instead of experiencing an air-gap flux density of equal but opposite polarity, there is definite variation in the flux pattern. This can be better understood if the difference in pole face shape is considered as shown in Figure 4-10 (b). When the middle pole is perfectly aligned with the magnet with pole face-2, the adjacent stator poles are aligned with the magnet of opposite polarity with pole face-1. Thus, the flux linkage experienced by the adjacent stator poles will not be identical and rather have a phase shift between them. In the ideal modeling this can be represented by introducing variation in the path permeance created by top and bottom segments of the stator as seen in Figure 4-11. The top layer permeance of stator pole 1 is identical to the bottom layer of the adjacent or stator 2 permeance as the stator poles are created by flipping the poles upside down. This creates a phase shift between the flux linkage making it impossible to use the parallel connection between the coils.

Figure 4-10: (a) Asymmetric flux fringing in front of the stator pole face (b) Difference in geometry between the top and bottom segments of stator pole face.
4.7 Finite Element Analysis of pole arc coefficient and skewing contribution

Though the derivations of harmonic components were done considering the idealized linear system, it is useful to predict the basic pattern of different harmonic components. But to find the exact magnitude and to properly optimize the design, finite element analysis is the best option. This takes the flux fringing, leakage and non-linearity into account and often provides results close to experimental values.
FEA results of optimization procedure for Ferrite magnet design are shown in Figure 4-12. Figure 4-12(a), (c) and (e) demonstrate the contribution of $PAC_m$ with no skewing in magnet placement. It is evident that the first harmonic component of back EMF or effective torque constant increases with increasing $PAC_m$. Depending on the higher order harmonic contribution, minimum THD is achieved for some value of $PAC_m$ (0.65 for $OSF = 1.4$) that is close to the ideally computed value. But the best THD can be achieved at a significant sacrifice of torque production and does not represent the best option for cogging torque reduction ($E_1 = 83$ for $OSF = 1.4, PAC_m = 0.65$)

By choosing the $PAC_m$ as 0.8 which resulted for best cogging torque performance and moderate first order back EMF, introducing a shift in the magnet placement will add a small drop in back EMF as seen by Figure 4-12 (b). But there is a significant improvement in THD and cogging torque becomes near non-existent as seen from the results.

The results for the rotor with NdFeB magnets are found to diverge more compared to the analytical results as it pushes the system towards saturation making the flux linkage to saturate. This causes the THD of the back EMF to be significantly higher compared to that with Ferrite magnet structure as shown in Figure 4-13. The primary objective was set to reduce cogging torque and an electrical shift of 40 degrees minimized the cogging torque when $PAC_m$ was set to 0.8. Due to saturation, the drop in first order back EMF or the torque constant is found to be less effected by the skewing process and with the optimized $PAC_m$ and skewing, the THD of the back-EMF remained larger than 13% for the motor with $OSF = 1.4$. 
Figure 4-12: Optimization process of Ferrite magnet rotor (a) Back EMF first harmonic variation for varying $PAC_m$ (b) Back EMF first harmonic variation for varying skew angles (c) Back EMF THD variation for varying $PAC_m$ (d) Back EMF THD variation for varying skew angles (e) Peak to peak cogging torque variation for varying $PAC_m$ (f) Peak to peak cogging torque variation for varying skew angles.
Figure 4-13: Optimization process of NdFeB magnet rotor (a) Back EMF first harmonic variation for varying $PAC_m$ (b) Back EMF first harmonic variation for varying skew angles (c) Back EMF THD variation for varying $PAC_m$ (d) Back EMF THD variation for varying skew angles (e) Peak to peak cogging torque variation for varying $PAC_m$ (f) Peak to peak cogging torque variation for varying skew angles.
4.8 Effects of Phase Shift Error

As shown in previous chapters, a multiphase TFM is necessarily formed by axially tacking multiple single-phase structures with appropriate phase shifts. As most TFMs are designed to achieve high torque density and thus have a large number of poles, a strict tolerance is indispensable especially in maintaining the required phase shift. As an ideal 120-degree phase shift was assumed in the analytical and FEA results, performance of a practical machine will inevitably deviates from this ideal case. Even if the machine is ideally optimized to the extent of producing near zero torque ripple, presence of phase shift error can eventually result in unusual harmonic components compared to the radial flux machines. If a machine is excited with pure sinusoidal excitation, torque components generated by three phases can be expressed with the following equations

\[
T_a(t) = \left( \frac{1}{\omega_m} \right) \left[ \sum_{n=1,3,5,...} E_{an} \cos(n\omega t - \frac{2\pi}{3} - nO_a) \right] \left[ I_M \cos(\omega t - \frac{2\pi}{3} - \beta) \right]
\]

\[
T_b(t) = \left( \frac{1}{\omega_m} \right) \left[ \sum_{n=1,3,5,...} E_{bn} \cos(n\omega t) \right] \left[ I_M \cos(\omega t - \beta) \right]
\]

\[
T_c(t) = \left( \frac{1}{\omega_m} \right) \left[ \sum_{n=1,3,5,...} E_{cn} \cos(n\omega t + \frac{2\pi}{3} - nO_c) \right] \left[ I_M \cos(\omega t + \frac{2\pi}{3} - \beta) \right]
\]

where \(O_a\) and \(O_c\) are the phase shift offset for phase-A and phase-C respectively assuming the central phase or phase-B as a reference. Presence of this error will negatively affect the torque performance of the machine even in the absence of any higher order harmonics in the back EMF. Since a perfect first order component of the back-EMF is responsible for both DC and 2\(^{nd}\) order torque component with pure sinusoidal excitation, presence of phase offsets will essentially-

i) Reduce DC or mean torque as the first order stator MMF for three phases will not be orthogonal to the rotor MMF for all three phases. In other words, the motor will not be operating with a perfect \(q\) axis current for all three phases.

ii) Since both zeroth order and second order torque essentially have the similar peak magnitude, a small phase shift error will lead a large second-order ripple since the second order components no longer being in a perfectly 120-degree phase shift condition.
A closer look into the cogging torque components will reveal that the stator and rotor pole combination in these machines causes a large second order component in the individual cogging torque generated by phases. Failure to attain a perfect 120-degree phase shift will also create a large second order component in the cogging torque.

4.9 Time Domain Signal and Harmonic Analysis

Due to structural issues, the first attempted NdFeB rotor failed to maintain air-gap and thus perform smoothly as a working prototype. A new rotor was built afterward with modified mechanical structure and larger air-gap. Despite lacking structural robustness, some initial data were extracted from the first failed prototype. Extensive tests were performed only on the two working rotors – one with Ferrite magnet and the other with NdFeB second prototype referred as NdFeB Gen-2 in the report.

4.9.1 Back EMF and Flux Linkage

The back EMF profile for both the Ferrite and NdFeB magnet variant of the TMFs are shown in Figure 4-14, Figure 4-15 and Figure 4-16 respectively. In Figure 4-14, detailed analysis of the harmonic component of the Ferrite magnet variant of the TFM is presented. It can be seen that there is a close match of the harmonic contribution between the FEA and the actual measured results. In Figure 4-14 (b), the measured back EMF is plotted with the actual measured signal and the filtered signal after removing the even harmonics. Since there was hardly any presence of even harmonics even at the measured signal, the two signals practically overlap. A closer look at the experimental back EMF reveals a small imbalance in the magnitude of the back EMF between the phases. In fact, the central phase (phase B) in the structure experience around 1.5% higher back EMF compared to the phases at top and bottom as seen in Figure 4-14(b). The percentage harmonic contribution in these phases, however, are in good correspondence with the simulation results and first few harmonic contributions for phase-B are shown in the bar chart in Figure 4-14(c). As seen from the harmonic orders, both simulation and experimental results show a presence of 3rd order harmonic of more than 2%. The higher order harmonic contributions are too small and at some point, falls beyond the tolerance level of the measurement system. Negligible presence of back-EMF harmonic establishes the robustness of the optimization process.
As mentioned in the previous section, the inherent structure of TFM and adaptation of high pole structure makes these motors susceptible to phase shift error between phases. This is visible in the back-EMF plot and harmonic analysis of the signal reveals a phase shift error of 2.4 degrees and 0.23 degree for the bottom and top phases respectively when the central phase is taken as reference. Measured phases for the back-EMF signals considering phase-B as reference are given in Table 4.3.

Figure 4-14: Back EMF profile for Ferrite magnet structure (a) Back EMF at 50 RPM from FEA (b) Experimental Back EMF at 50 RPM (c) Harmonic components as a percentage of fundamental.

The NdFeB magnet variant of the motor had significant THD in the back EMF even after optimization. The effect can be seen in both FEA and experimental results. Due to larger phase shift error, the first generation experimental TFM with NdFeB had large cogging torque making the speed control difficult for the dyno. To resolve the issue, the back-EMF data at variable speed was converted into a flux linkage curve which was later used to estimate the back-EMF at a
constant speed. This makes the approach more susceptible to measurement noise and leads to a larger harmonic component in the estimated back EMF. Estimated back-EMFs for the NdFeB rotor structure are given in Figure 4-15(b) where the noticeable presence of phase shift errors can be realized even with visual inspection. The dotted lines represent the back EMF measured after the removal of even harmonic components. The large discrepancy between the phases are mostly due to measurement errors and this makes the comparison of harmonic component ineffective. Just to have a coarse comparison, back EFM harmonic contributions from FEA and experimental contribution from phase-B is shown in the bar chart in Figure 4-15(c).

![Figure 4-15](image)

Figure 4-15: Back EMF profile for NdFeB Gen-1 magnet variant (a) Back EMF at 50 RPM from FEA (b) Experimental Back EMF at 50 RPM (c) Harmonic components as a percentage of fundamental.

The second NdFeB variant of the TFM proved to be mechanically robust and back EMF data of the second prototype is presented in Figure 4-16. Note that the larger air-gap also reduced the
expected harmonic component computed in FEA. The back EMF was measured at 300 RPM and found to be in close correspondence with the FEA results. The harmonic orders of the back EMF are given in Figure 4-16(c).

![Back EMF profile for NdFeB Gen-2 magnet variant](image)

(a) Back EMF at 50 RPM from FEA
(b) Experimental Back EMF at 300 RPM
(c) Harmonic components as a percentage of fundamental.

Figure 4-16: Back EMF profile for NdFeB Gen-2 magnet variant (a) Back EMF at 50 RPM from FEA (b) Experimental Back EMF at 300 RPM (c) Harmonic components as a percentage of fundamental.

The first harmonic component of the back EMF or flux linkage was compared for both the Ferrite and NdFeB structure and results are found to be in close correspondence with the FEA in spite of small phase imbalance. Though the actual torque producing capability is often over-estimated in FEA, the no-load condition results were found to be in close correspondence for both motors.
Table 4.2: Flux Linkage Constant for the prototype machines

<table>
<thead>
<tr>
<th></th>
<th>Ferrite</th>
<th>NdFeB Gen-1</th>
<th>NdFeB Gen-2</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>FEA</td>
<td>Experimental</td>
<td>FEA</td>
</tr>
<tr>
<td>Phase A</td>
<td>0.08055</td>
<td>0.0832</td>
<td>0.22</td>
</tr>
<tr>
<td>Phase B</td>
<td>0.0867</td>
<td></td>
<td>0.22</td>
</tr>
<tr>
<td>Phase C</td>
<td>0.0807</td>
<td></td>
<td>0.2203</td>
</tr>
</tbody>
</table>

Table 4.3: Phase Shift error in the prototype TFM structures

<table>
<thead>
<tr>
<th></th>
<th>Ferrite</th>
<th>NdFeB</th>
<th>NdFeB Gen 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phase A</td>
<td>122.4</td>
<td>131.9</td>
<td>118.2</td>
</tr>
<tr>
<td>Phase B</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Phase C</td>
<td>240.23</td>
<td>220.2</td>
<td>242.1</td>
</tr>
</tbody>
</table>

4.9.2 Cogging Torque

Since a multi-objective optimization was adopted and a good THD was achieved as shown in previous sections, the efficacy of the process can be truly realized by analyzing the cogging torque performance of the same machine. Unlike back-EMF, cogging torque of the machine relies on the contribution of three phases and hence an error in the phase shift can have a significant effect. Thus, the cogging torque performance will be analyzed first for an ideal case- which represents the achievable cogging torque if absolute precision can be maintained. Later, practical limitations will be introduced, and the effect will be observed.

FEA results for cogging torque associated with the Ferrite rotor structure is shown in Figure 4-17(a). As anticipated, with zero phase shift error, the cogging torque has only orders multiple of 6 as seen in the harmonic components of the cogging torque shown in the bar chart in Figure 4-17(b). Peak-to-peak value of the cogging torque is found to be only 25.4 mNm which is extremely small compared to the rated torque rating of the machine.

As the phase shift error can be measured from the back-EMF components, the same effect was introduced in the FEA and corresponding cogging torque is shown in Figure 4-18(a). As anticipated, even a small phase shift error introduced large second order harmonics in the machine.
cogging torque giving rise to a peak to peak cogging torque of 0.1332 Nm. Since this torque number is still small compared to the rated torque of the machine, it is a difficult task to accurately measure the cogging torque experimentally with the prototype. Such measured cogging torque is shown in Figure 4-18(b) where the blue band represents the cycle to cycle variation in the measured torque. The torque was measured for large enough cycle and a low pass filtering was used to cancel out the high-frequency cycle to cycle variations. Since a large part of the cycle to cycle variation comes from measurement noise, a large sample is expected to cancel out the effect and the red line in Figure 4-18(b) can be considered a good estimation of cogging torque in the machine. Peak-to-peak value of this measured cogging torque was found to be 0.0929 mNm.

![Cogging Torque Harmonics (Ferrite)](image)

Figure 4-17: FEA results for cogging torque for Ferrite magnet variant (zero phase error)
(a) Time domain signal from FEA (b) Harmonic components.

The NdFeB rotor structure was the first attempted prototype and various manufacturing issues made this structure prone to exhibit larger deviations from analytical and FEA based observations. The NdFeB grade used in the machine was much stronger compared to the Ferrite magnet and thus inherently have larger cogging torque. The cogging torque generated in this motor with no phase shift error is shown in Figure 4-19 (a) which has a peak to peak value of 0.5 Nm. The harmonic components of the cogging torque have similar contributions compared to the Ferrite magnet structure but with larger magnitude due to stronger magnets.
Figure 4-18: Cogging torque for Ferrite magnet variant with the presence of phase shift error (a) FEA (b) Experimental (c) Harmonic components comparison

When the FEA model was introduced with a similar phase shift as measured from back EMF, the large contribution of the second-order makes the peak to peak cogging torque exceed 16 Nm as...
shown in Figure 4-20 (a). The experimental motor has some additional structural issue that was not incorporated in the FEA model and hence shows a peak to peak cogging torque of 23 Nm.

![Graph](image1)

![Graph](image2)

![Graph](image3)

**Figure 4-20**: Cogging torque for NdFeB Gen-1 magnet variant with the presence of phase shift error (a) FEA (b) Experimental (c) Harmonic components comparison.

The second generation NdFeB rotor had significantly smaller phase shift error and thus smaller cogging torque compared to the first failed prototype. The results associated with this prototype are shown in Figure 4-21 and Figure 4-22 respectively. Figure 4-21 shows the expected cogging torque with a perfect phase shifted machine and the expected value can be seen to be lower than the first generation results due to increase in air-gap length. Figure 4-21 shows the experimentally measured cogging torque and similar procedure as described in the earlier section was followed for the measurement. The cogging torque measurement requires some significantly precise measurement capability and was not available in the test bench. Moreover, the IEEE standard for
measurement conditions could not be followed due to constraints associated with the dyno setup. Thus, the measured cogging torque is not guaranteed to be immune from measurement errors.

Figure 4-21: Cogging torque for NdFeB Gen-2 magnet variant (zero phase error) (a) Time domain signal from FEA (b) Harmonic components

Figure 4-22: Cogging torque for NdFeB Gen-2 magnet variant with the presence of phase shift error (a) FEA(b) Experimental (c) Harmonic components comparison
Chapter 5: Simulation Setup and Motor Model

In this chapter, an outline of the simulation environment, time domain approximate model, loss model and applicable control algorithm will be provided. This is necessary to get confidence over the finite element outcomes since these data will eventually play central role in the optimization process.

5.1 Simulation Setup

5.1.1 Finite Element Model

Ansys Electromagnetic Suite 17.1.0 was used as the primary design software throughout the TFM design process. Since 3D FEA is extremely time-consuming, necessary simplification exploiting the symmetry is required to perform the computation in a reasonable time. According to the previous discussion, the phase-phase gap between the adjacent phases was adjusted to reduce the mutual coupling between phases to a negligible extent. Thus, only one phase of the machine was simulated for all comparisons. Necessary post-processing can be used to come up with the complete three-phase motor performance. One pole of the motor model in the FEA software environment is shown in Figure 5-1. Exploiting the repetition, the boundaries were defined with appropriate magnetic field direction. A thin layer was defined to add higher mesh density along the air-gap region to increase simulation accuracy.

To control the meshing of the model, maximum allowable element size for different objects were defined to apply reasonable meshing of the 3D model. A finer mesh can be desirable for accurate computation but comes at a cost of immense simulation time. To mitigate the trade-off between simulation accuracy and computational burden, meshing was selected based on the type of simulation and corresponding impact in the optimization process. The initial crude optimization was performed employing relatively relax meshing of the model while the more mesh sensitive quantities such as cogging torque optimization was performed with finer mesh model. A typical meshing of the objects is shown in Figure 5-1(b).
5.1.2 Material Properties

The non-linear BH property of silicon steel is one of the dominant contributors in motor performance evaluation. The machine was aimed for high torque density operation and was pushed to the limit in terms of material utilization. Thus, a proper model of the BH characteristics is required. To incorporate proper reluctance model of the flux path, the stator was modeled as lamination stack with lamination placement along the circumferential direction. Due to the lamination stacking effect, the effective BH characteristics would get altered and must be modeled cautiously. If a stacking factor $F_S$ of 0.95 is assumed and ideally obtained flux density for corresponding MMF is denoted as $B_{solid}$, the effective permeability can be computed as

$$\mu_{solid} = \frac{B_{solid}}{H}$$ \hspace{1cm} 5-1

Now with 0.95 of stacking factor, the rest of the material can be considered as air and the air permeability of $\mu_o$ can be considered. Thus, effective flux density as a function of MMF can be denoted as

$$B_{SF} = H\mu_{solid} F_S + (1 - F_S)H\mu_o$$ \hspace{1cm} 5-2

The steel BH curve for solid steel and with 95% of stacking factor is shown in Figure 5-2.

---

Figure 5-1: Simulation model for one pole pair of TFM (a) Model with total solver domain and boundaries (b) Model meshing (Boundary removed for clarity)
5.2 Flux-Linkage Based Motor Model

Unlike conventional radial flux machines, transverse flux machines allow decoupling the individual phases by allowing appropriate separation between the phases. Though the separation was planned to avoid the imbalance in mutual coupling between the phases, this also allows adaptation of a simplified motor model for computation purpose. A typical motor can be expressed using the following differential equations

\[
\begin{bmatrix}
V_a \\
V_b \\
V_c
\end{bmatrix} = \begin{bmatrix}
R_a & 0 & 0 \\
0 & R_b & 0 \\
0 & 0 & R_c
\end{bmatrix} \begin{bmatrix}
i_a \\
i_b \\
i_c
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
M_{ba} & M_{ca} & i_a \\
M_{cb} & M_{cb} & i_b \\
M_{ca} & M_{cb} & i_c
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
\lambda_a \\
\lambda_b \\
\lambda_c
\end{bmatrix}
\]

5-3

The mutual coupling between phases makes the model more complicated and a simple table-based model ignoring the mutual effect can significantly suffer from inaccuracy. The TFM, on the other hand, can be complicated to control if the phases are mutually coupled since \( M_{ba}, M_{cb} \) and \( M_{ca} \) will not be necessarily equal. Since necessary separation between phases were maintained, these mutual inductances can be considered negligible and following motor model can be obtained

\[
\begin{bmatrix}
V_a \\
V_b \\
V_c
\end{bmatrix} = \begin{bmatrix}
R_a & 0 & 0 \\
0 & R_b & 0 \\
0 & 0 & R_c
\end{bmatrix} \begin{bmatrix}
i_a \\
i_b \\
i_c
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
L_a & 0 & 0 \\
0 & L_b & 0 \\
0 & 0 & L_c
\end{bmatrix} \begin{bmatrix}
i_a \\
i_b \\
i_c
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
\lambda_a \\
\lambda_b \\
\lambda_c
\end{bmatrix}
\]

5-4

This essentially decouple the phases and for a single phase we can assume

\[
V_a = R_a i_a + L_a \frac{di_a}{dt} + \frac{d}{dt} \lambda_a
\]

5-5
Consequently, if a simple table for flux and torque is generated as a function of phase current and rotor position, the complete machine can be simulated. Decoupling of phases enables using a simple 2D table unlike a 4D table otherwise required which increases the complexity of computation and table size in geometric proportion. Moreover, the symmetric nature of the machine further reduces the computation and table size since only half of the electrical cycle is sufficient for full machine performance evaluation.

![Figure 5-3](image)

**Figure 5-3:** Sample model data and results for flux linkage based single phase motor model (a) DC excitation (b) Resultant flux linkage for DC excitation (c) DC excitation electromagnetic torque (d) Three phase torque for sinusoidal excitation.

To generate the table, only half electrical cycle of single phase machine was simulated with DC excitation at the stator. The different DC excitation levels are shown in Figure 5-3(a) where the DC excitation currents are shown with dotted lines. Though only 7 excitation levels are shown in the figure, the actual model was generated with 0.5A interval of DC excitation variations. The corresponding flux linkage and generated torques are plotted with dotted lines in Figure 5-3(b) and (c) respectively. The symmetry around the rotor $d$-axis or 180° rotor position is evident from the figures. Once the two tables for flux linkage and generated torque is produced, generated flux and electromagnetic torque for any current wave-shape can be generated using linear interpolation.
Such sinusoidal current excitation is shown in Figure 5-3(a) with a thick line and for each point of the excitation, corresponding flux linkage and electromagnetic torque was computed and plotted with thick solid lines in the figures. For the other two phases, appropriate phase shifting in the table should be considered and similar flux and torque quantity can be reproduced.

To further improve the model accuracy, magnet temperature dependence was incorporated in the model. The full simulation was done at two different magnet temperatures and depending on the desired magnet temperature, a weighted average of the torque and flux profiles were used to compute motor performance. A brief block diagram of the model is shown in Figure 5-4.

5.3 Preliminary Simulation Results and Motor Control Basic

The simple flux-linkage based machine model allows motor performance evaluation for any arbitrary excitation as seen from the previous example. This indeed can come handy in evaluating motor performance with a specific control algorithm.

The most popular standard for controlling a permanent magnet synchronous machine is to use vector control in the synchronous reference frame. The orthogonal axes known as direct axis or $d$ – axis and quadrature axis or $q$ – axis are chosen in a manner that the back EMF of the machine is perfectly aligned with the $q$ – axis making only the $q$ – axis current responsible for torque production with machines without saliency. Ignoring the derivative terms, the ideal machine model in the $dq$ domain can be represented as

$$
\begin{bmatrix}
    v_q \\
    v_d
\end{bmatrix} =
\begin{bmatrix}
    R & \omega L_d \\
    -\omega e & R
\end{bmatrix}
\begin{bmatrix}
    i_q \\
    i_d
\end{bmatrix} +
\begin{bmatrix}
    \omega e \lambda_M \\
    0
\end{bmatrix}
$$

Figure 5-4: Temperature dependent motor simulation model.
Torque = \(\frac{3}{2}P\lambda_M i_q\)

Here, \(L_d\) and \(L_q\) are the \(d\) and \(q\) axis inductance and \(\lambda_M\) is the flux linkage constant and \(\omega_e\) is the electrical angular velocity. It should be noted that, the ideal PMSM model does not consider the core saturation and associated change in inductances. Though the TFM was built with surface mount magnets with no anticipated reluctance torque, saturation in the core resulted in some inductance variation and caused the motor to exhibit some reluctance torque.

![Graphs of Torque vs. Direct and Quadrature Axis Currents](image)

**Figure 5-5**: NdFeB TFM parameters as a function of direct and quadrature axis currents (a) Electromagnetic torque (b) Line to neutral RMS voltage at 100 RPM at 100°C coil temperature (c) \(d\) – axis inductance \(L_d\) (d) \(q\) – axis inductance \(L_q\).

The electromagnetic torque generated as a function of \(d\) and \(q\) axis currents are shown in Figure 5-5 (a) for NdFeB magnet and Figure 5-6(a) for Ferrite magnet structure. It can be seen that the
constant torque lines are not parallel to the horizontal axis indicating some reluctance torque component. The conventional motor control algorithm suggests using the smallest current vector to achieve certain torque command to minimize the copper loss also called the maximum torque per ampere control (MTPA). The MTPA lines in the figures are shown in cyan dotted lines and the corresponding $i_q$ and $i_d$ values represent the optimum current vectors for the corresponding torque command. The MTPA line along the q-axis represents a motor with zero saliency and no reluctance torque output. If the MTPA lines for the NdFeB rotor is compared with that of Ferrite rotor, the inclination towards a larger $d$-axis current indicates larger saturation for the NdFeB rotor. The $d$- and $q$-axis inductances are also presented as a function of $i_d$ and $i_q$ in the figures.

The RMS line to neutral voltage at a speed of 100 RPM is shown in Figure 5-5 (b) for NdFeB magnet and Figure 5-6(b) for Ferrite magnet respectively. It is evident from the plot that the voltage reaches zero for $i_d = 4.5A$ and $i_d = 1.6A$ for NdFeB and Ferrite rotor respectively. This closely match with the characteristic current described during the demagnetization analysis presented in Chapter 3.

The RMS voltage plots shown in the figures are generated for a fixed mechanical speed of 100 RPM and will scale up at higher operating speeds. Due to the limit in the maximum applicable voltage for a given DC bus voltage, at higher speed, some of these voltages will exceed the limit set by this inverter constraint. At these points, the available operating points in the plots will get smaller and will tend to shrink towards the center of the voltage limit ellipses scene in the figures. Thus, for a given torque command, the MTPA operating points shown with the cyan line will not be applicable after the speed exceeds a certain value often denoted as corner speed. Beyond corner speed, the MTPA line will shift left and new current vectors must be chosen for different torque commands.
Figure 5-6: Ferrite TFM parameters as a function of direct and quadrature axis currents (a) Electromagnetic torque (b) Line to neutral RMS voltage at 100 RPM at 100°C coil temperature (c) $d-$axis inductance $L_d$ (d) $q-$axis inductance $L_q$.

Typical variations in available torque output as speed increases are demonstrated in Figure 5-7 for two different speeds. The RMS voltage was calculated for speeds of 250 RPM and 450 RPM and the voltage vectors which exceed the inverter output limits were eliminated. To come up with the inverter voltage limit, a 250 V DC bus with space vector modulation scheme was considered. It should be noted that the ideal torque output of an EM machine is independent of operating speed and only depends on the current vector. The available torque decreases as the applicable current vector shrinks as less and less current vectors fall within the available voltage limit region as speed increases. If the ellipses forming the voltage boundaries are observed, it is found to be following a closer to ideal pattern (not tilted) for higher speed. This is due to the fact that, at higher speeds,
the resistive drop becomes negligible and the boundaries are decided more by the flux linkage terms. This is evident from the process that at higher speed, available torque output will decrease, and the current vectors will have more \( d \)-axis components for same torque yield. This control scheme is widely known as maximum torque per ampere with flux weakening control and is the most common motor control scheme.

Figure 5-7: Torque and voltage magnitude as a function of current vectors for different operating speeds. Voltage larger than the inverter limits have been eliminated. (a) Available torque output at 250 RPM (b) RMS line to neutral voltage at 250 RPM (c) Available torque output at 450 RPM (d) RMS line to neutral voltage at 450 RPM.

The Torque envelopes as a function of speed for the two motors are presented in Figure 5-8. As the characteristics currents fall within the rated current range, both motors are suitable for high-speed operation given that the mechanical limitation of the structure is not reached.
Losses in electric machines can be broadly separated into three major components

1) Copper loss or joule loss in the winding due to current excitation.
2) Magnetic loss components e.g. core losses in the lamination material, which includes Hysteresis loss in the steel lamination, Eddy current loss both in laminations, magnets or support structure. Some of these losses are also referred as stray load loss.
3) Mechanical loss components such as friction and windage losses

Joule losses are the simplest of all loss components and can be calculated accurately from the winding geometry and excitation current at given temperature. The most critical of all loss components can be attributed to the magnetic losses. Though there have been standardized test methods to assess the loss characterization of the lamination material called Epstein frame test, these test results are often performed at limited frequencies of 50/60Hz sinusoidal excitation at 1.0/1.5T peak magnetic flux density [37]. Moreover, the conventional test setups are often not capable of measuring lamination characteristics at high frequency and high flux density, the loss data for a variable speed drive system is often not obtainable from the manufacturers’ datasheet. However, if the steel characterization is performed over a wide frequency and flux density range, a typical magnetic loss resembles the plot shown in Figure 5-9.
Once provided with the electric steel loss characterization, a precise model of the loss properties is required to accurately predict the losses in FEA. Different approaches have been executed in perfecting the method [38][39] among which the conventional three-term model is used by the FEA tool used in the design process. In this conventional loss model, the total magnetic loss in the electric steel is represented by the following loss terms

\[ P_h = K_h f B^2 \]  
\[ P_c = K_c f^2 B^2 \]  
\[ P_e = K_e f^{1.5} B^{1.5} \]  
\[ P_{core} = P_h + P_c + P_e \]

where \( P_h \), \( P_c \) and \( P_e \) are the hysteresis, eddy current and excess loss components of the electric steel. Total magnetic loss in the lamination material is the sum of these three loss components. The measured loss through Epstein frame test is approximated using these three estimation equations and the coefficients \( K_h \), \( K_c \) and \( K_e \) are tuned to minimize the estimation error. The dotted lines in Figure 5-9 shows the steel losses represented through the three-term iron loss model. For NdFeB magnet models, the electrical conductivity in the magnet also contribute for eddy current losses inside the rotor magnets. This loss is also considered in Ansys Electromagnetic suites and the relationship between the frequency and the loss component closely follow the typical square
relationship similar to steel eddy current loss. The benefit of this loss model can be credited to the separation of losses based on their correlation with operating frequency. Only one fixed frequency simulation and the loss maps for different current vectors are adequate for computing magnetic losses at any frequencies. It should be noted that, in the actual machine, high eddy current losses in the machine can produce a negative electromagnetic torque which eventually reduces the torque throughput of the machine at a higher speed for same current vector excitation. But if the typical way of loss computation depends on the post-processing of magnetic field in the medium and hence the FEA tool yield same torque output for a fixed current excitation irrespective of speed and losses. Incorporating the magnetic losses in field calculation requires an iterative process and can become overwhelmingly computational intensive especially for a 3D FEA model.

Computed magnetic losses for NdFeB-TFM as a function of current vectors are shown in Figure 5-10. The losses shown are calculated for the mechanical speed of 400 RPM based on the M19 loss property shown in Figure 5-9 and the magnet eddy current loss was computed based on magnet material conductivity. The total magnetic losses in the machine are simply the sum of all these components as shown in Figure 5-11(a). Since the flux density in the machine is independent of operating speed, the loss relationship between two different speeds depends only on the corresponding operating frequencies. Thus, loss components for any operating frequencies can be simply interpolated using the three-term loss equations. Total loss at 600 RPM computed using the approach is shown in Figure 5-11(b).
Figure 5-10: Magnetic loss components in NdFeB-TFM at 400 RPM computed using FEA (a) Hysteresis loss (b) Eddy current loss (c) Excess loss (d) Magnet eddy current loss.

Figure 5-11: Total magnetic loss for NdFeB-TFM (a) 400 RPM (b) 600 RPM.
A similar approach was followed for the Ferrite-TFM and corresponding loss components computed for 400 RPM are shown in Figure 5-12. It should be noted that Ferrite magnets have close to zero conductivity and thus do not account for eddy current loss inside the magnet. If the total loss components are compared, both NdFeB and Ferrite based motor account for minimum losses at the region close to the characteristic current. This justifies the loss contour plot since this current represent total cancelation of magnetic flux and consequently should minimize the magnetic loss component.

![Figure 5-12: Magnetic loss components in Ferrite-TFM at 400 RPM computed through FEA](image)

(a) Hysteresis loss (b) Eddy current loss (c) Excess loss (d) Total magnetic loss.

To find the loss component at any torque-speed operating point, the corresponding current vector can be computed based on the available bus voltage as explained in the previous section. Once the current vector is determined, the associated magnetic loss component can be computed from the
pre-calculated table and necessary adjustment can be introduced based on the operating frequency. This simplifies the process and saves substantial computation time compared to simulating the at different torque-speed conditions with associated current vectors. So far by the measured loss through Epstein frame test and the curve fitting based approximate model gives an impression of quite a straight-forward approach for estimating magnetic losses in rotating machines. Though in practical machine construction, lamination stamping process, stator core bonding, and other manufacturing issues create a significant deformation in the steel property causing large deviation in the practical steel core loss compared to the Epstein frame test. Moreover, the difficulty in separating the loss components adds further difficulties in truly identifying the responsible contributor to the machine loss. The actual motor loss often surpasses the loss computed from FEA by multiple factors.
Chapter 6: Motor Construction

The unique ring-shaped winding pattern of transverse flux topology introduces a substantial simplification in the winding process. But the ring winding also necessitates an unconventional stator flux path and associated complexity. The use of SMC material can provide 3D flux path using a single piece core, but conventional steel lamination core imposes some restriction in that and requires an intricate stator support structure. This in terms increases the manufacturing cost and acts as one of the prime factors behind the unacceptance in commercial applications.

As part of the research, the emphasis was given on the simplification of the mechanical structure without sacrificing structural robustness. This chapter will describe process related to the support structure design, fabrication, and assembly of the full machine.

6.1 Support Structure Design

6.1.1 Stator Supporting Structure

Due to torque-pole linearity, TFMs are the desired topology for high torque application and high pole count machines are mostly anticipated. High pole count requires a large number of stator pole pieces to be assembled precisely. The designed support structure to hold the stator pole is shown in Figure 6-1(a). The stator contains six similar plates to hold the stator cores with appropriate spacing. The stator core was incorporated with the extension at the back end to provide mechanical holding force as shown in Figure 6-1(b). Though high strength epoxy was used to attach the poles rigidly with the supporting plate, the extension at the back will ensure mechanical holding force in case of failure of the epoxy holding force due to chemical degradation due to aging and thermal issues. The stator pole supporting plates shown in Figure 6-1(a) is shown from another perceptive in Figure 6-1(c). Each of these supporting plates has grooves for precise positioning of the stator poles as seen from Figure 6-1(a). The inner sections of the plates have twelve axial channels. Nine of these are to permit the winding ends to pass axially through the bottom phase. The holes to direct the winding ends have additional openings in the circumferential direction for the ring winding ends to pass through.

The intermediate holes between the sets of winding directing channels are drilled to allow long steel screws to provide mechanical thrust for holding the stator structures together. The topmost plates have appropriate threading to allow the screws to apply axial thrust. The construction of the stator will require the stator poles to adhere to the supporting plates with the help of a strong
adhesive. Once each plate has the full stator poles attached to it, ring winding can be wound. The complementary stator of the matching phase can be constructed in similar fashion. The final look of a single-phase stator without the winding is shown in Figure 6-1(d).

Figure 6-1: (a) Aluminum support structure for stator poles (b) One single stator pole (c) Aluminum support structure bottom view (d) One complete phase with stator poles with no winding inserted.

6.1.2 Phase Shifting Implementation

The pole count of the designed TFM was set as 32 which is significantly higher than typical radial flux machines. This high pole count corresponds to a scaling factor of 16 for any conversion between a mechanical angular position to an electrical one. Thus, a minuscule error in angular phase shift between the phases can lead to a large error in the electrical properties of the machine.
To precisely control the phase shift between the phases, the stator supporting plates were incorporated with precisely located dowel pins as seen in Figure 6-1(a) and (c). These pins ensure precise angular shift between the axially stacked phases if properly installed.

6.1.3 Bearing and Rotor Support

The stator support structure should also be responsible for housing the motor bearings with adequate holding strength. To support the bearings, the bottom and top stator support plates were constructed with appropriate housings as shown in Figure 6-2(a). The bearings are shown in red in the figure placed in the housing. Two bearings at the top and bottom of the stator give sufficient support to the rotor to maintain the air-gap. Figure 6-2(a) also shows the complete cross-section of the motor and gives a complete impression of the motor structure. The axial channel at the left with openings towards the stator poles can direct the winding towards the bottom plate and the structure has nine similar channels. The channel at the right, on the other hand, has a threaded end and is responsible for creating the axial thrust needed for structural robustness.

The structure of the rotor is rather straightforward as seen in Figure 6-2(b). The rotor structure used in the motor is comparable with simple outer rotor surface magnet rotor of conventional radial flux machines. Though these rotors are usually constructed with the top plate being shrink-fitted with the external shell, the first prototype of such nature experienced issues with the shrink fitting and failed to provide robustness against the strong magnetic force. Rather unconventional design shown in Figure 6-2(b) was thus implemented where the whole rotor structure was made from a single magnetic steel piece to avoid issues with shrink fitting of the top plate. The inner shaft was rather shrink fitted as shown in the figure.

Though the usual way of attaching magnets to a surface mount rotor is to have appropriate grooves in the rotor for the guidance for placement, an alternative cost-effective mean was used for the prototyping. A 3D printed structure as shown in Figure 6-2(c) was used inside the rotor to create a rectangular grid for the 192 magnets of the 32-pole machine. The outer rotor structure removed the risk of magnet detachment due to centrifugal force and the simple epoxy-based adhesive was sufficient to provide sufficient adhesion.
6.2 Thermal Modeling

The lifetime of a motor is a strong function of the operating temperature and a close assessment of steady-state operating temperature is a prerequisite before confirming the machine performance characteristics for continuous operation. During the design stage of an electric machine, thermal FEA can provide an estimate of the operating temperatures. In this section, a brief description of the thermal FEA setup procedures and associated assumptions will be described.

6.2.1 Copper Winding Modeling

Magnetic wires used in motors are made of copper with a thin layer of electrically insulating material. Though copper is a very good conductor of heat, the necessary insulation significantly
reduces the thermal conductivity through the wire cross-section. Moreover, the air-gap between the adjacent conductors adds additional thermal insulation making it difficult for heat to travel. However, along the axial direction of the wire, the pure copper provides a conductive path for heat making the wire to have an anisotropic thermal conductivity. To introduce such anisotropic behavior, a simplified single conductor model as shown in Figure 6-3(b) was used. The actual coil formation with several single wires was substituted by a single coil with insulation with the equivalent cross-sectional area.

Figure 6-3: Coil conductor in the motor (a) Actual coil formed with insulated wires (b) Equivalent single coil representation.

### 6.2.2 Periodicity, Heat Sources, and Convections

A simplified motor structure with periodic symmetry was used in the thermal FEA to reduce model size. Ansys transient thermal simulation model was prepared as shown in Figure 6-4(a). Copper loss was modeled as heat sources with equivalent volumetric heat density inside the copper conductors. The equivalent circuit-based model shows the heat generation as equivalent current sources. The convections in the air-gap regions and at stator and rotor outer surfaces were modeled using empirically calculated values. As seen from the circuit equivalent model in Figure 6-4(b), the heat flow paths are largely dependent on the convections occurring in different regions. The thermal conductivity of metal is significantly high and thus the surfaces responsible for the heat convections work as the choking points for the internally generated heat. Moreover, the metal-metal contacts have some thermal resistivity at the contact region which depends on the surface finish, clamping force and other factors creating additional thermal resistance along the axial direction of the motor. This should result in a temperature gradient between the top and bottom
phase and corresponding imbalance in winding resistances. However, these contact resistances are difficult to compute using FEA and were ignored in the simulation. The absence of these thermal resistances was compensated by the modification of the convection at the outer surface.

6.2.3 Thermal FEA Results

Temperature plot of the motor model for rated current operation is shown in Figure 6-5(a). Since the contacts between the phases were assumed as perfect conduction paths of heat, the temperature gradient can be seen absent in the stator. Thus, the stator maximum temperature will be very close to the average stator temperature. A larger temperature gradient can be seen in rotor due to the geometry and heat path as seen in Figure 6-5(b). A thermal image of the rotor during experimental heat run is shown in Figure 6-5(d) and a similar trend of local hot spots at the central shaft can be verified from experimental results as well. The magnet temperatures are shown in Figure 6-5(c) and can be found within the thermal limit of the magnet grade.
Figure 6-5: Thermal FEA temperature map and experimental thermal camera image (a) Periodic motor model temperature (b) Temperature map for rotor (c) Temperature map for magnets (d) Experimentally captured thermal camera image.

Transient temperature readings obtained from thermal FEA are shown in Figure 6-6. The coil inside the holds the hottest part of the machine and thus maximum stator temperature can be considered the coil temperature. For rotor, the average temperature represents a closer approximation of the magnet temperature evident from the temperature plots.
6.3 Prototype Construction and Test Setup

As mentioned earlier, other than the lamination stacks and magnets, the rest of the structural components were constructed in the university facility. Step by step procedure of the stator construction is presented in Figure 6-7. Figure 6-7(a) shows the first stage of the construction process where a simple aluminum disc was machined with a lathe machine. Appropriate grooves to support the stator pole pieces were machined with CNC drilling and partly finished structure is shown in Figure 6-7(b). The shape was introduced with necessary drilling later to create the channels for conductors and bolts. Figure 6-7(c) shows one half of a single-phase stator with the stator pole attached with epoxy adhesive. High-temperature resistant Kapton tapes were used as insulation between the wire and the stator poles. Additional insulations were added at the wire ends to provide extra support near the sharp bending of wires and particularly in the regions where the wires touch the sharp edges of the stator cores. One phase of a completed stator is shown in Figure 6-7(d). The nine channels numbered in the plate can be seen in the photo used for guiding the winding ends. Two completed single-phase stators are shown in Figure 6-7(e). A closer inspection reveals the presence of small dowel pins used for phase alignment in these structures. Finally, a completed 3-phase stator mounted on the base plate is shown in Figure 6-7(f).
Despite having the conventional rotor structure, major difficulties were experienced during the initial construction attempts. Motor manufacturing is a multidisciplinary division and requires expertise both in the electrical and mechanical domain. The shrink fitting system at the rotor top plate proved to be challenging due to lower tolerance in such design. Figure 6-8 (a) shows the first attempted rotor structure which eventually failed to maintain structural robustness in the presence of strong magnetic pull force excreted by the rare earth magnets. Though some preliminary data
were extracted with this first prototype and the back EMF, cogging torque and flux linkage profile proved to be in close correspondence with the FEA results.

The issues experienced in the first rotor were addressed and a new simple rotor structure was designed for the second prototype. Since there was no significant issue with the apparently more complicated stator structure, the same stator was used in the process. To reduce the risk of failure, the air-gap in the second prototype was raised to 1.5 mm and a weaker Ferrite magnet was used for added safety. Figure 6-8 (b) shows the constructed Ferrite magnet-based rotor. This rotor exhibited no mechanical issue and extensive testing was performed on this prototype. Successful operation of the second rotor prototype with Ferrite magnet provided confidence in the structural integrity of the rotor design and the third rotor with initially planned NdFeB magnet was constructed. To further decrease the risk of failure, air-gap length was increased to 1.6 mm for the third prototype. Figure 6-8 (c) shows the NdFeB based rotor and extensive tests were performed with this rotor structure as well.

Figure 6-8: (a) First NdFeB variant prototype (b) Ferrite Rotor (c) NdFeB Gen-2 rotor.

6.4 Experimental Setup and Results

Figure 6-9(a) shows the test bench used to run the experiments. A 10 HP induction motor driven by an ABB ACS-800 four quadrant drive was used as a load motor. Yokogawa WT-3000 power analyzer was used for the power measurement. The motor was incorporated with 3 thermocouples and the corresponding temperatures were measured during the measurement of efficiency. Agilent
A 3472A data acquisition unit was used to capture the temperature data in synchronism with the power analyzer output. A Himmelstein torque transducer rated for a maximum torque of 56.5 Nm (500 lb-inch) was used to measure the output torque. The direct method of output power measurement was used and thus care was given in calibrating the offset of the torque transducer. The oscilloscope was used primarily to capture instantaneous torque, voltages and position sensor output for detailed harmonic analysis of some selected data.

Since the no-load performance such as cogging torque and back EMF data have been provided in the previous chapter, the motor performance under current loading will be presented in this chapter. The typical motor tests such as heat run, retardation tests, and final loss map will be presented as well in later part of this chapter.

![Motor Test Bench](image)

**Figure 6-9: Motor Test Bench.**

### 6.4.1 Retardation Test

One of the most common tests for permanent magnet motors is to measure the no-load loss in the machine. The test was performed according to the guideline provided in IEEE 1812-2014 which give a decent idea about the motor no-load loss. The test relies on the fact that under a no-load
condition, without any external force, a running rotor will decelerate, and the rate of deceleration will depend on the no-load magnetic loss component for the given speed. In other words, the magnetic loss components in the machine will create a drag and bring the rotor to halt. Thus, the test gives the summation of magnetic loss and friction and windage loss. With a similar unmagnetized rotor, the friction loss can be separated for further analysis. In this experiment, the unmagnetized rotor was not available and the friction loss separation was not performed.

For an unloaded rotor,

\[ T_{\text{dec}}(\omega_m) = J_c \frac{d\omega_m}{dt} \]

were \( T_{\text{dec}} \) is the deceleration torque at mechanical speed \( \omega_m \) and \( J_c \) is the rotor inertia.

The loss at speed \( \omega_m \) can be computed in a straight-forward way as

\[ W_{\text{Loss}}(\omega_m) = \omega_m T_{\text{dec}}(\omega_m) \]

To calculate the loss at a speed of \( \omega_m \), the rotor must be spun at a speed higher than this speed and completely disconnected from the drive circuit to ensure zero current operation. The speed of the motor should be measured throughout the process to compute the deceleration accurately. A screen shot of oscilloscope data of a sample test result is shown in Figure 6-10. To increase the precision in speed and hence deceleration measurement, the instantaneous speed was measured in the DSP controller and plotted using the DAC output as shown with a green line in the oscilloscope screenshot. The magenta plot represents the mechanical position reading from the encoder and the yellow plot shows the measured current of phase A. These data were also captured in .csv format and post processing techniques were used to precisely compute the no-load loss.

Figure 6-10: Retardation test oscilloscope data sample screen shot.
Actual test data for the retardation test with Ferrite rotor structure is given in Figure 6-11. Figure 6-11(a) shows the phase current during the test and can be seen to have sudden termination indicating the opening of the three-phase switch. The motor controller DSP was kept running and the measured mechanical position is shown in Figure 6-11(b). The calculated speed and the polynomial representation of the decelerating segment is given in Figure 6-11(c). The polynomial fitting of the speed data is recommended by IEEE Std. 1812 to get a smooth deceleration pattern. Figure 6-11(d) shows the deceleration torque $T_{dec}$. It should be noted that the initial segment of this torque should be discarded due to the initial error in the speed measurement due to filter delay. This can be avoided by selecting the decelerating region from a lower mechanical speed. $T_{dec}$ as a function of speed is shown in Figure 6-11(e). Finally, Figure 6-11(f) shows the no-load loss measured using the retardation test. The final drop in the loss towards higher speed represents the initial erroneous data due to incorrect speed measurement and should be discarded.

The procedure was repeated for the NdFeB rotor as well and the results are given in Figure 6-12. The no-load loss for this stronger magnet TFM can be found to be several times compared to the Ferrite rotor variant. This is expected since the steel core flux density is significantly higher for the NdFeB rotor and hence higher core loss is evident.

A different approach for measuring the no-load loss was undertaken to confirm the measurement accuracy. With the motor terminal open, the motor was spun at various speed with the help of the load motor. This was possible as the load motor was driven by a four-quadrant drive and thus can operate in both motoring and generating mode in any direction. The torque transducer was used to measure the shaft torque. With the open circuit motor, the shaft is expected to experience the drag torque produced by the magnetic loss components and this torque must be supplied by the load motor. Since the drag torque experienced at any speed corresponds to the torque associated with the magnetic and friction loss at no-load, this drag torque and the deceleration torque $T_{dec}$ measured from the retardation test are essentially the same quantity. By multiplying the operating speed with the measured drag torque $T_{Drag}$, the no-load loss at any operating speed can be calculated.
Figure 6-11: Retardation test results for Ferrite rotor TFM (a) Phase Current (b) Mechanical position reading (c) Measured speed and polynomial fit (d) Deceleration torque (e) Deceleration torque as a function of speed (d) Computed no-load loss
Figure 6-12: Retardation test results for NdFeB rotor TFM (a) Phase Current (b) Mechanical position reading (c) Measured speed and polynomial fit (d) Deceleration torque (e) Deceleration torque as a function of speed (d) Computed no-load loss

Figure 6-13 (a) and (c) shows the drag torque measured for Ferrite and NdFeB rotor respectively for different rotational speed. The process was repeated for both clockwise and counter clockwise
rotation. The computed loss components as a function of speed are given in Figure 6-13 (c) and (d) respectively. The loss measured from the retardation test are also provided in the same plot for comparison.

![Figure 6-13](image)

Figure 6-13: No load drag-torque measurement and associated no-load loss. (a) Drag-torque for Ferrite-TFM (b) Drag-torque for NdFeB-TFM (c) No load loss for Ferrite TFM (d) No-load loss for NdFeB-TFM.

### 6.4.2 Low-Speed Torque with MTPA Control

As seen in earlier sections, transverse flux machines with faulty design choice can lead to a machine with extra-ordinary no-load performance such as very high back EMF despite being worthless with actual stator excitation. Thus, it is necessary to test the motor torque with maximum current rating and test torque producing capability. Both motors were excited with different current excitations at low speed for both clockwise and counter-clockwise direction. The speed was kept low for two main reasons – i) To reduce current control error and maintain as close as possible to
ideal current excitation. This is important as perfect current control system was assumed during the FEA. ii) The larger speed also creates larger drag torque due to magnetic losses in the system. This drag torque is also present at low speed and the torque measured at the shaft does not truly represent the electromagnetic torque produced in the motor. As seen in the previous section, the motor exhibits a drag torque with zero current at the winding, the torque measured at the shaft with current excitation is not a true representation of the generated torque. A fraction of the generated electromagnetic torque is used to overcome the friction and drag torque for magnetic loss components. This does not necessarily justify adding the drag torque measured in the previous method with the measured torque to compensate the loss in torque. With current loading, the flux distribution inside the machine will be different and thus, the no-load drag torque does not represent the actual drag torque in the loaded condition. Essentially, there is no correct method to compensate this in the measurement and the shaft output torque will be considered as the electromagnetic torque component for comparison purpose.

![Figure 6-14: Low-speed torque with MTPA current both in clockwise and counterclockwise direction (a) Ferrite Rotor (b) NdFeB Rotor.](image)

The measured torques at the motor shaft for different current excitations are provided in Figure 6-14. For both Ferrite and NdFeB rotor, good correspondence is found when compared to the FEA data. The current vectors in the experiment and FEA were chosen for maximum torque per ampere (MTPA) operation. Since FEA completely ignores the loss contribution in torque output, the close correspondence in mean torque measurement represents slightly higher torque production in the experimental motor. This can be attributed to many factors such as the miss-match in $BH$ property.
in the experimental motor compared to the FEA computation. The true stacking factor of the stator cores was approximated and can differ slightly.

Figure 6-15: Instantaneous torque at 15 RPM for Ferrite rotor structure (a) 7.6 Nm output torque (b) 6 Nm torque output (c) 1 Nm torque output
A time domain representation of torque for three different torque levels are shown in Figure 6-15. Comparing the torque oscillation in the three plots, it is obvious that the measurement system loses accuracy in small ripple measurement at higher mean torque level. At lower torque value of 1 Nm, the analog signal representing the torque value was small enough and the scaling was appropriate for the ripple measurement. Moreover, the required constant speed operation cannot be achieved in the used dyno at higher torque levels. The induction motor used for maintaining speed exhibits speed oscillation giving rise to inertial torque variation. Thus, the measured torque ripple at higher torque magnitude is not reliable and should be discarded.

![Figure 6-16: Torque Speed envelope and motor power](image)

(a) Torque-Speed for Ferrite TFM (b) Power for Ferrite-TFM (c) Torque-Speed for NdFeB TFM (d) Power for NdFeB-TFM.

### 6.4.3 Torque Speed Envelope

The torque-speed envelope was computed with 250V DC bus for both motors. Though the initial torque measured for mean torque comparison was done at low magnet temperature, the rotor was
significantly warm during the data extraction to compute the envelope. Thus, the torque yield for similar current excitation reduced and a smaller envelope was achieved compared to low-temperature envelope achievable otherwise. The torque speed and power envelope for Ferrite TFM is shown in Figure 6-16(a) and (b) respectively. Similar curves for the NdFeB TFM are given in Figure 6-16(c) and (d) respectively. The corner speed and thus envelope depend on the voltage limit of the inverter and is a strong function of modulation index. The inverter was operated with space vector pulse width modulation (SV-PWM) and is the most widely used scheme in inverter operation. Due to the presence of dead-band, the actual voltage limit of the inverter is slightly lower than the ideally calculated magnitude. Consequent adjustment of the modulation index was done in the simulation model to correspond to the similar voltage limit.

6.4.4 Heat Run and Winding Temperatures

To validate the motor performance as rated value, the motor must reach the thermal steady state with the declared rated values without exceeding the recommended maximum limit of the materials. The most crucial failure at high temperature can be failure of winding insulation due to thermal degradation, maximum bearing, magnet and epoxy temperature ratings when applicable. The IEEE Std. 112 [40] recommended criterion for the achievement of thermal steady state suggests a temperature rise of less than 1°C in a 30 minutes time span.

To perform the tests, the motor was equipped with three thermocouples for three different phases. For phase B and C, the thermocouples were placed directly on the winding with thermally conductive paste to increase heat transfer between the coil and the thermocouple terminals. For phase A, the thermocouple was placed at one stator pole as shown in Figure 6-17. Two additional thermocouples were used- one to measure the temperature of the motor mounting plate and the other one to measure the ambient temperature.
Heat run test data for the Ferrite-TFM is presented in Figure 6-18. The temperature readings for different locations are numbered one to five. During the heat run test for Ferrite TFM, the motor was stopped in a regular interval without stopping the stator current and the thermocouple assigned for ambient temperature was connected to the rotor outer surface. This attempt was made to get some approximate idea about the rotor temperature as other suggested methods were not available. Due to outer rotor structure, the rotor exhibit much lower temperature compared to the stator and the primary heat path can be seen to be through the axial direction. This is evident from the temperature readings for phase-A, B, and C. Though the thermocouple for phase-A is much lower from the other two phases due to the fact that the thermocouple for phase-A was rather placed in stator pole, a definite gradient in the coil temperature can be observed as well and will be explained later. Figure 6-18(b) shows the gradual decrease in torque with rising temperature with a sharp increase at the end due to the decrease in operating speed and consequent reduction in friction and drag torque.
Figure 6-18: Experimental test data for Ferrite-TFM heat run test (a) Measured temperatures (b) Shaft torque (c) Mechanical speed (d) Efficiency.

Similar plots for the NdFeB-TFM are given in Figure 6-19. Both motors exhibit similar trend in terms of the drop in torque and temperature rise. The NdFeB-TFM though reaches higher temperature with similar current excitation levels due to the presence of significantly higher magnetic loss components. Both motors were excited with 4A peak current magnitude which represents an RMS current density of $5.66 \, A_{RMS}/mm^2$. Though the permitted maximum operating temperature of the winding insulation was much higher (200 °C), the steady state current was limited to avoid failure due to local hot points.
Moreover, the temperature readings of the attached thermocouples does not necessarily represent effective coil temperature due to thermal contact resistance between the thermocouple terminal and the winding. Different thermocouples in different winding locations can give a variety of results and an effective procedure is required to establish a mapping between the thermocouple readings and actual winding temperature. The recommended method to get a true estimation of winding temperature according to IEEE Std. 112 [40] suggests using the ‘Rise by Resistance’ method which essentially the measurement of winding resistance at the end of the heat run to compute the temperature. The procedure gets complicated for a structure with temperature imbalance between phases as the line-to-line resistance and corresponding winding temperature can lead to confusion in the presence of temperature imbalance. The temperature imbalance in the winding resistance and hence temperature can be realized if the RMS voltage of the phases during
the heat run test is observed in Figure 6-20. It should be noted that part of the imbalance can be caused by the axial misalignment between the phases and associated flux linkage variance.

Figure 6-20: Phase Voltage variation during heat run due to temperature variation and temperature imbalance (a) Ferrite-TFM (b) NdFeB TFM.

The estimated winding temperatures at steady state using the ‘rise by resistance’ method are plotted for Phase-C and Phase-B. The method suggests measuring the winding temperatures at room temperature for reference. After reaching the steady state, the terminals were connected to a measurement system and corresponding winding resistance was measured at certain time intervals. Thus, from the reference resistance at a known temperature and the measured resistance during the process can be used to estimate the winding temperature. The blue lines in Figure 6-21 represent the estimated temperature with this method after stopping the current excitation. If the exact time of current excitation termination is known, a simple interpolation can be used to estimate the steady state temperature at the end of heat run. These estimated temperatures are shown with red squares in the figure with the actual thermocouple readings with yellow crosses. It is evident that there exists a large discrepancy between the thermocouple measurement and the estimated temperature using the resistance value. This discrepancy can lead to large error if precise estimation of copper loss is desired for the separation of magnetic loss components.
Figure 6-21: Temperature estimation at steady state using rise by resistance method (a) Phase-C (b) Phase-B.

6.5 Loss Separation

Even with the prototype available, accurately separating the losses for a better understanding of the responsible contributors can be a difficult task considering the non-idealities and practical constraints. Most importantly, with the change in the operating temperature, the motor parameters change significantly creating additional difficulties in the measurement and separation method. Although sounds straightforward, separation of copper loss components from total loss can be challenging due to the change in winding resistance and the difficulty in measuring the change with temperature variation.

Different approaches have been previously undertaken to separate the loss components in permanent magnet machines\cite{41}\cite{42}\cite{43}\cite{44}. In some of these cases, finite element analysis (FEA) based approach was adopted \cite{45}\cite{44}. A calorimetric measurement-based approach was proposed in \cite{43} which enables identifying the core loss at different frequencies. In \cite{44}, separation of eddy current loss was demonstrated but the process requires a separate rotor with removed magnets which poses additional challenges in setting up the experimental setup. A combination of FEA and experimental method and more elaborate loss separation has been proposed in \cite{41} which attempted separation of loss components depending on their order of relationship with excitation current and operating frequencies. Regardless of the methods applied, the isolation of the joule loss component from the total loss is a prerequisite and the accuracy in this procedure is essential. In this section, a simple and straightforward approach will be demonstrated to precisely separate
the joule loss component from the total loss. The procedure requires no additional hardware setup and can be easily automated. Application of the procedure provides an efficient means to accurately model the joule loss component which can be further used in all other test procedures.

6.5.1 Thermal Contribution in Loss Components

Among different factors of operating conditions, variation in temperature is the dominant contributor in adding complexity to loss measurement. Change in temperatures can affect the losses in different ways, and the following factors can be considered the most significant –

i) Copper loss variations due to change in winding resistance
ii) Magnetic loss variations due to change in magnet properties such as change in remnant flux density
iii) Lamination material properties (such as conductivity causing eddy current loss) that further changes the magnetic loss components.

Although seems simple, isolating the copper loss component from the total loss can be difficult due to temperature variation during the experiment. Moreover, local hotspots and non-uniform winding temperature complicate the process of temperature measurement and resistance estimation. In some literature, methods have been proposed to estimate the winding resistance for induction machine based on motor models [46][47] but these methods do not guarantee an accurate estimation. Signal injection-based methods can be another option, but this requires injecting a DC current in the system that can result in torque ripple. In addition, the DC injection requires a neutral connection in the system adding complexity to the system [46][48]. In the proposed method, a straightforward approach will be provided to get a more accurate representation of winding resistance as a function of embedded detector temperature readings.

6.5.2 Temperature Contribution in Joule Loss

In most cases, joule loss or copper loss is the simplest form of loss to be measured given that the excitation frequency is low enough not to produce any AC copper loss. In cases where all coils are uniformly placed and experience similar natural or forced cooling, winding resistances of all phases can be considered nearly identical. Thus, a uniform sinusoidal excitation will result in a DC waveform for copper loss component. In such cases, a line-to-line resistance measurement according to IEEE Std. 112 [40] is sufficient for winding resistance estimation at a certain temperature. The IEEE guideline for permanent magnet motors IEEE Std. 1812-2014 [49] also
suggests using the procedure similar to the poly-phase induction motor. However, the uniform winding cooling might not be the case for all motor structures and the unpredictability of heat extraction both for forced and natural cooling can add complexity to the system.

The experimental TFM is mounted with Phase A winding being the closest to the mounting plate and vice versa. Since air acts as a thermal insulator, the air-gap of the motor passes a small amount of heat making the mounting plate the true heat sink in the system. This causes a large temperature gradient in the system resulting in a significant difference between the phase resistances. The problem can be realized in radial flux machines as well with conventional windings especially for forced cooling system depending on the motor mounting orientation and placement of cooling nozzles. The experimentally measured copper loss for the prototype is shown in Figure 6-22 (a). The motor was excited with a perfectly sinusoidal current of 0.015 Hz to eliminate any magnetic loss components. The corresponding copper loss exhibits a second order oscillation and the peak copper loss coincide closely with the peak value of the phase C current. This indicates a significant increase in phase C resistance compared to other phases. Since there are only three unknowns, a time-varying acquisition of the loss data can be used to form the linear equations to estimate phase resistance values. Thus, assuming constant resistivity for three consecutive time steps

\[ [R_a(t) R_b(t) R_c(t)]^T = A(t)^{-1}P_{cu}(t) \]

where

\[ A(t) = \begin{bmatrix} I_a^2(t-T_S) & I_b^2(t-T_S) & I_c^2(t-T_S) \\ I_a^2(t) & I_b^2(t) & I_c^2(t) \\ I_a^2(t+T_S) & I_b^2(t+T_S) & I_c^2(t+T_S) \end{bmatrix} \]

and

\[ P_{cu}(t) = [P_{cu}(t-T_S) P_{cu}(t) P_{cu}(t+T_S)]^T \]

Figure 6-22 (b) shows the computed resistance for the 200 second time interval using Eqn. 6-3.

The significant variation in phase resistance is obvious from the plot where phase-C resistance is found to be more than 7% higher compared to that of phase-A. \( R_{inst} \) in Figure 6-22 (b) represents the instantaneous effective resistivity considering the current vector and can be expressed as

\[ R_{inst(t)} = \frac{P_{cu}(t)}{I_a^2(t) + I_b^2(t) + I_c^2(t)} \]

For a reasonable speed operation, this instantaneous resistivity value is not required and the cycle average value is sufficient which can be readily available by averaging the 3 phase resistance and denoted as \( R_{Eff} \) in Figure 6-22 (b). \( R_{Eff} \) eliminates the current dependence on the resistivity and
is the true effective for balanced excitation. The scope of this paper concentrates in a simple and effective way to represent this $R_{\text{eff}}$ as a function of measured temperatures at different locations.

![Figure 6-22](image_url)

Figure 6-22: (a) Three phase sinusoidal excitation of 0.015 Hz. The dotted line represents the copper loss variation (b) Three phase resistance found using Eqn. 6-3.

### 6.5.3 Experimental Setup and Test Procedure

The proposed procedure requires a simple setup with no modification in the motor. The motor is simply connected to the inverter driver with no load connected to the shaft. Instead of using position feedback, the motor is excited with a constant frequency sinusoidal excitation with extremely low frequency. In the actual experiment, a sinusoidal excitation of 0.015 Hz was used, which resulted in a slow rotor movement with constant angular speed (0.056 RPM). This enables to assume zero mechanical output power as the rotor is always aligned with the d-axis yielding no mechanical torque output. Four thermocouples were placed to measure the temperatures at different locations. The temperature readings for the thermocouples placed at phase A, B, C, and the mounting plate are shown in Figure 6-23 (a). It should be noted that for phase-A, the thermocouple was placed between the coil and the stator pole making this reading slightly smaller than actual coil temperature. Even for other phases, depending on the contact region of the thermocouples, motor mounting orientation and other mechanical factors, the sensor temperature readings are not necessarily the effective coil temperature. To come up with the actual effective coil temperature $T_{\text{eff}}$, a weighted average of the temperature readings was taken as

$$T_{\text{eff}} = \sum \sigma_n T_n$$  \hspace{1cm} 6-5
\[ R_{\text{Estimate}} = R_{\text{EstimateREF}} \left[ 1 + (T_{\text{effREF}} - T_{\text{eff}}) \alpha_{\text{Cu}} \right] \]  
\[ J = \frac{1}{N} \sum_{n} \left( R_{\text{Estimate}} - R_{\text{Eff}} \right)^2 \]

where \( R_{\text{EstimateREF}} \) is the effective resistance \( R_{\text{Eff}} \) computed using Eqn. 6-3. at any reference point and \( T_{\text{effREF}} \) is the effective coil temperature using Eqn. 6-5 at the same reference point. \( T_n \)'s are the temperature sensor readings for different thermocouples and \( \alpha_{\text{Cu}} \) is the copper resistivity temperature coefficient. The weight functions are obtained to minimize the objective function \( J \) which is the mean squared error (MSE) between the computed \( R_{\text{Eff}} \) and the estimated value \( R_{\text{Estimate}} \). Once the weight functions are evaluated for the motor with certain sets of embedded sensors, the same weight functions can be used throughout the experiment to precisely isolate joule losses from the total loss. This ensures a full mapping effective resistance with thermocouple readings for the whole thermal operating range of the motor and it takes the non-uniformity of resistance into account while coming up with the weight functions. The plots collected during the resistance mapping test procedure are shown in Figure 6-23 (a) through (d). Temperature reading during the test procedure are shown in Figure 6-23 (a). Since the input power consist of only joule loss at this frequency, the effective phase resistance values were computed using Eqn. 6-3. and shown in Figure 6-23 (c). This provided the desired \( R_{\text{Eff}} \) also shown in Figure 6-23 (c) and by minimizing the objective function, appropriate weight functions were computed to come up with \( R_{\text{Estimate}} \) as a function of \( T_n \)'s. The estimated copper loss using this estimated resistance is shown in green in Fig. 3(c) and denoted as \( P_{\text{CuEstimate}} \). It should be noted that, the measured copper loss \( P_{\text{Cu}} \) has the second order sinusoidal variation (as shown in Figure 6-22(a)) captured in the measurement due to extremely low frequency current and relatively faster sample rate for power measurement. But the mean value of the copper loss has precise correspondence with the estimated value. Unless the temperature sensors are not physically moved to different locations, the weight function computed in the procedure can be used for future motor testing.
Figure 6-23: (a) Thermocouple temperature reading for resistance mapping test procedure (b) Three phase measured currents (c) Calculated phase resistances for three phases and the weight function estimated value $R_{\text{Estimate}}$ (d) Measured and estimated copper loss throughout the process.

6.6 Efficiency and Losses

6.6.1 Measurement Procedure

Inverter-driven permanent magnet motors are often used for variable speed drive applications and using a single operating point efficiency does not justify the performance of the machine. Depending on the design target, the highest efficiency operating point can be carefully shifted and an optimized machine for a variable speed drive application can be achieved. Both the NdFeB and Ferrite TFM were operated at a variety of torque-speed points and efficiencies were computed over the entire torque speed region.
To state an efficiency number for certain torque-speed condition, the ideal way should be to wait for the machine to reach thermal steady state and thus calculate efficiency. In a variable speed drive system, however, thermal steady state for each operating point can be different due to different copper loss contribution resulting from a variety of operating currents. Even keeping the current magnitude constant does not guaranty similar thermal state due to change in magnetic loss components and change in physics behind the convective heat transfer. Coming up with steady state condition for each operating point would be extremely time consuming and thus the efficiencies for the map population were extracted during the thermal transient. Initial rapid temperatures were avoided to reduce error due to temperature variation. Since most of the measured quantities were acquired by the power analyzer, the input and output power, voltages, currents and measured torque speed values were in perfect synchronism. Only the temperatures were acquired using the data acquisition unit, synchronization was ensured by selecting equal sampling rate of 1 second for both devices. Both devices were synchronized with the control computer and synchronization was ensured.

Samples of the acquired data are shown in Figure 6-24. Figure 6-24(a) shows the temperature readings from the thermocouples. Mean torque and speed data are shown in Figure 6-24 (b) and (c) respectively. Finally, the calculated efficiency through the power analyzer is plotted in Figure 6-24(d). The motor control algorithm in the DSP used simple vector control with two PI controllers regulating the current. The reference current vectors were manually inputted during the operation and the MTPA points from FEA were taken as reference for initial reference calculation. A small tuning was performed during the operation to account for practical constraints such as position sensor offset, thermal effects and so on. The motor was not operating at MTPA condition during the full operation and thus the efficiency can vary for the same torque speed point. To get the best operation for any torque-speed point, the best efficiency sections were chosen, and a 5 second averaging were performed to eliminate measurement glitches.
Figure 6-24: Data acquisition and synchronization for efficiency map computation (a) Temperature readings from data acquisition unit. (b) Shaft torque from power analyzer (c) Motor speed from power analyzer (d) Efficiency from power analyzer.

### 6.6.2 NdFeB-TFM Efficiency and Loss Map

Experimentally measured efficiency for the NdFeB TFM is shown in Figure 6-25(a). Despite the unusually large air-gap due to mechanical issues, the motor exhibited a maximum efficiency of more than 87.5%. It should be noted that the primary concern during the design process was achieving high torque density and not the most efficient machine. A reasonable air-gap can, however, give much better efficiency due to the smaller current requirement for similar torque output. A prime concern in TFM is the low power factor and the proper design measures proved to be effective in a significant improvement in power factor. Since power factor is a strong function of the current vector and can become unity in flux weakening region, careful measure should be taken to state the appropriate power factor number. Since the motor operation deviated from the
MTPA operation, additional flux weakening caused the voltage to drop and consequent power factor rise during the operation. Thus, the apparent power limit was calculated from the inverter limit and power factor was measured from the computed apparent power limit. In a continuous torque of 20 Nm, power factor was more than 0.8 near corner speed and was 0.68 for peak torque operation at corresponding corner speed.

The phase RMS voltage and RMS currents are given in Figure 6-25(c) and (d) respectively. Since the motor exhibited imbalances in phase voltages as seen in the previous section, the mean value of the three voltages was used in the plot. The RMS current and the effective coil temperature shown in Figure 6-25(e) are two quantities needed for copper loss estimation. The procedure and coefficient used in the previous section for resistance estimation were used in measuring the effective coil temperature and by subtracting the estimated copper loss from total loss, the summation of magnetic loss contribution and friction loss was calculated and shown in Figure 6-25(f). It should be noted that the perfect loss separation would result in a perfect zero core loss contribution at zero speed. Over-estimated copper loss can lead to a negative loss in this low speed operating region and vice versa. The very small section of this negative loss region can be seen near the zero-speed region, but the amount of error is negligibly small. The inability of the dyno to maintain constant speed near this zero-speed region and associated torque pulsation also adds difficulty in the output power measurement and contributes to some measurement issues as well. Thus, the true contributor to this faulty core loss measurement at low speed can be difficult to determine.

6.6.3 TFM Operation and Loss-Comparison with FEA

According to the control algorithm described in Chapter-5, the motor always operates at the most efficient current vector available at the torque-speed point for certain DC bus voltage. The FEA operating points were post-processed and thus the absolute best current vectors can be chosen with zero susceptibility to control error. In practical operation, however, for any given torque command at a certain speed and available DC bus voltage, coming up with the perfect current vector can be challenging. Moreover, at speeds larger than corner speed, an inaccurate current vector selection can lead the PI controller to saturate in cases where the flux weakening is not sufficient enough to lower the terminal voltage. Pushing additional $d$-axis current will however will lead to higher copper loss and hence lower efficiency.
Figure 6-25: Experimentally measured data over the full torque-speed operating points. (a) Motor Efficiency (b) Mechanical Power, the cyan line indicates continuous operation limit (c) RMS phase voltage (c) RMS Current (e) Effective coil temperatures (f) Summation of core and friction loss.
A simple comparison of the RMS current magnitude for the entire torque-speed map is shown in Figure 6-26(a). This plot shows the difference between the experimentally measured current and the suggested current computed from FEA operating points. At low speed, the difference is minimal and an indication of close correspondence of the FEA performance with the experimental one. This also indicates achievement of a good control algorithm in choosing the appropriate current vector for the given torque command. At higher speed, however, the experimentally measured current was found significantly higher for the same torque output. This can be the result of non-ideal control such as failure to choose the most appropriate current vector, the presence of current harmonics at high-speed flux weakening region. Variation of temperature and associated drop in magnet strength can be also an attributing factor. This indicates that comparison of experimental FEA with the experimental one with such discrepancy will not be an acceptable method and the difference in current magnitude must be addressed. To mitigate the issue, the motor model was forced to operate at these non-ideal current vectors and the error between experimental and FEA RMS currents are shown in Figure 6-26(b).

Even with the modifications in the current vectors, the core-loss components in FEA are usually under-estimated as the loss data available are usually the supplied for ideal conditions using Epstein test method. The punching and bonding process adds significant degradation in the steel property and the lead to a significantly higher loss. Moreover, the ideal sinusoidal current assumption in FEA is necessarily not true particularly at high-speed flux weakening region where
the inverter reaches the limit. The computed efficiency and core loss from the material property shown in Chapter 5 are given in Figure 6-27(a) and (b) respectively. The large underestimation in core loss and corresponding high efficiency is visible in the efficiency map shown in the figure. At this stage, the loss separation performed in the previous stage can play a significant role in tuning the loss model of the FEA for future use. Instead of using the material property Steinmetz model, a tuning factor can be added to core loss computation such as

\[
P_{\text{core FEA}} = K_{hT\text{une}} K_h f B^2 + K_{cT\text{une}} K_c f^2 B^2 + K_{eT\text{une}} K_e f^{1.5} B^{1.5}
\]

As seen from the equation, each loss contributor was added with a tuning factor. An objective function was set to minimize the mean squared error between the measured and FEA core loss such as

\[
J_{\text{core}} = \text{MSE}(P_{\text{core FEA}} - P_{\text{core Exp}})
\]

Values of the tuning factors were determined by minimizing the objective function to match the core loss in the best possible way. It should be noted that the frictional loss component was ignored in the process as the unmagnetized rotor was not available for friction loss separation. Loss map computed after adding the tuning parameters and the core loss map are given in Figure 6-27(c) and (d) respectively.

**6.6.4 Ferrite-TFM Loss Map and Efficiency**

Similar procedures were implemented for the Ferrite TFM to obtain the complete efficiency map. Since the motor was primarily designed for NdFeB magnet, the optimum efficiency point for the Ferrite structure does not fall within the desired operating range of the prototype. It can be seen that the NdFeB exhibited most efficient operation near the neck of the torque-speed envelope to ensure good efficiency over the entire range. Significantly lower flux density operation reduced the magnetic loss component and thus pushed the optimum efficiency towards higher speed as shown in Figure 6-28 (a). Thus, the Ferrite-TFM can be operated at a higher speed and suitable for more high-speed low torque applications. Though if Ferrite is considered during the design process, much thicker magnets would have been used and that could lead to a higher torque machine. As anticipated, the power factor of the Ferrite TFM can be found to be smaller compared to the NdFeB counterpart. Though this machine operated at significantly higher efficiency. It should be noted that the torque transducer was rated for 56 Nm maximum torque measurement.
and the accuracy of measuring low torque such as 1 Nm is lower. The low torque region is, therefore, more susceptible to measurement errors.

![Efficiency (FEA-No Tuning)](image1)

![Core Loss (FEA-No Tuning)](image2)

![Efficiency (FEA-Tuned)](image3)

![Core Loss (FEA-Tuned)](image4)

**Figure 6-27:** (a) Efficiency map computed from FEA (b) Core loss map (c) Efficiency map after coefficient tuning (d) Core loss with tuned coefficients.

The summation of core and friction loss was separated for the Ferrite TFM as well and shown in Figure 6-28 (f). The lower accuracy in torque and thus output power measurement can lead to higher inaccuracy and core-loss separation and some negative core loss regions are seen in the loss map. However, the numerical value of this negative quantity was small enough as seen from the range of numbers in the color-bar in the loss data.
Figure 6-28: Experimentally measured data over the full torque-speed operating points. (a) Motor Efficiency (b) Power Factor, the cyan line indicates continuous operation limit (c) RMS phase voltage (c) RMS Current (e) Effective coil temperatures (f) Summation of core and friction loss.
6.6.5 Loss Comparison with FEA

A similar procedure was applied for the Ferrite TFM to match the current vector operation of the experimental motor with the FEA model. Due to lower core loss, the primary contribution for Ferrite-TFM was coming from the joule loss and was estimated accurately during the experiment. Thus, the efficiency plot computed from FEA data shows fewer deviations from the experimental map as seen in Figure 6-29(a). However, the efficiency in still more than 6% higher and the similar tuning method was applied. The tuned efficiency map and core loss map for the Ferrite TFM are shown in Figure 6-29(c) and (d) respectively.

![Figure 6-29: (a) Efficiency map computed from FEA (b) Core loss map (c) Efficiency map after coefficient tuning (d) Core loss with tuned coefficients.](image-url)
6.7 Performance Summary

Despite some initial mechanical issues, the final prototypes performed well and show a close correspondence with FEA prediction. Especially the torque characteristics and the torque-speed envelope showed decent performance. Some key motor parameters and performance numbers for both motors are given in Table 6-1.

Table 6-1: Prototype Machine Performance

<table>
<thead>
<tr>
<th>Design Parameters</th>
<th>NdFeB</th>
<th>Ferrite</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outer Diameter (mm)</td>
<td>137</td>
<td>135</td>
</tr>
<tr>
<td>Total Axial Length (mm)</td>
<td>124</td>
<td>124</td>
</tr>
<tr>
<td>Number of poles</td>
<td>32</td>
<td>32</td>
</tr>
<tr>
<td>Air gap length (mm)</td>
<td>1.6</td>
<td>1.5</td>
</tr>
<tr>
<td>Number of Turns</td>
<td>110</td>
<td>110</td>
</tr>
<tr>
<td>Wire Gauge (AWG)</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>Magnet Type</td>
<td>N42SH</td>
<td>Ferrite D3850</td>
</tr>
<tr>
<td>DC Bus Voltage (Volt)</td>
<td>250</td>
<td>250</td>
</tr>
<tr>
<td>RMS Current (A) (@peak Torque)</td>
<td>5.67</td>
<td>5.67</td>
</tr>
<tr>
<td>RMS Current Density (A/mm²) (@peak torque)</td>
<td>11.34</td>
<td>11.34</td>
</tr>
<tr>
<td>Peak Torque (Nm)</td>
<td>31</td>
<td>11</td>
</tr>
<tr>
<td>Continuous Torque (Nm)</td>
<td>21</td>
<td>8</td>
</tr>
<tr>
<td>Corner Speed (@peak torque) (RPM)</td>
<td>240</td>
<td>240</td>
</tr>
<tr>
<td>Corner Speed (@cont. torque) (RPM)</td>
<td>260</td>
<td>280</td>
</tr>
<tr>
<td>Power Factor (@ peak torque, 240 RPM)</td>
<td>0.68</td>
<td>0.4</td>
</tr>
<tr>
<td>Power Factor (@ cont. torque, 360 RPM)</td>
<td>0.8</td>
<td>0.5</td>
</tr>
</tbody>
</table>
Chapter 7: Summary and Conclusion

7.1 Work Summary

Transverse flux machines had been there in concept for more than hundred years and had always been known for high torque producing capability. But the lack of industry adoption can be attributed to several factors such as construction complexities, low power factor, high torque ripple and so on. Overall contribution in this research can be summarized as follows

1) A novel topology was designed with 2D flux path in the stator eliminating the necessity for unconventional material for motor construction such as SMC. The associated complexities occurring for using laminations were mitigated by design innovations such as over-lapping between the stator poles.

2) The moving part was simplified to a great extent for ensuring structural robustness and to reduce manufacturing complexities. Novelties were introduced in the stator support structure as well which demonstrated good manufacturability.

3) The well-known high torque ripple contribution was analyzed and factors contributing to it were identified with proper modeling. Effects of manufacturing tolerance and non-idealities were incorporated in the analysis. Despite the manufacturing issues, reasonable torque ripple performance was achieved in the constructed prototype.

4) The most repealing factor for TFM- the power factor issue was mitigated strongly. The comprehensive study of the leakage paths proved to be an effective mean for identifying the concerns. More importantly, each issue such as rotor and stator leakage were addressed with a systematic approach and noteworthy improvement can be observed in the experimental tests.

Moreover, some subsidiary works associated with the designing and testing stage can be considered valuable as well and

1) Simplified machine models were developed to reduce simulation time. These models can be an effective mean for motor performance analysis and controller development.

2) An improved method for accurate joule loss estimation was developed. The method is equally applicable for any type of electric machines and required no additional hardware.

3) The improved joule loss estimation was applied for separating the loss components. An objective function based optimization process was developed to improve the FEA accuracy in magnetic loss prediction.
7.2 Conclusion

Through innovative design approach, the research brought out the true potential of transverse flux machine in terms of torque producing capabilities. In near future, better manufacturing capabilities and improved machine tolerance can bring in opportunities for innovations in machine design. Most of the commonly used motors have gone through generations of development and the current status is the consequence of continual contributions from the research community. TFM s, on the other hand, were disregarded and minimal attempts can be found for truly addressing the associated concerns. Recent wide-scale electrification strategy in different sectors will definitely increase the demand for high torque machines and with proper design initiative, TFM s can be a strong contender in the competition. The potential applications and scope of innovations make TFM a deserving candidate in the research area. The performance improvements in this research revealed promising capabilities in terms of torque density and power factor and removed much of the factors that prevented this topology from entering the industry.
REFERENCES


