ANALYSIS AND CONTROL OF FIVE-PHASE PERMANENT MAGNET ASSISTED SYNCHRONOUS RELUCTANCE MOTOR DRIVE UNDER FAULTS

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ABSTRACT

This dissertation addresses advanced control methodologies for the five-phase permanent magnet assisted synchronous reluctance motor (F-PMa-SynRM) drive under various open phase fault conditions. F-PMa-SynRMs are principally reluctance-type machines which contain fewer magnets than permanent magnet machines. The major advantage of F-PMa-SynRMs is their inherent fault-tolerant capability, which makes them suitable for critical applications in the automotive and aerospace industries. However, under different open phase faults, F-PMa-SynRMs lose their primary advantages due to reduced average torque and higher torque ripple creating severe vibrations that may cause immediate system shutdown. Additionally, the parameter estimation becomes challenging due to the temperature variations and the presence of current harmonics. In these situations, it is essential to develop advanced fault-tolerant control (FTC) methods for F-PMa-SynRMs targeting the maximization of average torque, minimization of torque ripple, and accurate estimation of temperature. Also, during the FTC, a fault detector is necessary to implement any feedback control methods.

An optimal phase advance control method is proposed to maximize the reluctance torque with a minimum phase current under different open phase fault con-
ditions. Then, an active torque ripple minimization (TRM) technique adopting three major steps is proposed as follows: (i) active current harmonic identification, (ii) percentile harmonic injection, and (iii) vector rotation of healthy phases. After that, an analytical method is proposed to estimate magnet temperature in the F-PMa-SynRM without needing any temperature sensors. Finally, this dissertation develops a simplified fault detection method based on five-phase symmetrical component (SC) theory.

Extensive simulation using MATLAB and finite element method is done to validate the theoretical claims. For further validation, experiments have been conducted on a 2.9 kW F-PMa-SynRM in different operating conditions. A five-phase IGBT-based inverter, digital control based on Texas Instruments digital signal processor F28335, and a dynamo set-up have been developed to support experimental tests.

While applying the proposed FTC, a maximum of 92.2% torque is obtained experimentally under a single-phase open fault condition; in comparison, previous methods could achieve up to 80% torque. The torque ripple is improved from 22% to 11% using the proposed TRM method under the single-phase fault condition. Also, under the two-phase fault condition, 11% improvement is observed in the torque ripple with the proposed TRM method. The magnet temperature is also estimated under different operating conditions, and to have no more than 3.3% error compared to a direct measurement. Finally, the performance of the fault detection method is found to be almost 100% consistent with the theoretical studies.
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CHAPTER I
INTRODUCTION

This chapter starts with an overview of the research and the proposed solutions. Later, this chapter explains the fundamental theories of the electric machines. It includes the definitions of the electromagnetic characteristics, operating principles, the generation of the magnetic field due to the current, and the rotational torque. Finally, the operational mechanisms of a single-phase, a three-phase, and a five-phase electric machine are described utilizing those fundamental principles.

1.1 Overview

This dissertation attempts to improve the performance of a five-phase permanent magnet assisted synchronous reluctance motor (F-PMa-SynRM) drive under different fault conditions. F-PMa-SynRMs are hybrid types of electric machines that provide multiple superior features than the conventional three-phase machines. Especially, the open phase fault tolerance features are observed very beneficial for industrial applications. However, there are challenges with F-PMa-SynRMs which limit their wide acceptability in industries.

The major challenge observed is that under different fault conditions, F-PMa-SynRMs fail to generate enough average torque to continue the rotation. For
this reason, further operation of the system severely gets altered and requires immediate system shutdown. The later challenging task observed is that the torque ripple increases enormously which creates a dangerous vibration in the system. Also, the magnets in F-PMa-SynRM$\text{s}$ face non-uniform flux linkages which cause higher possibilities of demagnetization. For this reason, the temperature of magnets in F-PMa-SynRM$\text{s}$ require continuous monitoring. Finally, for a closed loop operation in any motor drive system, a continuous fault detection method is very important.

Targeting each of the challenges described above, this dissertation proposes several solutions to improve the overall drive performance under different fault conditions. Firstly, a phase current optimization method is developed to maximize the average reluctance torque. In this method, the remaining phase currents are increased, which is justifiable to avoid the higher saturation effects. Secondly, to minimize torque ripple, the phase current harmonics are analyzed and a harmonic suppression method is developed. In this method, a fast and accurate harmonic identification is adopted in the control loop to support the harmonic suppression. Then, for the magnet temperature estimation, a novel frequency determination algorithm is developed to support a cleaner signal injection method. Finally, a symmetrical components (SCs) theory is developed for the five-phase system to perform the fault detection. The fault detector carries the fault information to the main controller which initiates the necessary modification in the phase currents.
1.2 Fundamental Laws

Ampere’s law, the Faraday’s law, and Lorentz law are the fundamental laws which are essential to describe the principle of any electric machine. In this section, the definitions of these laws are provided.

1.2.1 Ampere’s Law

The integral form of the Ampere’s law is defined as follows:

\[ \oint_C H \cdot dl = I \]

where \( H \) (ampere per meter, \( A/m \)) is the magnetic field intensity, \( dl \) is the small length of the magnetic path, \( B \) (Tesla, \( T \)) is the magnetic field density, \( \mu \) is the permeability, \( I \) is the applied current. In other words, this theory suggests that for current flow in a conductor, there will a magnetic field encircling the imaginary contour \( C \). This phenomenon is shown in the Fig. 1.1. The direction of the field and the current follows the right-hand grip rule. That means the thumb points in the direction of the current flow, and the other fingers give the direction of the magnetic field.

1.2.2 Faraday’s Law

Faraday’s law suggests that a magnetic field will interact with an electric circuit to produce an electromotive force (EMF). This phenomenon is also known as electromagnetic induction process. Mathematically it is defined as follows:
\[ \epsilon = -N \frac{d\phi}{dt} \]  

(1.2)

where \( \phi = BA \), \( \epsilon \) is the EMF, \( N \) is the number of turns, \( \phi \) is the magnetic flux (unit is Weber or \( \text{wb} \)). In other words, this theory explains that if there is a magnetic material generating magnetic flux lines, and those flux lines are linking with a nearby coil with a change in time, then there will be an induced voltage (EMF) across the terminals of that coil. The minus sign in (1.2) represents the Lenz’s law which explains that the direction of the induced current is always such to oppose the change in the magnetic flux that generated it. This phenomenon is depicted in the Fig. 1.2. It shows that if a magnet with two poles as north (N) and south (S) is moved towards an electrical circuit that has a coil and a load (lamp), there will be current flowing through the circuit due to change in the flux linkage between the coil and the magnet.
1.2.3 Lorentz Law

A charged particle moving in a magnetic field shall experience a force along the right angles of the magnetic field and the moving direction. This phenomenon is explained in the Fig. 1.3a. If a current $I$ is flowing through the right arm of the conductor which is in the magnetic field $B$, it will experience a force $F$ in the downward direction. Similarly, as the current is coming out through the left arm of the conductor, it will experience a force in the upward direction. The direction of the force follows the left-hand rule which is shown in the Fig. 1.3b.
1.3 Electromechanical System

The electrical machine is one of the best examples of an electromechanical system. It can either convert the electrical energy to mechanical energy which is known as motoring action or transforms the mechanical energy to electrical energy which is known as a generating action. During the energy conversion process, the fundamental theories of electromagnetism are applied which are discussed in the previous section. The Fig. 1.4 shows a simple construction of an electromechanical system. It consists of a static part which is known as a stator and a movable part which is known as a rotor. The stator or the rotor can be a permanent magnet or an electromagnet. In this Fig. 1.4, the stator is considered as an electromagnet and the rotor as a permanent magnet. In this system, if a direct current $I$ is applied to the coil, as per ampere’s law, a magnetic field will be produced, and the energy is coming from the current shall be stored in the field. This energy is called stored magnetic field.
energy. This field energy is utilized to attract the rotor causing the rotation of the rotor. However, if the current is direct, the rotation will only occur to align the south pole of the rotor to the north pole of the stator. There will not be any further rotation unless the direction of the stator current is changed. If the direction of the stator current is changed continuously, there will be a continuous rotation of the rotor for the alternation in the electromagnetic poles of the stator. This principle is the fundamental mechanism of any DC machines. In conventional DC machine, the rotor is made of electromagnets. In these machines, brushes and commutators are utilized to change the magnetic polarity through the input current.

The stored energy in this field can be calculated from the Fig. 1.5. Fig.

![Diagram of flux linkage (\(\lambda\)) vs current (A) showing area above and below the curve indicating energy and co-energy.](image)
1.5 shows that the flux linkage ($\lambda = N\phi$) increases with the increment in the applied current. However, this relationship is almost linear before the saturation point (point C). After that, the flux linkage does not increase with the current. Fig. 1.5 also shows that there are two areas which can be defined as energy and co-energy. Mathematically, the energy stored in the filed due to the current can be defined as follows:

$$W_f = \int_0^{\lambda_0} id\lambda$$

(1.3)

From the Fig. 1.5, the energy can be calculated as follows:

$$W_f = \frac{1}{2}Li^2$$

(1.4)

Similarly, the co-energy can be defined as follows:

$$W'_f = \int_0^{i_0} \lambda di$$

(1.5)

In the linear region $W_f = W'_f$. However, there is no physical meaning of the co-energy except the benefit of calculating force and torque from it.

For rotational motion, the electromagnetic torque $T_e$ can be calculated from the co-energy as follow:

$$T_e(I, \theta) = \frac{\partial W'_f}{\partial \theta}$$

(1.6)

1.3.1 Inductance

For a linear magnetic system the inductance is defined as follows:
\[ L(\theta) = \frac{\lambda}{I} \]  

where \( L(\theta) \) is the inductance of the coil which is a function of the position for rotational motion. In Fig. 1.5, the slope of the \( \lambda - i \) characteristics curve is defined as the inductance. This figure depicts, beyond the saturation point \( C (I_c, \lambda_c) \), the slope decreases which means the inductance reduces at higher current. If more and more current is pushed beyond that point, there will be no increment in the flux linkage. In such condition, there will be more losses in the system.

1.3.2 Reluctance

Magnetic reluctance is analogous to the resistance of the electrical circuits and is defined as follows:

\[ R = \frac{F}{\phi} \]  

(1.8)

where \( F \) is the magnetomotive force (MMF), \( \phi \) is the magnetic flux. Reluctance can also be defined as follows:

\[ R = \frac{l}{\mu A} \]  

(1.9)

where \( l \) is the length of the material, \( A \) is the cross-sectional area, and \( \mu \) is and the permeability of the material.

1.3.3 Magneto-Motive Force

In the magnetic circuit, magnetomotive force (MMF) is the ability to create magnetic flux and. This is analogous to the electromotive force (EMF) of the electrical circuits.
Figure 1.6: Single phase current.

MMF is also presented in terms of number of turns and the current. Mathematically it is presented as below:

\[ MMF = NI \]  

(1.10)

where \( N \) is the number of turns in the coil, \( I \) is the current. Under fixed MMF from the stator, \( \phi \) and \( R \) are inversely proportional. In other words, higher the reluctance lower the magnetic flux and vice-versa.

1.4 Fundamental Principle of Single Phase Electrical Machine

In the previous section, the fundamental concept of the rotation of a rotor is described by considering the applied current is a direct current. In the Fig. 1.4, if
the applied current is regarded as an alternating current (AC) as shown in the Fig. 1.6, there will be an alternating change in the magnetic polarity in the stator field (positive/negative current changes the flux direction in the stator). Due to this alternation, the rotor shall experience an attraction and repulsion based on the stator magnetic polarity. Therefore, the rotor shall try to rotate. As the electric power is provided through one phase current, this simple system is regarded as single-phase machine. Mathematically the single-phase system is defined as follows:

\[ X_a = X_{am} \sin(\omega t) \]  

where \( X \) can be voltage or current, \( X_{am} \) is the magnitude of phase A, \( \omega \) is the electrical speed in rad/s, and \( t \) is the time.

A single-phase system with the stator winding orientation and equivalent circuit model is shown in Fig. 1.7. Here the cross (×) presents the current entering into that point, and dot (.) presents the current coming out of that point. In Fig. 1.7a, \( A \) is denoting the winding is entering into that point, and \( A' \) is denoting the current is winding is coming out of that point. In this condition, if the current is positive (during the positive half cycle of Fig. 1.6), as per the Ampere’s law (1.1), then there will be magnetic flux generating around the current point \( A \). The magnetic flux lines and the direction of them are shown in the Fig. 1.7a. Similarly, at the same time, the current is coming out of the point \( A' \). For the similar cause, there shall be flux lines around the point \( A' \). The magnetic flux lines around that point and the direction of them are shown in the Fig. 1.7a. For these flux lines around \( A \) and \( A' \),
Figure 1.7: Single-phase system: (a) stator winding orientation, and (b) equivalent circuit model.

there shall be a resultant magnetic field generating a north pole and a south pole as shown in the Fig. 1.7a.

During the negative half cycle of Fig. 1.6, the direction of the flux lines change as the current direction changes. In this condition, the resultant magnetic field shall be changing causing opposite poles around the points A and A' as it was
during the positive half cycle. In this way, the magnetic poles change their location on the stator side based on the current direction. If a permanent magnet or an electromagnet is placed inside the rotor fixing on a shaft, that rotor shall face an attraction or repulsion based on the alternating magnetic poles in the stator.

The advantage of single-phase electrical machines are given as below

1. Advantages are

   (a) they are simple in structure,

   (b) easier to control,

   (c) less expensive to manufacture than most of the other types of motors,

   (d) do not require frequent maintenances or repairs, and

   (e) if they need any repair, it can be done fast and easy way.

2. Disadvantages are

   (a) they are not self-starting motors,

   (b) they require starter capacitors or often configured with a split phase,

   (c) they generate a smaller amount of average power

   (d) suitable for low power applications such as domestic appliances, and

   (e) the output torque of is highly pulsating which limits its uses in high power applications.
1.5 Fundamental Principle of Three-Phase Electrical Machine

The principle of a three-phase machine is similar to a single phase machine except for the number of stator excitations. In a three-phase system, there are three excitations are provided which are electrical $120^\circ$ apart from each other. The three-phase system is widely utilized in high power applications. The major advantages of a three-phase electrical machines are as follows:

1. generate high power comparing single-phase machines
2. robust and sturdy
3. can operate in wide range of industrial applications
4. Torque ripple is much lower where torque ripple is defined as follows:

\[
T_{\text{ripple}} = \frac{T_{\text{max}} - T_{\text{min}}}{T_{\text{avg}}} 
\]  

(1.12)

where \(T_{\text{max}}\), \(T_{\text{min}}\), and \(T_{\text{avg}}\) are the maximum, minimum, and average values of the torque respectively.

5. No need for starter capacitors.

Mathematically the three-phase signals (voltage/current) can be defined as follows:

\[
X_a = X_{am} \sin(\omega t) \\
X_b = X_{bm} \sin(\omega t + 2\pi/3) \\
X_c = X_{cm} \sin(\omega t + 4\pi/3) 
\]  

(1.13)

where \(X_{abc}\) can be voltage or current, \(X_{am}\), \(X_{bm}\), and \(X_{cm}\) are the magnitude of the phase A, B, and C, \(\omega\) is the electrical speed in rad/s, and \(t\) is the time. The graphical presentation of the three-phase current is given in the Fig. 1.8.

A three-phase system with stator winding orientation is shown in the Fig. 1.9. In Fig. 1.9a, it is shown the windings (using conductors) in the stator are designed to maintain a spatial difference of 120° between each phase. The corresponding circuit equivalent is shown in the Fig. 1.9b considering the winding are connected as Y. It shows, there are three phase currents as \(I_a\), \(I_b\), and \(I_c\), which are generated from the three-phase voltages as \(V_a\), \(V_b\), and \(V_c\), respectively. The three equivalent phase resistances are \(R_a\), \(R_b\), and \(R_c\), and inductances are \(L_a\), \(L_b\), and \(L_c\).
1.5.1 Generation of the Rotating Magnetic Field

The fundamental principle of the operation of a three-phase AC machine which causes the rotating magnetic field is presented with the help of the Fig. 1.8 and Fig. 1.10. In Fig. 1.8, the three-phase current signals are marked with seven vertical positions to present different situation at the different time ($T_1 - T_7$). The corresponding field positions are presented in Fig. 1.10. Here the crosses present the current entering into those points and dots present the current coming out of those points.

At time $T_1$, the phase A current is zero, Phase B current is positive, and phase C current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.10a. At time $T_2$, the phase A current is positive, Phase B current is positive, and phase C current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.10b. At time $T_3$, the phase A current is positive, Phase B current is negative, and phase C current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.10c. At time $T_4$, the phase A current is positive, Phase B current is negative, and phase C current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.10d. At time $T_5$, the phase A current is negative, Phase B current is negative, and phase C current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.10e. At time $T_6$, the phase A current is negative, Phase B current is positive, and phase C current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.10f. At time $T_7$, the phase A current is negative, Phase B current is positive, and phase C current is negative. The resultant magnetic field is shown in Fig. 1.10g.
Figure 1.9: Three-phase system: (a) stator winding orientation, and (b) equivalent circuit model.
Figure 1.10: Rotating magnetic field in three-phase system (Fig. 1.8) at time: (a) T1, (b) T2, (c) T3, and (d) T4.
Figure 1.10: Rotating magnetic field in three-phase system (Fig. 1.8) at time: (e) T5, (f) T6, and (g) T7.
1.6 Fundamental Principle of Five-Phase Electrical Machine

The five-phase system has a higher number of stator phase excitations than the three-phase system. Those five excitations are electrical $72^\circ$ apart from each other. Due to these denser excitations, the five-phase system provides additional advantages over three-phase system such as

1. lesser torque ripple comparing,

2. higher fault tolerant

3. higher reliability

4. high power density

5. lesser per phase current stress

6. suitable for critical applications such as aerospace and automotive industries.

However, the five-phase system requires ten switches (MOSFET/IGBT) in the inverter side while three-phase system requires only six of them.

Mathematically the five-phase signals can be defined as follows:

\[
Y_a = Y_{am} \sin(\omega t) \\
Y_b = Y_{bm} \sin(\omega t + 2\pi/5) \\
Y_c = Y_{cm} \sin(\omega t + 4\pi/5) \\
Y_d = Y_{dm} \sin(\omega t + 6\pi/5) \\
Y_e = Y_{em} \sin(\omega t + 8\pi/5) \tag{1.14}
\]
where $Y_{abcde}$ can be voltage or current, $Y_{am}$, $Y_{bm}$, $Y_{cm}$, $X_{dm}$, and $Y_{em}$ are the magnitude of the phase A, B, C, D, and E, $\omega$ is the electrical speed in rad/s, and $t$ is the time.

The graphical presentation of the five-phase currents is given in the Fig. 1.11.

1.6.1 Generation of the Rotating Magnetic Field

The generation of the rotating magnetic in a five-phase system is different from the three-phase system as the stator excitations are provided in a shorter time interval (considering same rotor speed). Considering the stator winding connected as a star, the orientation the windings in a five-phase system is presented in the Fig. 1.12a. Unlike the three-phase system, the spatial difference in the stator windings is maintained 720 between the phases. The equivalent circuit model of a five-phase system is shown in the Fig. 1.12b. It shows, there are five phase currents as $I_a$, $I_b$, $I_c$, $I_d$, $I_e$.
and $I_e$, which are generated from the five-phase voltages as $V_a$, $V_b$, $V_c$, $V_d$, and $V_e$, respectively. The five equivalent phase resistances are $R_a$, $R_b$, $R_c$, $R_d$, and $R_e$ and inductances are $L_a$, $L_b$, $L_c$, $L_d$, and $L_e$, respectively.

The fundamental principle of the generation of the rotating magnetic field in five-phase systems remains similar to a three-phase system except for excitation numbers. The Fig. 1.13 explains the generation of the magnetic field in a five-phase system. The phase current in the Fig. 1.11 is utilized to describe the process. In Fig. 1.11, the five-phase currents are marked with ten vertical lines to represent different time conditions ($T_1 - T_{10}$).

At time $T_1$, the phase A current is zero, phase B current is positive, phase C current is positive, phase D current is negative, and phase E current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.13a.

At time $T_2$, the phase A current is positive, phase B current is positive, phase C current is negative, phase D current is negative, and phase E current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.13b.

At time $T_3$, the phase A current is positive, phase B current is positive, phase C current is negative, phase D current is negative, and phase E current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.13c.

At time $T_4$, the phase A current is positive, phase B current is negative, phase C current is negative, phase D current is negative, and phase E current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.13d.

At time $T_5$, the phase A current is positive, phase B current is negative, phase C current is negative, phase D current is negative, and phase E current is positive.
C current is negative, phase D current is positive, and phase E current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.13e.

At time T6, the phase A current is negative, phase B current is negative, phase C current is negative, phase D current is positive, and phase E current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.13f.

At time T7, the phase A current is negative, phase B current is negative, phase C current is positive, phase D current is positive, and phase E current is positive. Due to these currents, the resultant magnetic field is shown in Fig. 1.13g.

At time T8, the phase A current is negative, phase B current is negative, phase C current is positive, phase D current is positive, and phase E current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.13h.

At time T9, the phase A current is negative, phase B current is positive, phase C current is positive, phase D current is positive, and phase E current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.13i.

At time T10, the phase A current is negative, phase B current is positive, phase C current is positive, phase D current is negative, and phase E current is negative. Due to these currents, the resultant magnetic field is shown in Fig. 1.13j.
Figure 1.12: Five-phase system: (a) stator winding orientation, and (b) equivalent circuit model.
Figure 1.13: Rotating magnetic field in five-phase system (Fig. 1.11) at time: (a) T1, (b) T2, and (c) T3.
Figure 1.13: Rotating magnetic field in five-phase system (Fig. 1.11) at time: (d) T4, (e) T5, and (f) T6.
Figure 1.13: Rotating magnetic field in five-phase system (Fig. 1.11) at time: (g) T7, (h) T8, (i) T9 and (j) T10.
1.7 Salient and Non-salient Pole Machines

Based on the reluctance path between the stator and rotor the synchronous electrical machines can be divided into a salient pole and non-salient pole electrical machines. Fig. 1.14 shows a typical salient and non-salient pole electric machine. In a non-salient pole machine (Fig. 1.14a), there exists a uniform air gap flux linkage as all the magnets are mounted on the surface of the rotor. In other words, the reluctance path between the stator and rotor is uniform all over the surface. This type of machine is also known as surface mounted permanent magnet machines. In a salient pole machine (Fig. 1.14b), the air gap flux linkage is not uniform as the magnets are buried inside the rotor making the reluctance path non-uniform as well. This type of
machines is also known as interior permanent magnet machines.

Permanent magnet assisted synchronous reluctance motor is also known as the salient pole machine. In these machines, the amount of the magnet is lessened and buried inside the rotor. Also, in these machines, the reluctance path is not uniform which makes them salient pole machines.

1.8 Introduction to Direct Axis and Quadratic Axis

In an electrical machine, the direct axis (d axis) and quadratic axis (q axis) play an essential role during the operation. Fig. 1.15 shows the $d - q$ axes graphically in a reluctance machine. Conventionally, the $d$ axis is chosen towards the favorable flux path where the reluctance path is the lowest. On the other hands, the $q$ axis is chosen towards the unfavorable flux path where the reluctance path is the highest. The ratio between the $d$ and $q$ axes inductances is called the saliency ratio.
1.9 Introduction to Field-oriented Control

Field-oriented control (FOC) is the most popular control method for any permanent magnet machines. In this control method, the stator and rotor field is oriented to maximize the motor performances. Usually, the current through the stator winding generates the rotating filed as discussed in the previous section. The rotor magnetic field shall try to be aligned to the stator magnetic field. The rotating magnetic field in the stator is controlled by electronics switches to be ahead of the rotor magnetic field. Therefore the rotor keeps spinning. Fig. 1.16 shows the fundamental concept of the FOC.

There is an angle between the stator magnetic and rotor magnetic field which is defined as $\theta$. This angle plays a significant role to generate torque. For a surface mounted permanent magnet motor $\theta$ is maintained $90^0$ to generate maximum torque by controlling the stator excitations.
1.9.1 Field-oriented Control for Five-phase System

Similar to the three-phase machines, the motivation of the five machine control scheme originates from the control strategy of a separately excited DC motor. For example, in a separately excited DC motor, the flux and torque are individually controlled through the field current and armature respectively. Unlike the DC motor, in a permanent magnet AC system (three-phase or five-phase), there are no current flows in the rotor. Therefore, both the flux and torque control is done through the stator current control.

For this purpose, the stator currents are transformed into two orthogonal components known as the direct axis \((d\text{ axis})\) and quadrature axis \((q\text{ axis})\) components (please see section). For PMa-SynRM, the \(d\) axis components are responsible for the flux
control and the $q$ axis is component is responsible for the torque control. During the current control, the phase currents are transformed into $d$ axis current ($I_d$) and $q$ axis current ($I_q$) respectively. This transformation is necessary for the FOC of any three-phase or five-phase system.

1.9.2 Transformation of the Phase Components to $d-q$ Components

Two individual steps are followed during the transformation of the phase components to the $d-q$ axes components. The first step is to transform the $abcde$ components to $\alpha-\beta$ axes components which is also known as the CLARK transformation for a three-phase system. The second step is to transform the $\alpha-\beta$ axes components to the $d-q$ axes components which is also known as the PARK transformation. Altogether
Figure 1.19: Transformation: (a) Phase component($abcde$), (b) phase to $\alpha - \beta$ frame, and (c) $\alpha - \beta$ to $d - q$ frame.

the $\alpha - \beta$ axes are considered as the stationary reference frame, and the $d - q$ axes are considered as the synchronously rotating reference frame.

During the control strategy, the rotor information is detected by the encoder. The stator phase A current is considered as the reference axis and the $\alpha$ axis aligned towards this reference axis. The transformation from the $abcde$ frame to the $\alpha - \beta$ frame is shown in the Fig. 1.17. Mathematically the transformation can be derived
as follows:

\[
Y_{s\alpha} = Y_a
\]

\[
Y_{s\beta} = Y_b \cos(18^0) + Y_c \cos(54^0) + Y_d \cos(126^0) + Y_e \cos(198^0)
\]

where \( Y \) represents the voltage or current.

During this transformation, the \( d \) axis is aligned along the rotor axis, which is degree shifted from the stationary reference frame \( \alpha \). This transformation in the \( d-q \) frame is as follows:

\[
Y_{sd} = Y_{s\alpha} \cos \theta + Y_{s\beta} \cos(90^0 - \theta)
\]

\[
Y_{sq} = Y_{s\alpha} \cos(90^0 + \theta) + Y_{s\beta} \cos \theta
\]

Unlike the three-phase system, the five-phase system is controlled through two different synchronously rotating reference frames which are known as \( d_1 - q_1 \) frame (for the fundamental) and \( d_n - q_n \) frame (for the harmonic). The order of the harmonic depends on the design characteristics. It is assumed if the \( d_1 - q_1 \) frame rotates at the synchronous speed, \( d_n - q_n \) frame rotates at \( n \) times the synchronous speed.

1.9.3 Mathematical Presentation of the Voltages in \( d-q \) axes frame

Including the fundamental and the harmonic signal, the overall mathematical transformation matrix can be presented as in (1.17). In (1.17), the first row represents the transformation of the fundamental signals in the \( d \) axis and the second row represents the transformation of the fundamental signals in the \( q \) axis. Similarly, the third row
represents the transformation of the harmonic signals \((m^{th} \text{ order})\) in the \(d\) axis and
the fourth row represents the transformation of the harmonic signals \((m^{th} \text{ order})\) in
the \(q\) axis. Finally, the fifth row represents the zero sequence components.

\[
\begin{bmatrix}
Y_d \\
Y_q \\
Y_{dm} \\
Y_{qm} \\
Y_0
\end{bmatrix} = \begin{bmatrix}
\cos \theta & \cos(\theta - 2\pi/5) & \cos(\theta - 4\pi/5) & \cos(\theta - 6\pi/5) & \cos(\theta - 8\pi/5) \\
\sin \theta & \sin(\theta - 2\pi/5) & \sin(\theta - 4\pi/5) & \sin(\theta - 6\pi/5) & \sin(\theta - 8\pi/5) \\
\frac{2}{5} & \cos m\theta \cos m(\theta - 2\pi/5) & \cos m(\theta - 4\pi/5) & \cos m(\theta - 6\pi/5) & \cos m(\theta - 8\pi/5) \\
\frac{1}{2} & \sin m\theta \sin m(\theta - 2\pi/5) & \sin m(\theta - 4\pi/5) & \sin m(\theta - 6\pi/5) & \sin m(\theta - 8\pi/5) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}
\begin{bmatrix}
Y_a \\
Y_b \\
Y_c \\
Y_d \\
Y_e
\end{bmatrix}
\]

Using (1.17), the transient voltages in the \(d - q\) frame can be derived as follows:

\[
\begin{align*}
V_d &= I_d R_s + L_d \frac{dI_d}{dt} - E_d \\
V_q &= I_q R_s + L_q \frac{dI_q}{dt} - E_q \\
V_{dm} &= I_{dm} R_{sm} + L_{dm} \frac{dI_{dm}}{dt} - E_{dm} \\
V_{qm} &= I_{qm} R_{sm} + L_{qm} \frac{dI_{qm}}{dt} - E_{qm}
\end{align*}
\]

where, \(V_d\) and \(V_q\) are the \(d\)-axis and \(q\)-axis voltages, \(L_d\) and \(L_q\) are the \(d-q\) axis
inductances, \(E_d\) and \(E_q\) are back EMF in the \(d\) and \(q\) axis respectively, \(R_s\) is the
stator resistance, and \(m\) is the harmonic order. The equivalent circuit model of
(1.18) is drawn in the Fig. 1.20.

This chapter described the fundamental mechanism of a single-phase, a three-
phase, and a five-phase system. Electric machines are also classified based on the
Figure 1.20: Equivalent circuit model (a) $d$ axis (fundamental), (b) $q$ axis (fundamental), (c) $d$ axis (harmonic), and (d) $q$ axis (harmonic).

rotor structures. The following chapter describes the classification of those electric machines, the F-PMa-SynRMs, and the challenges associated with them.
In this chapter, different types of electric machines and their applications are discussed. Especially, detailed information on the five-phase permanent magnet assisted synchronous reluctance motors (F-PMa-SynRMs) is provided. Finally, the research motivation is explained to address the challenges which are associated with the F-PMa-SynRM.

2.1 Introduction

Electric machines are widely utilized in every aspect of our lives including the low voltage residential appliances, medium voltage commercial applications and high voltage process industries. Additionally, modern development in the automobile and aerospace industries boosted the global utilization of the electric machines significantly. It is reported that electric motors driven systems (EMDS) consumes the single most substantial electrical end-use, which accounts for 43% and 46% of all global electricity consumption [1]. In united states, the percentage is even higher as electrical motors make up the single largest end use of electricity consuming roughly 60% in industrial applications and 70% in the process industries [2]. Due to this mas-
sive demand, it is predicted by the Transparency Market Research that the global market \[3\] value of the electric machines shall reach to $120.68 billion by 2019.

2.2 Classification of Electric Machines

There are different types of electric machines that have been developed to match the fabrication cost limit, operating performances, and higher reliability. The major classification of the electrical machines is presented in the Fig. 2.1. Electrical machines are mainly classified into two major groups: brush machines and brushless machines. Alternatively, the brush machines are also known as direct current (DC) machines which utilize the commutator to continue the current flow in the same direction.
Based on the field generation, the DC machines are divided into two major types which are known as wound field and permanent magnet (PM) field type DC machines. The wound type DC machines are further divided into three categories which are known as shunt field, series field and separately excited DC machines. All these DC machines are well known for their easier construction and more straightforward control. However, the primary disadvantage of these machines is the involvement of the commutator with a brush that requires frequent maintenance and is responsible for the additional mechanical loss.

The brushless machine was first invented in 1885 by Galileo to overcome the issues of brush machines. This machine was also called as the asynchronous machines. Unlike the DC machines, the voltage in the rotor of these asynchronous machines is induced which later helped them get known as induction machines widely. Induction machines (IMs) are much more straightforward to construct and do not require frequent maintenance which helped them to get widely accepted in large industrial applications. However, due to their lower efficiency, in high-performance applications, the adaptation of IMs has been limited.

Synchronous machines have started gaining attention for their higher efficiency and a wide range of rotor construction. Based on the rotor structure, these machines can be categorized as wound rotor, PM rotor (SPM), reluctance rotor which does not contain any magnets, and a hybrid rotor which contains both magnet and reluctance path. Wound rotor synchronous machines have current flow in the rotor that generates additional losses in the rotor. Except for Wound rotor, other syn-
chronous machines do not have current flow in the rotor which supports them to provide higher efficiency. PM machines contain large magnets which are mounted on the surface of the rotor, and, therefore not suitable for high-speed operation. Also, the magnet cost is higher on these machines. Reluctance rotor machines are proposed for higher speed operation as they do not contain any magnets in the rotor. However, these machines generate vibration leading to higher torque ripple. A hybrid type of machines is proposed to support the reduced magnet cost and higher speed operation. Based on the usage of the magnetic material hybrid type of machines are further divided into two categories which are known as interior type PM (IPM) machines, and PM assisted synchronous reluctance machines (PMa-SynRM). In IPM, the magnetic torque is dominant as it contains relatively higher magnetic materials, whereas, in PMa-SynRM, the reluctance torque is dominant and contain less magnetic materials. Due to the lesser magnetic material buried in PMa-SynRMs, they are suitable for low-cost and high-speed applications.

After asynchronous and synchronous machines, doubly salient machines are the next large family of the brushless machines. They are known as the switched reluctance machines (SRMs), and stator PM machines. Based on the control over flux, the stator PM machines are divided into hybrid-excited PM (HEPM), flux-reversal PM (FRPM), and flux-switching PM (FSPM), doubly-salient PM (DSPM), flux-mnemonic PM (FMPM). There are other applications specific brushless synchronous machines are proposed which are known as vernier and stepper machines.
2.3 Five-phase Permanent Magnet Assisted Synchronous Reluctance Motor (F-PMa-SynRM)

Permanent magnet assisted synchronous reluctance motor (PMa-SynRM) is a hybrid of a synchronous reluctance motor (SynRM) and an interior permanent magnet motor (IPM). Unlike the IPM, SynRM does not contain any magnets inside the rotor [4, 5]. In the sixties, the design of different SynRMs has been studied extensively [6, 7]. However, the control methods of SynRM were challenging as the modern vector control has not been adopted at that time. The modern version of the SynRM is proposed in the early nineties [8, 9]. Though these SynRMs were low-cost, they had deficient power factor with relatively higher torque ripple. To improve the power factor and torque ripple, a small permanent magnet is added in the SynRM which later became familiarized as PMa-SynRM [10–12].

The rotors of an IPM, a SynRM, and a PMa-SynRM are shown in the Fig. 2.2. It is shown, a PMa-SynRM rotor has been evolved from an IPM and a SynRM rotor. The conventions of the \( d \) – \( q \) axes in all the configuration are also shown in the Fig. 2.2. Conventionally, in IPM, the \( d \) axis aligned along the permanent magnet axis and the \( q \) axis are quadratic of the \( d \) axis. In SynRM, the \( q \) axis aligned along the flux barrier and the \( d \) axis is quadratic to the \( q \) axis. However, in PMa-SynRMs, the \( q \) axis aligned to the magnet axis and the \( d \) axis is quadratic to the \( q \) axis as their characteristics mostly match with SynRMs.

For being hybrid in structure, PMa-SynRMs show higher cross saturation ef-
ffects, and higher torque ripple while comparing with similar PM machine. Extensive research is done on PMa-SynRM to address these nonlinear issues by incorporating advanced control and design methods [13–16]. Also, the stator of a PMa-SynRM can be chosen as single-phase, three-phase or multi-phase. Fundamentally, the choice of phase number depends on the required performances such as power density, torque ripple, and system reliability (fault tolerant capability). Here, five-phase stator excitation is considered to improve the torque ripple and the fault-tolerant capability.

Figure 2.2: Rotor of a PMa-SynRM.
However, under different open phase faults, the average torque reduces significantly along with higher torque ripple which promotes severe vibration. In this dissertation, different control methodologies are developed to improve its torque characteristics under various open phase faults.

2.3.1 \( d \) and \( q \) axis Modeling of The F-PMa-SynRM

Fig. 2.2 shows that the permanent magnet in the rotor of a PMa-SynRM is placed in the \( q \) axis. So, the back EMF generating for the \( q \) axis current shall be reflected in the \( d \) axis voltage. Considering the transformation matrix in the previous chapter, the transient equations of the F-PMa-SynRM can be derived as follows:

\[
\begin{align*}
V_{ds} &= I_{ds}R_s + L_{ds}\frac{dI_{ds}}{dt} - \omega_r(L_{qs}I_{qs} - \lambda_{PM}) \\
V_{qs} &= I_{qs}R_s + L_{qs}\frac{dI_{qs}}{dt} + \omega_r(L_{ds}I_{ds}) \\
\lambda_{qs} &= I_{qs}L_{qs} - \lambda_{PM} \\
\lambda_{ds} &= I_{ds}L_{ds}
\end{align*}
\]

(2.1)

where \( V_{ds} \) and \( V_{qs} \) are the \( d \)-axis and \( q \)-axis voltages, \( I_{ds} \) and \( I_{qs} \) are the \( d \) axis and \( q \) axis currents, \( \lambda_{ds} \) and \( \lambda_{qs} \) are the magnetic flux linkages in the \( d \)-axis and \( q \)-axis, and \( \lambda_{PM} \) is the permanent magnet flux linkage, \( \omega_r \) is the Back-EMF resulting from the permanent magnet, \( L_{ds} \) and \( L_{qs} \) are the \( d \)-\( q \) axis inductances, \( \omega_r \) is the rotor speed, and \( R_s \) is the stator resistance.

Utilizing (2.1), the electromagnetic torque can be derived as follows:

\[
T = \frac{5P}{4}(\lambda_{PM}I_d + (L_d - L_q)I_dI_q)
\]

(2.2)
where $P$ is the number of poles.

2.3.2 Finite Element Modeling of the F-PMa-SynRM

A finite element analysis (FEA) model (2-D) of the F-PMa-SynRM has been developed which is shown in Fig. 2.3. The motor model contains less permanent magnet (Neodymium) in $q$-axis direction and the flux barriers for generating reluctance torque. The model has fifteen slots, four poles, and double layer distributed winding (start connected) with 0.94 winding factor. It has been further optimized using the differential evolution strategy (DES) to minimize machine loss and cost [17]. This FEA model will be utilized in the simulation section to validate all the theoretical claims.
2.3.3 Classification of Open Phase Faults

Fig. 2.4 shows the open phase fault model of the five-phase PMa-SynRM. There can be three types of potential open phase faults in a five-phase machine. These faults are single-phase open fault (SPF), two-phase adjacent open fault (TPAF) and two-phase non-adjacent open fault (TPNF). For example, Fig. 2.4 shows various open phase faults including the SPF (Phase $A = 0$), TPAF (phases $AE = 0$), and TPNF (phases $BE = 0$) open faults. During these faults, the torque production capability of the F-PMa-SynRM significantly reduces.
2.4 Research Purposes and Contributions

This dissertation primarily focuses on the fault-tolerant capability of an F-PMa-SynRM to improve the torque characteristics under above mentioned open phase fault conditions. During these faults, the F-PMa-SynRM fails to generate enough torque in the output. Until now, most of the fault tolerant techniques are proposed based on feeding higher phase current in the remaining healthy phases which theoretically is not a proper solution for an F-PMa-SynRM. With higher phase current, the F-PMa-SynRM prone to reach saturation quickly resulting in reduced amount of reluctance torque. To solve this critical issue, this dissertation proposes a technique to maximize the reluctance with a limited amount of current which eventually avoids any saturation effects, thus can provide relativity higher average torque.

Another method is proposed to reduce the torque ripple during those fault conditions. The torque ripple is mainly caused by the unpredicted harmonics in the phase currents, which is highly destructive in reluctance type of machines due to their non-linearity in nature. To address this critical challenge, a real-time active torque ripple minimization (TRM) technique has been proposed under open-phase faults. The proposed solution efficiently consists of three different steps as following (1) current harmonic identification, (2) percentile current harmonic injection, and (3) vector rotation of healthy phases. A correlation-based signal identifier has been proposed to support the real-time analysis of phase harmonics.

This study also includes an analysis of the rotor magnet temperature esti-
For efficient control performance, the accurate temperature estimation has to be dynamically feedback to the central controller as it directly affects the parameter estimation. Due to the inflexibility of the rotor magnet, the direct measurement of the magnet temperature is challenging. This dissertation presents an analytical approach to the temperature estimation of the magnetic material utilized in the F-PMa-SynRM, which is based on a signal injection method through a frequency determination algorithm. The proposed method presents an analysis of the frequency spectrum to isolate the fundamental and fault frequencies and their respective harmonics to determine a range of available frequencies. The available frequencies are further filtered based on the noise to signal ratio constraints.

Finally, this dissertation also proposes a fault detection method which mathematically analyses the above mentioned open-phase fault conditions and develops a symmetrical components (SCs) theory to extract the unique feature of those faults. This analysis will provide the types of faults by logically analyzing the pattern of magnitude and phase angle changes of the fundamental signal in the SCs. Finally, those fault information is utilized in the forward path controller for further modification in the phase currents.

2.5 Dissertation Outline

The dissertation is presented in different chapters as follows:

Chapter 3 discusses the relevant literature reviews on the existing techniques that are considered during the fault tolerant control of multi-phase electric machines.
Additionally, previous works on different torque ripple minimization (TRM) methods, fault detection (FD) techniques, and magnet temperature estimation methods are presented in the same chapter.

Chapter 4 discusses the proposed solution for the maximization of the reluctance torque in F-PMa-SynRM under different open phase faults. Here, detailed analytical modeling under faults and a phase advance technique is developed to improve the reluctance torque.

Chapter 5 discusses the proposed technique for the minimization of the torque ripple in F-PMa-SynRM under different open phase faults. Here, an active harmonic suppression through the harmonic identification is proposed to estimate the magnitude of the current harmonics.

Chapter 6 discusses a method for the magnet temperature estimation in F-PM-SynRM. A novel frequency determination algorithm has been proposed to estimate the magnet temperature accurately.

Chapter 7 discusses the proposed method for the open phase fault detection in the F-PMa-SynRM without requiring any additional hardware. Here, symmetrical components theory for a five-phase system has been developed and utilized to identify those open phase faults.

Chapter 8 presents the simulation results of the proposed methods.

Chapter 9 presents the experimental validation of the proposed methods.

Chapter 10 concludes with the findings of the proposed methods and future works which can potentially extend this dissertation.
CHAPTER III
LITERATURE REVIEW

This chapter provides the literature reviews of the techniques which are proposed earlier to perform under fault tolerant operation of multi-phase motor drive systems. It covers the previous works on different fault tolerant control methods of the five-phase system, and various torque ripple minimization techniques, and their limitations. After that, literature reviews on magnet temperature estimation are also provided. Finally, various fault detection methods which are applied in three-phase and five-phase system are discussed.

3.1 Maximization of Reluctance Torque under Different Open-phase Faults in F-PMa-SynRM

A significant amount of research has been done for the reliable control of electric machines for critical service applications for medical, military, and transportation purposes. To maximize the fault tolerance capability while minimizing cost, the multiphase motor has been suggested for transportation applications [18, 19]. Among all the multiphase systems, a five-phase system is a minimum configuration comparing other multiphase systems that inherit greater fault tolerance capability than a conventional three-phase system [17, 20, 21]. A system with more than five phases would
have to deal with more power switch failures in operation which could potentially increase the complexity of the fault detection and fault tolerant control. Along with the five-phase configuration, a PMa-SynRM, which inherits the advantages of both a synchronous reluctance machine (RSM) and a permanent magnet synchronous machine (PMSM) [22,23], can provide a higher range for reluctance torque control [17,24–27] by providing optimal phase advance during operation [28]. However, in the event of open phase faults (single or multiple phases) in a PMa-SynRM, the formulation of the optimal phase advance to provide maximum reluctance torque becomes complex which has not been studied in detail until now.

3.1.1 Fault-tolerant Operation through Hardware Modeling

There are several strategies which have been reported previously for reliable fault-tolerant control (FTC) of electric machines proposing the alternative design of the hardware [29–32]. In [29], under fault conditions, redundant phases in the inverter side or additional machines in parallel have been considered to perform the safe and fault-tolerant operation. This method accurately fits in where zero tolerance is adopted under faults, and the additional hardware cost is not a primary concern. Back electromotive force based modeling for different winding configurations of a multi-phase machine has been discussed in [30]. The proposed method mostly discussed from the winding design point of view of a surface mounted permanent magnet machine. Therefore, the difficulties with the reluctance type of machine such as saturation effect are not well pictured in the method. Multi-phase drive modeling with space
vector modulation strategy has been discussed in [31]. The method included two separate five-leg inverters with two isolated DC sources which can not apply to the limited cost applications. A double-star PMSM has been proposed for FTC in [32]. Nevertheless, the study did not confirm the solution holds for maximization of torque while minimizing the saturation effects in an F-PMa-SynRM.

3.1.2 Fault-tolerant Operation through Active Control

Fault tolerant control has also been achieved through a pure control approach (no hardware modification) for both types of three phase and multi-phase machines. High-frequency injection with sensorless control [33] and adaptive back-stepping observer based field orient control [34] are suggested for a three-phase PMSM. To the best of author knowledge, multi-phase phase FTC methods are different and relatively complex than of the three-phase FTC. FTC for multi-phase machines is suggested in various literature [35–49]. In [35], a genetic algorithm is studied for a five-phase flux-switching machine under short circuit fault, thus, has not been validated for complete open phase faults. FTC for multi-phase induction machines is studied in [36–38]. However, the nonlinear behavior of Multiphase PMSM or PMa-SynRM cannot be standardized by the multi-phase induction machine behavior under faults. FTC for regular permanent magnet type of multi-phase machines (mostly generate magnet torque) is studied in [39–48]. In [39], FTC with trapezoidal back-EMF modeling [39] is suggested. This method is not applicable to a machine which has sinusoidal MMF. FTC focusing on the minimization of torque ripple is suggested in [40,42].
In [40,42], the methods have been applied to reduce the torque ripple while sacrificing the average torque. In [43], FTC under flux weakening operation has been studied for a regular permanent magnet machine. This study did not analyze the maximum torque realization under faults. On the other hand, to sustain the equal amount of torque under faults, five-phase permanent magnet machine has been studied in [44–48]. For this purpose, in [44–46], higher phase currents have been allowed in the healthy phases while disconnecting the faulty phases. However, by increasing the phase currents in a PMa-SynRM (the majority of the torque is being contributed from the reluctance [50, 51]), critical machine parameters could reach a saturation which would lead to decreased torque, lowered efficiency, increased temperature. To overcome the saturation issues while retaining the advantages of an F-PMa-SynRM, the magnitude of the phase currents must be limited under faults. Additionally, the reluctance torque can be maximized under faults so that a five-phase PMa-SynRM system can run with less disruption.

3.2 Torque Ripple Minimization under Open-phase Faults in F-PMa-SynRM

Multi-phase PMa-SynRM has become promising candidates for electric propulsion applications in electric and hybrid vehicles, aerospace industries, and military equipment due to its low torque ripple [52]. Unlike surface mounted PM machine, during the fault-tolerant operation of PMaSynRM, the cross-coupling effects in the MMF significantly increase due to the compound nature of the higher reluctance and lower magnetic properties [49]. This phenomenon introduces random phase current harmonics
which directly affects the electromagnetic torque with higher ripple sub-sequentially causing severe vibration [53,53].

3.2.1 Torque Ripple Minimization for Conventional Electrical Machines

Several methods are suggested in the conventional three-phase machine to reduce torque ripple. For example, neural network based feedback signals estimator has been proposed in [54–57]. This approach works with the higher computational complexity with extensive training of the control sequences. Thus, the training outcome can become nondeterministic with a significant dependence on the choice of the initial parameters. Model-based predictive current control has been proposed in [58–61]. However, under faults, the non-ideality in the motor phase currents due to the unpredictable MMF harmonics, increases the challenges of the model-based predictive method. Further solutions which are asymmetry switching patterns [44], classical proportional-integral-based iterative learning [40], on-line estimation of phase resistance [62], fuzzy-logic controller [63], and adaptive internal model [64] have been advised for reducing torque ripple. Nevertheless, all these methods have been reported for a three-phase machine under the healthy condition, thus requires further justification to perform under complicated phases faults in an F-PMa-SynRM.

3.2.2 Torque Ripple Minimization for Multiphase Machines

Various torque ripple minimization (TRM) methods are suggested for multi-phase machines in [41,45,65–67]. In these methods, DC link current control [41], model-based predictive control [45], and higher level inverter design [65,66] are advised
respectively to reduce the torque ripple. However, the addition of redundant hard-
ware or model prediction under faults in multi-phase machines is not economical and 
relatively challenging. Further suggestions on TRM under faults in multi-phase ma-
chines are given by designing different stator [50], and a reduced order transformation 
matrix [51] are presented for reducing torque ripple with a limited explanation on the 
phase harmonics effects, which is significantly higher in five-phase PMa-SynRMs. 

TRM for surface mounted permanent magnet type of multi-phase machines are stud-
ied in [39,40,44–48,68,69]. In [68], the sliding mode controller is suggested to reduce 
torque ripple under open-phase faults by allowing the phase current is chattering to 
some extent. In [69], a vectorial approach limited with harmonic effects has been 
studied. Nevertheless, the nonlinear behavior of multiphase PMSM or PMa-SynRM 
cannot be standardized by the multi-phase induction machine behavior under faults.

3.3 Magnet Temperature Estimation in F-PMa-SynRM

The temperature dependency of the demagnetization of magnets [70] gravely affects 
the performance of permanent magnet motors [71–73]. With the growing trend of 
electric motor designs toward High Power Density (HPD) configurations in thermally 
critical applications such as electric vehicles [74, 75], this problem is further exacer-
bated due to the higher amount of heat generated and the reduced surface area for 
heat dissipation [76]. Additionally, the use of relatively temperature sensitive rare 
earth magnet materials such as Neodymium Iron Boron (NdFeB) in permanent mag-
net motors is increasing because of their high power density, mechanical strength, and
relatively lower cost [70]. However, these magnets not only have high-temperature coefficients of remanence and coercivity but additionally have a relatively low Curie Temperature of $\approx 580 - 600K$. At this temperature, these magnets lose their ferromagnetic properties and begin to exhibit para-magnetism which would permanently damage the motor.

### 3.3.1 Signal Injection versus Direct Measurement of Magnet Temperature

This magnet temperature dependency on the performance, reliability, and efficiency of permanent magnet motors necessitates the need for effective and convenient thermal analysis and temperature measurement of permanent magnet motors. The signal injection method presents a convenient, non-invasive and easily implementable method for magnet temperature estimation that can be used for on-line magnet temperature determination [77]. While temperature sensors may be employed as tools for direct temperature measurement, they are extremely cumbersome, invasive and inconvenient in their application as a tool for on-line magnet temperature measurement of permanent magnet motors. Due to the continually rotating rotors on which the magnets are usually mounted, the measurement setup often requires a cumbersome wireless data acquisition system which increases the complexity of the system [78, 79].

### 3.3.2 Estimation of Magnet Temperature Using FEA/CFD/LPTM

Finite Element Analysis (FEA) and Computation Fluid Dynamics (CFD) methods available as elaborate commercial software may be used to estimate magnet temperature at various operating conditions. However, these tools cannot be used for
on-line real-time temperature estimation because of their excessive computation and simulation time. Lumped Parameter Thermal Models (LPTM) [80–84] on the other hand, are reduced finite element models in which the various geometries of the motor are lumped to form equivalent thermal network which is evaluated to determine the approximate temperatures at the various parts of the motor. While this method may be implemented as an on-line tool using Digital Signal Processors (DSP) [85] due to their significantly reduced simulation time, the major drawback of the LPTM is its exclusivity to a specific motor geometry and configuration.

3.3.3 Estimation of Magnet Temperature Using Signal Injection Method

The signal injection method exploits magnet resistance temperature dependence to determine its temperature. This method has been employed to determine the rotor temperature in the case of Induction Motors (IM) [86], and magnet temperature in the case of Surface Permanent Magnet Synchronous Motors (SPMSM) [87,88] and Interior Permanent Magnet Synchronous Motors (IPMSMs) [88]. In the case of the IM [86], a low frequency model was implemented and extensive sensitivity analysis experiments were conducted to isolate relatively lower frequencies at which the the rotor temperature dependency was easily detectable at a fixed operating condition using this method.

In [87], a high-frequency model is employed for a surface mounted magnet machine with better spectral separation of the injected signal from the operating frequency. It is proposed to inject the high-frequency intermittently to reduce the
adverse effect of the normal operation. In [88], pulsating high-frequency injection is proposed only in the $d$ axis for salient pole machines, which reported as insensitive to the speed and inductance variation. The high-frequency model proposed in [89] took into consideration the high-frequency resistances which are typically considered negligible at higher frequencies due to the more significant inductive component of the corresponding high-frequency impedance. In these methods, however, only the general requirements of the injected signal are determined with a proposed carrier signal frequency. To the best of the author’s knowledge, limited studies are done on the adaptive frequency determination algorithm to dynamically determine a suitable frequency of the injected signal to estimate the magnet temperature. To the best of the author’s knowledge, limited studies are done on the adaptive algorithm to dynamically determine a suitable frequency of the injected signal to regulate a relatively higher accuracy of the magnet temperature.

3.4 Open Phase Fault Detection in F-PMa-SynRM

Extensive research has been accomplished to improve the reliability of electric machines under critical service applications such as aerospace and automotive [90, 91]. Notably, electric motors have been popularly applied to the hybrid and electric vehicles where the reliable operation is mandatory. For such critical applications, advanced fault detection electric machine has been predominantly required. To maximize the reliability of the drive system, a multi-phase motor system has been suggested in [21, 92–94].
3.4.1 Fault Detection for Conventional Electrical Machines

Many methods have been extensively studied in last few decades for the fault detection and diagnosis of electric motors [95,96]. Fuzzy logic or neural network theory [97–101] is one of these conventional methods. This method requires an expert system [102,103] based on the rules set up by the accumulated experience. Open phase fault or short circuit fault has been analyzed with a motor dynamic model which uses parameter estimation [104,105] and state estimation [106,107]. This method requires estimation of accurate physical parameters to identify a precise system model. In [107], a simple diagnosis process has been proposed for three-phase BLDC which analyzed the switching status during the fault conditions. In [108], a fault detection process has been analyzed which uses extra circuitry to identify the faults. Advance signal processing has been attempted for three-phase machines in [45,109–113]. In [111–113], along with a signal processing technique, three-phase reference frame theory has been introduced to quantize the fault signature into a DC quantity. These techniques incorporate large computation burden on the controller.

3.4.2 Fault Detection based on Motor Current Signature Analysis

In recent years, advanced fault detection in three-phase motor drives has received much attention. Motor current signature analysis using kernel density estimation has been discussed in [109]. A mixed logical dynamic motor drive model has been discussed in [114]. An extended form of the Kalman filter associated with an appropriate model of permanent magnet synchronous machine has been discussed in [115]. Inter-
turn short circuit faults are detected in [116–118]. In [116,117], advanced analysis on the frequency pattern in motor current and back electromotive force is analyzed to identify faults. In [118], injection of pulsating voltage which induced high-frequency current is suggested which applies to the low-medium speed motor drive system. Intelligent particle filter which is applicable in a nonlinear system is discussed in [119].

3.4.3 Fault Detection for Multi-phase motor

Until now, the fault detection and condition monitoring theory have been commonly developed by assuming a three-phase system which may not be effectively applied to multi-phase machines. For a five-phase system and its diagnosis, an advanced fault detection algorithm is required which considers a minimum of ten power switches and their combinational complexities. Few studies have been made in [21,120,121] on fault detection in a five-phase motor. In [21], a model-based observer is proposed to estimate parameters to identify open transistor faults in a five-phase permanent magnet motor drive. In [120], combined space vector spectrum analysis has been done with a complex theoretical explanation for detecting an inter-turn fault in a five-phase machine. In [121], a centroid based switch fault detection method has been discussed utilizing the reference frame theory in a five-phase system.

3.4.4 Fault detection based on Symmetrical Theory

To accommodate the theoretical complexities [120,121] with easier fault detection, the conventional SCs theory can be applied as an excellent tool for the condition analysis of a general multi-phase system [25]. The SCs theory has been utilized as
a powerful tool in stability calculations, which principally permits the identification of any unbalances in the electrical power system [46, 47]. The detection procedure is more intuitive and robust compared to those methods which use parametric system modeling [21, 103, 104] or complex signal analysis [116–118]. The theory can be utilized to analyze the system degradation if there is an unbalanced condition arises inside the machine. However, the application of SCs theory to the five-phase system has been limitedly studied and therefore requires further analysis.
CHAPTER IV

MAXIMIZATION OF RELUCTANCE TORQUE UNDER OPEN PHASE FAULTS

In this chapter, an optimally designed F-PMa-SynRM [17] has been analyzed under different open phase fault conditions (Fig. 2.4). Under open-phase faults, the magnitude of the remaining phase currents is limited to an allowable range while developing the algorithm to reduce saturation effects. Then, the phase currents are balanced considering this limited magnitude. To further maximize the reluctance torque in F-PMa-SynRMs, different phase advances are proposed under different open phase faults.

4.1 Phase Current in Five-phase System

In a five-phase system, under the healthy condition, the phase currents can be derived as follows:

\[ I_{s\lambda} = I_{m\lambda} \sin \left( \theta - \sigma \frac{2\pi}{5} \right) \]

(4.1)

where, \( I_{m\lambda} \) is the phase current magnitude, \( \lambda \) is the phase number \((1, 2, 3, 4, 5)\), and \( \sigma \) is the integer \((0, 1, 2, 3, 4)\).
Under fault conditions, the zero sequence (ZS) current \( \left( \sum_{\lambda=1}^{5} I_{s\lambda} \neq 0 \right) \) becomes prominently large and the average torque reduces which denotes high unbalance in the system.

4.2 Space Vector Analysis under Faults

In this section, five-phase space vector analysis is presented in Fig. 4.1. The five arrowed lines (\( \vec{A}, \vec{B}, \vec{C}, \vec{D} \) and \( \vec{E} \)) define the direction of the currents in the windings in the form of vector Under normal condition, at a specific time, the currents in the five windings have the directions and values as shown in Fig. 4.1a. They could be summarized together using vector addition as \( \vec{I}_s = \vec{A} + \vec{B} + \vec{C} + \vec{D} + \vec{E} \). The current vector \( \vec{I}_s \) can be transformed into two-dimensional quantities as \( \vec{I}_{s\alpha} \) and \( \vec{I}_{s\beta} \) which are assumed as aligned towards the stationary reference axes, \( \alpha-\beta \) axis.

Fig. 4.1b shows the space vector orientation under single phase open fault. Under faults, the torque component \( \vec{I}_{s\alpha f} \) and the flux components \( \vec{I}_{s\beta f} \) reduces significantly hence reducing the total as \( \vec{I}_{sf} \). To maintain the torque, the amplitude of the remaining phase currents are increased 140\%, and the phase angles are also modified. Similarly, Fig. 4.1c and Fig. 4.1d show that the total current can be maintained same as a healthy condition by allowing 200-300\% rated current in the healthy phases. However, increasing the current in that amount in a PMa-SynRM may reduce the reluctance torque.
Figure 4.1: Space vector: (a) Healthy condition, (b) single phase open fault, (c) two adjacent phases (phase A and E) open fault and (d) two non-adjacent phases open fault (phase B and E).
4.3 Finite Element Modeling of F-PMa-SynRm under Faults with Saturation Effects

Under SPF, the FEA simulation (flux plot) for different current magnitude (100%, 150%, and 200%) has been done to see the saturation effect in the machine core and the rotor tooth. The results are shown in the Fig. 4.2. Comparing all, Fig. 4.2c shows the much saturation effect reaching more than 2.16 Tesla in many of the areas. This increment in the magnetic flux saturation would reduce the d-q axis inductances which are shown in Fig. 4.3.

Fundamentally the saliency ratio for the F-PMa-SynRM is the ratio of the d-axis inductance to the q axis inductance. The rated current of the machine is 15.2A (RMS). Considering the phase advance is 60 deg the d axis rated current is 10.7A and the q axis rated current is 18.6A. It is shown that, under 200% rated d-axis current (21.4 A), the saliency ratio reduces approximately 30% (saliency ratio 3.7 to 2.7), which significantly reduced the reluctance torque in F-PMa-SynRM. The cross saturation effect is noticeably prominent which further reduces the reluctance torque.

4.3.1 Introduction of Phase Advance with Phasor Diagram

Phase advance is the angle between the total phase current and the d axis which is also aligned to the reference axis. The advantages of the optimal phase advance on the torque production can be explained in a better way by using the phasor diagram. Before explaining the phasor diagram, it is important to show the transient equations of the five-phase PMa-SynRM which can be derived as follows:
\[ V_{dsf} = I_{dsf}R_s + L_{dsf}\frac{dI_{dsf}}{dt} - \omega_r(L_{qsf}I_{qsf} - \lambda_{PM}) \]
\[ V_{qsf} = I_{qsf}R_s + L_{qsf}\frac{dI_{qsf}}{dt} + \omega_r(L_{dsf}I_{dsf}) \]
\[ \lambda_{qsf} = I_{qsf}L_{qsf} - \lambda_{PM} \]
\[ \lambda_{dsf} = I_{dsf}L_{dsf} \]

where, \( \lambda_{dsf} \) and \( \lambda_{qsf} \) are the magnetic flux linkages in the d-axis and q-axis, and \( \lambda_{PM} \) is the permanent magnet flux linkage under open phase faults. From the mathematical model, the phasor diagram with phase advance(\( \gamma \)) is shown in Fig. 4.4. Fig. 4.4a shows the phasor diagram under healthy condition. In Fig. 4.4a, the Back-EMF (\( \omega_r\lambda_{PM} \)) is aligned towards the d-axis, \( \vec{V}_{ds} \) and \( \vec{V}_{qs} \) are the d-axis and q-axis voltages, \( L_{ds} \) and \( L_{qs} \) are the d-q axis inductances, \( \omega_r \) is the rotor speed, \( \lambda_{PM} \) is the permanent magnet flux, and \( R_s \) is the stator resistance.

Fig. 4.4b shows the phasor diagram under TPNF. It is shown that the current vector under faults, \( \vec{I}_{sf} \), is much smaller than the healthy condition current \( \vec{I}_s \) with a different phase advance (\( \gamma_{f1} \)). Under this condition, the phasor diagram shrinks with lower \( \vec{V}_{qsf} \). This lower \( \vec{V}_{qsf} \) leads to generate a low output torque. However, in this condition, the phase advance (\( \gamma_f \)) can be adjusted to generate more q axis voltage causing more q axis current to maintain higher saliency ratio. This higher saliency ratio during the fault, can generate higher reluctance torque.
Figure 4.2: Flux density distribution under single-phase open: (a) 100% rated condition, (b) 150% rated condition, and (c) 200% rated condition.
Figure 4.3: Inductances under single-phase open fault: (a) d-axis and (b) q-axis.
Figure 4.4: Phasor diagram: (a) Healthy condition (b) Phase B and E open.
4.4 Analysis of the Proposed Fault Tolerant Control Strategy for Five-Phase PMa-SynRM

This section analyzes the proposed FTC strategy along with optimal phase advance derivation for a PMa-SynRM under different open phase faults.

4.4.1 Conditions for Derivation of Optimal Phase Currents under Faults

The fundamental principles to derive the fault tolerant strategy are as follows:

1. the amplitude of the current remains within lesser saturation and sustainable range (it is assumed K% of the rated value in this paper)

2. ZS current of the healthy phase currents is zero

3. torque ripple is less than 10% and

4. maximum torque should be maintained during FTC.

Fig. 4.5 shows the optimization process by which the phase currents can be optimized under different open phase faults. In the optimization process, the first step is to limit the magnitude of the healthy phase currents under faults. The limit of the magnitude can vary for different machine types. This limit in the magnitude can be justified by utilizing finite element method that considers the nonlinearities properly inside the machine. To avoid the saturation effect for the machine utilized in this analysis, the limit on the phase currents magnitude (K) has been set to 150% of the rated value.
4.4.2 Derivation of Balanced Phase Currents under Faults Considering Limited Magnitude

This section provides an analytical derivation of the balanced phase currents under different open phase faults considering the system is running with limited magnitude ($K \leq 150\%$).

Single Phase Open Fault

The per unit phase currents under single phase open fault (phase A) become as follows:

\begin{align*}
I_{s1} &= 0 \\
I_{s\lambda} &= k_{1\lambda} \sin(\theta - (2\lambda - 2)\pi/5 - \phi_{1\lambda})
\end{align*}

(4.3)

where $\lambda$ is the phase number (2-5), $k_{1\lambda}$ are the limited magnitude of phase B, phase C, phase D and phase E, respectively, $\phi_{1\lambda}$ are the phase angle adjustment to maintain the balance in the system under SPF. To consider the current phasor symmetry, the phase angle of phase B and phase C can be kept constant as $\phi_{12} = 0$ and $\phi_{13} = 0$. 

Figure 4.5: Summary of the phase current optimization under different faults.
whereas the phase angle of phase D, \( \phi_{14} \) and phase E \( \phi_{15} \) should be changed. However, to satisfy the \( \sum I_s = 0 \) along with a limited current magnitude for all phases as \( K = 140\% \), the phase angle can be derived as follows:

\[
\begin{align*}
  k_{12} \sin(\theta - 2\pi/5) + k_{13} \sin(\theta - 4\pi/5) &= -k_{14} \sin(\theta - 6\pi/5 - \phi_{14}) - k_{15} \sin(\theta - 8\pi/5 - \phi_{15}) \\
  \phi_{14} &= \phi_{15} = \pi/5
\end{align*}
\]  

(4.4)

where \( k_{12} = k_{13} = k_{14} = k_{15} = K \).

Two-phase Adjacent Open Fault

The per unit phase currents under two-phase adjacent open fault (phase A and phase E) become as follows:

\[
\begin{align*}
  I_{s1} &= 0, I_{s5} = 0 \\
  I_{s\lambda} &= k_{2\lambda} \sin(\theta - (2\lambda - 2)\pi/5 - \phi_{2\lambda})
\end{align*}
\]  

(4.5)

where \( \lambda \) is the phase number \( (2-4) \), \( k_{2\lambda} \) are the limited magnitude of phase B, Phase C, and Phase D, respectively, \( \phi_{2\lambda} \) are the phase angle adjustment to maintain the balance in the system under TPAF.

In this condition, the phase angle of phase C can be kept constant \( (\phi_{23} = 0) \) whereas the phase angle of phase B and phase D can be changed. To satisfy the \( \sum I_s = 0 \) along with a limited current magnitude for phase B and phase D as \( K = 150\% \), the phase angle can be derived as follows:
\[ k_{22} \sin(\theta - 2\pi/5 - \phi_{22}) + k_{24} \sin(\theta - 6\pi/5 - \phi_{24}) = -k_{23} \sin(\theta - 4\pi/4) \]  

(4.6)

\[ \phi_{22} = -\phi_{24} = 0.654 \text{rad} \]

where \( k_{22} = k_{24} = K, \) \( k_{23} = 1, \) and \( \phi_{22} = -\phi_{24}. \)

Two-phase Non-adjacent Open Fault

The per unit phase currents under two-phase non-adjacent fault (phase B and phase E) become as follows:

\[ I_{s\lambda} = k_{2\lambda} \sin(\theta - (2\lambda - 2)\pi/5 - \bar{\phi}_{2\lambda}) \]

(4.7)

\[ I_{s2} = 0, I_{s5} = 0 \]

where \( \lambda \) is the phase number \((1, 3, 4)\) \( k_{2\lambda} \) are the limited magnitude of phase A, phase C, and phase D, respectively, \( \bar{\phi}_{2\lambda} \) are the phase angle adjustment to maintain the balance in the system under TPAF.

In this condition, phase A has been kept constant \((\bar{\phi}_{21} = 0)\) while the phase angle of phase C and phase D can be changed. To satisfy \( \sum I_s = 0 \) along with a limited current magnitudes, and considering \( k_{23} = k_{24} = K, \) \( \bar{k}_{21} = 1, \)and \( \bar{\phi}_{23} = -\bar{\phi}_{24}, \) the phase angle can be derived as follows:

\[ -k_{23} \sin(\theta - 4\pi/5 - \bar{\phi}_{23}) - k_{24} \sin(\theta - 6\pi/5 - \bar{\phi}_{24}) = k_{21} \sin(\theta) \]  

(4.8)

\[ \phi_{22} = -\phi_{24} = 0.60 \text{rad} \]
Figure 4.6: Balanced phase current for FTC under different faults: (a) single-phase, (b) two adjacent phase, and (c) two non-adjacent phase.

As discussed in (4.4) to (4.8), the balanced phase excitations are shown in Fig. 4.6a, Fig. 4.6b, and Fig. 4.6c under SPF (assuming phase $A = 0$), TPAF
4.4.3 Derivation of the Optimal Phase Advance under Different Faults

One of the critical parameters of F-PMa-SynRM is the phase advance ($\gamma$). The torque equation of a F-PMa-SynRM is given in (4.9). It can be further maximized by differentiating as $\frac{dT_e}{d\gamma} = 0$ and solving for $\gamma$ as follows:

$$T_e = \frac{NP}{4} [\lambda_{PM} I_s \cos \gamma + (L_{ds} - L_{qs}) I_s^2 \cos \gamma \sin \gamma]$$

$$\gamma = \sin^{-1}(-\lambda_{PM} + \frac{\sqrt{\lambda_{PM}^2 + 8(L_{ds} - L_{qs})^2 I_s^2}}{4(L_{ds} - L_{qs}) I_s^2})$$

where $N$ is the active number of phases (under healthy condition $N = 5$), $P$ is the number of poles, $I_s$ is the magnitude of current, $\gamma$ is the phase advance, $L_{ds}$ and $L_{qs}$ are the d and q axis inductances. In this paper, the phase advances under SPF, TPAF and TPNF are considered as $\gamma_{f1}$, $\gamma_{f2}$ and $\gamma_{f3}$.

Single Phase Open Fault

Under SPF, the balanced phase currents which are found in the section III-B-1 (Fig. 4.6a), can be transformed in the d-q axis frame by following two steps. The first step is to transform the phase current into the $\alpha - \beta$ axis frame as follows:

$$I_{s\alpha-SPF} = \sqrt{A_1^2 + B_1^2} \cos (\gamma_{f1} - \tan^{-1}(A_1/B_1))$$

$$I_{s\beta-SPF} = \sqrt{C_1^2 + D_1^2} \cos (\gamma_{f1} - \tan^{-1}(C_1/D_1))$$

(4.10)
Figure 4.7: Transformation under single phase open fault considering phase advance:
(a) phases to alpha-beta components, (b) alpha-beta to d-q axis components.
where,
\[
A_1 = (0.3I_{s2-new} - 0.8I_{s3-new} - 0.3I_{s4-new} + 0.6I_{s5-new})
\]
\[
B_1 = (-0.9I_{s2-new} - 0.6I_{s3-new} + 0.9I_{s4-new} + 0.8I_{s5-new})
\]
(4.11)
\[
C_1 = (1.2I_{s2-new} + 0.6I_{s3-new} - 0.9I_{s4-new} - 0.6I_{s5-new})
\]
\[
D_1 = (0.7I_{s2-new} - 0.8I_{s3-new} - 3I_{s4-new} + 0.8I_{s5-new})
\]

The \(\alpha - \beta\) transformation under SPF is shown graphically in Fig. 4.7a. In second step, the \(\alpha - \beta\) axis components are transformed into \(d - q\) axis components as follows:
\[
I_{ds-SPF} = I_{s\alpha-SPF} \cos(\theta) + I_{s\beta-SPF} \cos(\pi/2 - \theta)
\]
(4.12)
\[
I_{qs-SPF} = I_{s\alpha-SPF} \cos(\pi/2 + \theta) + I_{s\beta-SPF} \cos(\theta)
\]

The \(d - q\) transformation under SPF is shown graphically in Fig. 4.7b. From (4.10) and (4.12), the torque under SPF can be calculated as follows:
\[
T_{e-SPF} = \frac{N_p}{4} \mu_p M (E_1 \cos(\gamma_{f1} - \delta) \cos \theta + F_1 \cos(\gamma_{f1} - \epsilon_1) \sin \theta)
\]
\[
+ (L_{ds-SPF} - L_{qs-SPF})(-E_1^2 \cos^2(\gamma_{f1} - \delta) \cos \theta \sin \theta)
\]
\[
+ E_1 F_1 \cos(\gamma_{f1} - \delta_1) \cos(\gamma_{f1} - \epsilon_1)(\cos^2 \theta - \sin^2 \theta) + F_1 \cos^2(\gamma_{f1} - \epsilon_1) \sin \theta \cos \theta)
\]
(4.13)

where \(E_1 = \sqrt{A_1^2 + B_1^2}, F_1 = \sqrt{C_1^2 + D_1^2}, \delta_1 = \tan^{-1}(A_1/B_1), \) and \(\epsilon_1 = \tan^{-1}(C_1/D_1),\) respectively.

The torque under this SPF can be further maximized by differentiating as
\[
\frac{dT_{e-SPF}}{d\gamma_{f1}} = 0
\]
and solving for \(\gamma_{f1}\) as follows:
\[ G_1 \sin(\gamma f_1 - \delta_1) + H_1 \sin(\gamma f_1 - \epsilon_1) + 2I_1 \cos(\gamma f_1 - \delta_1) \times \sin(\gamma f_1 - \delta_1) + J_1 \cos(\gamma f_1 - \delta_1) = -\sin(\gamma f_1 - \epsilon_1) - J_1 \cos(\gamma f_1 - \epsilon_1) \sin(\gamma f_1 - \delta_1) \times -2K_1 \cos(\gamma f_1 - \epsilon_1) \times \sin(\gamma f_1 - \epsilon_1) \]

(4.14)

where,

\[ G_1 = \frac{NP}{4} \lambda_{PM} E_1 \cos \theta \]
\[ H_1 = \frac{5}{4} \lambda_{PM} F_1 \sin \theta \]
\[ I_1 = -\frac{5}{3}(L_{ds-SPF} - L_{qs-SPF}) E_1^2 \cos \theta \sin \theta \]
\[ J_1 = \frac{5}{4}(L_{ds-SPF} - L_{qs-SPF}) E_1 F_1 (\cos^2 \theta - \sin^2 \theta) \]
\[ K_1 = \frac{5}{4}(L_{ds-SPF} - L_{qs-SPF}) F \sin \theta \cos \theta \]

(4.15)

By solving (4.14) the optimal value of the \( \gamma f_1 \) is calculated.

Two-Phase Non-adjacent Open Fault

Under TPNF, the balanced phase currents which are found in section III-B-3 (Fig. 4.6b) are transformed into the \( \alpha - \beta \) axis components as follows:

\[ I_{s\alpha-TPNF} = \sqrt{(A_2)^2 + (B_2)^2} \cos(\gamma f_3 - \tan^{-1}(A_2/B_2)) \]

\[ I_{s\beta-TPNF} = \sqrt{(C_2)^2 + (D_2)^2} \cos(\gamma f_3 - \tan^{-1}(C_2/D_2)) \]

(4.16)

where,

\[ A_2 = (I_{s1-new} - 0.34I_{s3-new} - 0.36I_{s4-new}) \]
\[ B_2 = (0.87I_{s3-new} + 0.95I_{s4-new}) \]
\[ C_2 = (0.93I_{s3-new} - 0.47I_{s4-new}) \]
\[ D_2 = (I_{s1-new} + 0.34I_{s3-new} + 0.36I_{s4-new}) \]

(4.17)
Utilizing (4.16) and (4.17), the torque under TPNF can be calculated as follows:

\[
T_{e-TPNF} = \frac{NP}{4} \lambda_{PM} (E_2 \cos(\gamma_{f3} - \delta_2) \cos \theta + F_2 \cos(\gamma_{f3} - \epsilon_2) \sin \theta)
+ (L_{ds-TPNF} - L_{qs-TPNF})(- (E_2)^2 \cos^2(\gamma_{f3} - \delta_2) \cos \theta \sin \theta + E_2 F_2 \cos(\gamma_{f3} - \delta_2) \cos \theta)
- \delta_2) \cos(\gamma_{f3} - \epsilon_2)(\cos^2 \theta - \sin^2 \theta) + F_2 \cos^2(\gamma_{f3} - \epsilon_2) \sin \theta \cos \theta]
\]

(4.18)

where \( E_2 = \sqrt{A_2^2 + B_2^2} \), \( F_2 = \sqrt{C_2^2 + D_2^2} \),
\( \delta_2 = \tan^{-1}(A_2/B_2) \), and \( \epsilon_2 = \tan^{-1}(C_2/D_2) \), respectively.

The torque under can be further maximized by differentiating as \( \frac{dT_{e-TPNF}}{d\gamma_{f3}} = 0 \) and solving for \( \gamma_{f3} \) as follows:

\[
G_2 \sin(\gamma_{f3} - \delta_2) + H_2 \sin(\gamma_{f3} - \epsilon_2) + 2I_2 \cos(\gamma_{f3} - \delta_2) \sin(\gamma_{f3} - \delta_2)
+ J_2 \cos(\gamma_{f3} - \delta_2) \sin(\gamma_{f3} - \epsilon_2) = -J_2 \cos(\gamma_{f3} - \epsilon_2)
\]

(4.19)

\[
\sin(\gamma_{f3} - \delta_2) - 2K_2 \cos(\gamma_{f3} - \epsilon_2) \sin(\gamma_{f3} - \epsilon_2)
\]

where,

\[
G_2 = \frac{NP}{4} \lambda_{PM} E_2 \cos \theta
H_2 = \frac{5}{4} \lambda_{PM} F_2 \sin \theta
I_2 = -\frac{5}{4} (L_{ds-TPNF} - L_{qs-TPNF})E_2^2 \cos \theta \sin \theta
J_2 = \frac{5}{4} (L_{dsf} - L_{qsf})E_2 F_2 (\cos^2 \theta - \sin^2 \theta)
K_2 = \frac{5}{4} (L_{ds-TPNF} - L_{qs-TPNF}) F_2 \sin \theta \cos \theta
\]

(4.20)

By solving (4.19), the optimal value of the \( \gamma_{f3} \) is determined.
Two-Phase adjacent Open Fault

Under TPAF, the balanced phase currents which are found in section III-B-2 (Fig. 4.6c) are transformed into the $\alpha - \beta$ axis components as follows:

$$I_{sa-TPAF} = \sqrt{(A_3)^2 + (B_3)^2 \cos(\gamma f_2 - \tan^{-1}(A_3/B_3))}$$

(4.21)

$$I_{s\beta-TPAF} = \sqrt{(C_3)^2 + (D_3)^2 \cos(\gamma f_2 - \tan^{-1}(C_3/D_3))}$$

where,

$$A_3 = (0.83I_{s2-new} - 0.81I_{s4-new} - 0.28I_{s4-new})$$

$$B_3 = (-0.56I_{s2-new} - 0.59I_{s3-new} + 0.96I_{s4-new})$$

(4.22)

$$C_3 = (0.57I_{s2-new} + 0.59I_{s3-new} - 0.96I_{s4-new})$$

$$D_3 = (0.82I_{s2-new} - 0.81I_{s3-new} - 0.28I_{s4-new})$$

Utilizing (4.21) and (4.22), the torque under TPNF can be calculated as follows:

$$T_e - TPAF = \frac{NP}{4}[\lambda_{PM}(E_3 \cos(\gamma f_2 - \delta_3) \cos \theta + F_3 \cos(\gamma f_2 - \epsilon_3) \sin \theta)$$

$$+ (L_{ds-TPAF} - L_{qs-TPAF})(-E_3^2 \cos^2(\gamma f_2 - \delta_3) \cos \theta \sin \theta + E_3 F_3 \cos(\gamma f_2 - \delta_3)$$

$$\cos(\gamma f_2 - \epsilon_3)(\cos^2 \theta - \sin^2 \theta) + F_3 \cos^2(\gamma f_2 - \epsilon_3) \sin \theta \cos \theta)]$$

(4.23)

where $E_3 = \sqrt{A_3^2 + B_3^2}$, $F_3 = \sqrt{C_3^2 + D_3^2}$.

$\delta_3 = \tan^{-1}(A_3/B_3)$, and $\epsilon_3 = \tan^{-1}(C_3/D_3)$, respectively.

The torque under can be further maximized by differentiating as $\frac{dT_e - TPAF}{d\gamma f_2} = 0$

and solving for $\gamma f_2$ as follows:
By solving (4.24), the optimal value of the $\gamma_{f2}$ is determined.

$G_3 \sin(\gamma_{f2} - \delta_3) + H_3 \sin(\gamma_{f2} - \epsilon_3) + 2I_3 \cos(\gamma_{f2} - \delta_3) \sin(\gamma_{f2} - \delta_3)$

$+ J_3 \cos(\gamma_{f2} - \delta_3) \sin(\gamma_{f2} - \epsilon_3) = -J_3 \cos(\gamma_{f2} - \epsilon_3)$  \hspace{1cm} (4.24)

$\sin(\gamma_{f2} - \delta_3) - 2K_3 \cos(\gamma_{f2} - \epsilon_3) \sin(\gamma_{f2} - \epsilon_3)$

where,

$G_3 = \frac{N_P}{4} \lambda_{PM} E_3 \cos \theta$

$H_3 = \frac{5}{4} \lambda_{PM} F_3 \sin \theta$

$I_3 = -\frac{5}{4}(L_{ds-TPAF} - L_{qs-TPAF})E_3^2 \cos \theta \sin \theta$  \hspace{1cm} (4.25)

$J_3 = \frac{5}{4}(L_{ds-TPAF} - L_{qs-TPAF})E_3 F_3 (\cos^2 \theta - \sin^2 \theta)$

$K_3 = \frac{5}{4}(L_{ds-TPAF} - L_{qs-TPAF})F_3 \sin \theta \cos \theta$

By solving (4.24), the optimal value of the $\gamma_{f2}$ is determined.
4.4.4 Overall Control Strategy

The five-phase PMa-SynRM is energized using the five-phase inverter system. The five-phase currents are stored and checked for any faults in the motor phases utilizing fault detection methodologies [122]. Also, these currents are transformed into the two-dimensional quantities as $I_d$ and $I_q$. Using $I_d$ and $I_q$, the phase advance is calculated. To generate the maximum reluctance torque the phase advance is compensated from a look up table. After that, this optimal phase angle information is sent to the main controller. Also, the fault detection block provides the fault information to the main controller. Using optimal phase angles and fault categories information, the controller takes steps to generate optimized reference phase currents to the system. Fig. 4.8 describes the overall control strategy under different open phase fault condition.
CHAPTER V
MINIMIZATION OF TORQUE RIPPLE UNDER OPEN PHASE FAULTS

In this chapter, an active harmonics suppression in the phase currents is proposed to reduce the torque ripple while maintaining at least 88% average torque under a two-phase open fault condition. The algorithm is designed to control the unexpected phase harmonics which shows up for the higher non-linear saturation effects in the PMa-SynRM. Additionally, a correlation-based signal identification method has been derived from performing smart estimation of the phase current harmonics with high accuracy.

5.1 Back EMF and Phase Current

The fundamental voltage equation of a five-phase system is given as below:

\[
\begin{bmatrix}
  v_{abcde}
\end{bmatrix} = \begin{bmatrix}
  R_{abcde}
\end{bmatrix} \begin{bmatrix}
  i_{abcde}
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
  \lambda_{abcde}
\end{bmatrix}
\]

(5.1)

where \(v_{abcde}\), \(R_{abcde}\), \(i_{abcde}\), and \(\lambda_{abcde}\) are the phase voltage, resistance, currents and flux linkages, respectively. Usually, \(v_{abcde}\) is a sinusoidal voltage that is controlled from the input side. So any harmonic effect in the flux linkage directly affects the back-EMF voltage and results in harmonics in the phase currents. Therefore, the
back EMF is defined as follows:

\[ E_{s\lambda} = E_{1\lambda} \sin(2\pi ft - \alpha 2\pi /5) \]

\[ + \sum_{\nu=1}^{n} E_{(2\nu+1)\lambda} \sin((2\nu + 1)(2\pi ft - \alpha 2\pi /5)) \]  

(5.2)

where, \( E_{s\lambda} \) is the back EMF, \( E_{1\lambda} \) is the magnitude of the fundamental harmonic, \( E_{(2\nu+1)\lambda} \) is the magnitude of the harmonics, \( \lambda \) is the phase number, \( \alpha \) is the integer as 0 − 4 based on phase number, and \( \nu \) is the integer (1,2,3..n).

Here, the fundamental EMF signal is presented with only odd harmonics. Even harmonics are not expected due to the circular symmetry of the machine. Due to these back-EMF harmonics, the generalized phase currents can be presented from (5.1) as follows:

\[ I_{s\lambda} = I_{1\lambda} \sin(2\pi ft - \alpha 2\pi /5) \]

\[ + \sum_{\nu=1}^{n} I_{(2\nu+1)\lambda} \sin((2\nu + 1)(2\pi ft - \alpha 2\pi /5)) \]  

(5.3)

where \( I_{s\lambda} \) is the healthy phase currents, \( I_{1\lambda} \) is the magnitudes of the fundamental current, and \( I_{(2\nu+1)\lambda} \) are the magnitudes of the harmonics. The machine is designed in such way that under healthy condition the back EMFs (phase) contain the third harmonics (due to the magnet position in the rotor [47]). Under open phase faults, the magnitudes of these third harmonics vary dynamically and do not match each other. This paper focuses on controlling and suppressing their effects and improves the overall torque performances. In this paper, higher than third order harmonics are ignored as their effects are non-significant.
\[
\begin{bmatrix}
I_d \\
I_q \\
I_{dm} \\
I_{qm} \\
I_0
\end{bmatrix}
= \begin{bmatrix}
\mu_1 \cos \theta & \mu_2 \cos(\theta - \frac{2\pi}{5}) & \mu_3 \cos(\theta - \frac{4\pi}{5}) & \mu_4 \cos(\theta - \frac{6\pi}{5}) & \mu_5 \cos(\theta - \frac{8\pi}{5}) \\
\mu_1 \sin \theta & \mu_2 \sin(\theta - \frac{2\pi}{5}) & \mu_3 \sin(\theta - \frac{4\pi}{5}) & \mu_4 \sin(\theta - \frac{6\pi}{5}) & \mu_5 \sin(\theta - \frac{8\pi}{5}) \\
\frac{\mu_1}{5} & \frac{\mu_2}{5} & \frac{\mu_3}{5} & \frac{\mu_4}{5} & \frac{\mu_5}{5}
\end{bmatrix}
\begin{bmatrix}
I_{s1} \\
I_{s2} \\
I_{s3} \\
I_{s4} \\
I_{s5}
\end{bmatrix}
\]

5.2 $d - q$ Axis Equivalent Circuit

Conventionally, any synchronous machine is represented in synchronously rotating frame known as $d - q$ axes. For synchronous reluctance machine, the $d$-axis is aligned along the rotor magnetic flux, and the $q$-axis is orthogonal to $d$-axis [123]. Following this fundamental approach, the $d - q$ axes components in PMa-SynRM can be derived as follows:

\[
\begin{align*}
V_d &= R_a(I_d + \Delta i_d) + E_q + \Delta e_q \\
V_q &= R_a(I_q + \Delta i_q) + E_d + \Delta e_d \\
E_q + \Delta e_q &= -\omega_r(L_q(I_q + \Delta i_q) - \lambda_{PM}) \\
E_d + \Delta e_d &= \omega_r L_d(I_d + \Delta i_d)
\end{align*}
\]

where $R_a$ is the winding resistance, $V_d$, $I_d$, and $E_d$ are the $d$ axis input voltage, current and back-EMF which are contributions from the fundamental signals, $V_q$, $I_q$, and $E_q$ are the $q$ axis input voltage, current and back-EMF which are contributions from the fundamental signals, $\Delta i_d$ and $\Delta e_d$ are the current and back EMF in the $d$ axis due
Figure 5.1: $d - q$ axes equivalent circuit under fault condition: (a) $d$-axis, and (b) $q$-axis.

to the harmonic effects and $\Delta i_q$ and $\Delta e_q$ are the current and back EMF in the $q$ axis due to the harmonic effects, respectively, $L_d$ is the $d$-axis inductance, $L_q$ is the $q$-axis inductance, $\lambda_{PM}$ is the permanent magnet flux linkage. Utilizing the mathematical model in (5.5), the $d - q$ axes equivalent circuits of the F-PMa-SynRM is modeled as shown in Fig. 5.1. Therefore, the torque ripple can be derived as follows:

$$\Delta T_e = K([\lambda_{PM} + (L_d - L_q)I_q]\Delta i_d + [(L_d - L_q)I_d]\Delta i_q)$$  \hspace{1cm} (5.6)

$P$ is the number of poles, $K$ is $\frac{5P}{4}$. In (5.6), it is observed that the torque ripple largely depends on the $d - q$ axes currents ripple ($\Delta i_d, \Delta i_q$), which are mainly due to the harmonic effects. Therefore, by reducing the current ripple through harmonic suppression, the torque ripple can be reduced. Following that, the magnitude and phases of these harmonics are needed to identify properly. Detailed analysis of the
proposed torque ripple minimization and the harmonic identification is given in the following sections.

5.3 Torque Ripple Minimization under Open Phase Faults in a Five-phase PMa-SynRM

Fig. 5.2 shows the summarized steps to develop the proposed TRM method. Firstly, the third harmonic in the phase currents including the magnitude and phase are identified under different faults. Secondly, the injected signals are selected based on the percentile dominance of the harmonics. Finally, total phase currents (fundamental and third harmonic) are balanced before feeding the machine. Those steps sequentially explained in the following sections.

5.3.1 Step 1: Harmonic Identification in Phase Current

This section provides the harmonic analysis by analyzing the $d-q$ axis currents under fault conditions.
Normalized Estimation of the Phase Current Harmonics in $d - q$ axes

To estimate the normalized phase current harmonics (magnitude of the third harmonic/magnitude of the fundamental) in (5.3) can be presented as below

$$I_{s\lambda} = \sin(2\pi ft - \alpha 2\pi/5) + \rho_\lambda \sin(3(2\pi ft - \alpha 2\pi/5)) \quad (5.7)$$

where $\rho_\lambda$ is the normalized magnitude of the third harmonic.

The conventional two-dimensional transformation of the five-phase currents is derived in (5.4). In (5.4), the symbols are as follows: $\mu_1 = \mu_2 = \mu_3 = \mu_4 = \mu_5 = 1$ under healthy condition, otherwise ‘zero’ under each phase open fault condition, $I_{s\lambda}$ are the five-phase currents in (5.3), $m = (2\nu + 1)$ and $I_0$ zero sequence currents, $I_d$-$I_q$, and $I_{dm}$-$I_{qm}$ are the $d - q$ axes currents in the primary and secondary $d - q$ axes planes, respectively.

Assuming under SPF in phase A ($\mu_1 = 0$), utilizing the transformation matrix in (5.4) and the phase currents in (5.7), the $d - q$ axes currents under faults can be derived as follows:

$$I_d = \frac{2}{5} \sum_{\lambda=2}^{5} \cos(\theta - (2\lambda - 2)\pi/5) \times [\sin(\omega t - (2\lambda - 2)\pi/5) + \rho_\lambda \sin(3(\omega t - (2\lambda - 2)\pi/5))] \quad (5.8)$$

$$I_q = \frac{2}{5} \sum_{\lambda=2}^{5} \sin(\theta - (2\lambda - 2)\pi/5) \times [\sin(\omega t - (2\lambda - 2)\pi/5) + \rho_\lambda \sin(3(\omega t - (2\lambda - 2)\pi/5))] \quad (5.9)$$

By doing trigonometric summation in (5.8) and (5.9), these can be further expanded as follows:
\[ I_d = (0.3 \rho_5 - 0.3 \rho_2 - 1.1 \rho_3 + 1.1 \rho_4) + \cos 2\theta \]
\[ \times (0.5 \rho_5 - 0.5 \rho_2 - 3 \rho_3 + 3 \rho_4) + \sin 2\theta \]
\[ \times (-0.5 - 0.4 \rho_5 - 0.4 \rho_2 + 0.15 \rho_3 + 0.15 \rho_4) + \cos 4\theta \]
\[ \times (-0.5 \rho_5 + 0.5 \rho_2 + 3 \rho_3 - 3 \rho_4) + (\sin 4\theta) \]
\[ \times (0.15 \rho_5 + 0.15 \rho_2 - 0.4 \rho_3 - 0.4 \rho_4) \]
\[ I_q = (2 + 0.15 \rho_2 + 0.1 \rho_3 + 0.1 \rho_4 + 0.15 \rho_5) + (\cos 2\theta) \]
\[ \times (0.5 - 0.4 \rho_5 - 0.4 \rho_2 + 0.1 \rho_3 + 0.1 \rho_4) + (\sin 2\theta) \]
\[ \times (-0.3 \rho_5 + 0.3 \rho_2 - 0.45 \rho_3 + 0.45 \rho_4) - (\cos 4\theta) \]
\[ \times (0.4 \rho_5 + 0.4 \rho_2 - 0.15 \rho_3 - 0.15 \rho_4) - \sin 4\theta \]
\[ \times (0.5 \rho_5 - 0.5 \rho_2 - 0.3 \rho_3 + 0.3 \rho_4) \]

where \( \rho_2, \rho_3, \rho_4, \) and \( \rho_5 \) are the normalized magnitude of the third harmonics in phase B, phase C, phase D, and phase E, respectively. From (5.10) and (5.11) it is found that the currents in the \( d \)-axis and \( q \)-axis contain even harmonic under fault condition. Considering a phase difference in current harmonics, (5.10) and (5.11) can be further simplified as follows:

\[ I_d = \tilde{I}_d + \sum_{\kappa = 2 \nu(even)}^{2 \nu+2} \hat{i}_{d,1,\kappa} \cos(\kappa 2 \pi f t + \zeta_{d,\kappa,actual}) + \hat{i}_{d,2,\kappa} \sin(\kappa 2 \pi f t + \sigma_{d,\kappa,actual}) \]

\[ I_q = \tilde{I}_q + \sum_{\kappa = 2 \nu(even)}^{2 \nu+2} \hat{i}_{q,1,\kappa} \cos(\kappa 2 \pi f t + \zeta_{q,\kappa,actual}) + \hat{i}_{q,2,\kappa} \sin(\kappa 2 \pi f t + \sigma_{q,\kappa,actual}) \]

(5.12)
where $\bar{I}_d$ and $\bar{I}_q$ are the average currents, $i_{d,1,\kappa}$ and $i_{q,1,\kappa}$ are the magnitude of the harmonic currents in the $d$-axis and $q$-axis, $\zeta_{dqs,actual}$ and $\sigma_{dqs,actual}$ are unexpected actual phase differences, $\kappa$ is even numbered (in this case $\kappa = 2, 4$).

Magnitude Identification of the Harmonic

For faster and accurate identification the magnitude of the phase current harmonics, a correlation-based signal identifier (time domain) is proposed. For this purpose, the $d-q$ axis currents in (5.12) is multiplied with multiple reference signals to identify the normalized harmonic amplitudes. For example, the $d-q$ axes currents are multiplied by the following reference signals in (5.13), and the correlation output from these reference signals is calculated in (5.14).

$$I_{dq,1,\kappa \text{ref}} = [\cos(\kappa \theta - \zeta_{dqs,\text{predicted}})]$$

(5.13)

$$I_{dq,2,\kappa \text{ref}} = [\sin(\kappa \theta - \sigma_{dqs,\text{predicted}})]$$

$$X_{corr}I_{dq,1,\kappa} = \frac{T_f}{\int_0^f I_{dq} \times [\cos(\kappa \theta - \zeta_{dqs,\text{predicted}})]dt}$$

(5.14)

$$X_{corr}I_{dq,2,\kappa} = \frac{T_f}{\int_0^f I_{dq} \times [\sin(\kappa \theta - \sigma_{dqs,\text{predicted}})]dt}$$

where $T_f$ is the period of the fundamental, $\zeta_{dqs,\text{predicted}}$ and $\sigma_{dqs,\text{predicted}}$ are predicted phases of the $\kappa^{th}$ harmonic in the $d-q$ axes currents. The correlation between the reference and actual signals linearly varies from $-1$ to $1$ depending on the phase difference. The predicted phase ($\zeta_{dqs,\text{predicted}}$ or $\sigma_{dqs,\text{predicted}}$) is varied from 0 degree
to 180 degree, and subsequent correlation is estimated. The correlation becomes maximum when the predicted phase is equal to the actual phase. Fig. 5.3 shows a block diagram of the identification method.

From the output of the correlation-based signal identifier and (5.10), (5.11), (5.12) the per phase normalized magnitude of the phase current harmonics will be related as follows:

$$i_{dq,r,\kappa}^\wedge = f_{r\kappa}(\rho_1, \rho_2, \rho_3, \rho_4)$$

(5.15)

where $r = 1, 2$ and $\kappa = 2, 4$ (4 equations with 4 unknowns). By solving (5.15), the normalized magnitude of the harmonics $\rho_2$, $\rho_3$, $\rho_4$, and $\rho_5$ is calculated.
5.3.2 Step 2: Harmonic Injection under Open Phase Faults

The information from the harmonic identifier is utilized \((\rho_2-\rho_5)\) to make a decision on the harmonic signal injections. In this condition, two important factors are the normalized amplitude (including the sign) and the phase of the harmonic currents. To suppress the current harmonic effect, the amplitude of the injected current signals need to be the opposite magnitude to the identified signals. If \(\rho_2, \rho_3, \rho_4, \) and \(\rho_5\) are all positive quantity, then the injected signals can derived as follows:

\[
I_{s\lambda} = k_{1\lambda} \sin(2\pi ft - \alpha 2\pi/5) - \gamma_{3\lambda} \sin(3(2\pi ft - \alpha 2\pi/5)) \quad (5.16)
\]

where \(k_{1\lambda}\) are the per unit amplitude of fundamental component, \(\gamma_{3\lambda}\) is equal to \(k_{1\lambda}\rho_{(\lambda)}, \lambda = 2, 3, 4, 5\) and \(\alpha\) is the integer as \(0 - 4\), respectively for the open phase fault in phase A. By following the similar procedure, the identification of the harmonics and decision on the signal injection under TPAF can be done.

5.3.3 Step 3: Balancing Excitation Considering the Harmonic Injection

To maintain minimum zero sequence current, balanced excitation is needed even in the faults conditions. Following that, this section utilizes identified fundamental the harmonics in the phase current (from the previous section) and balances the phases to provide a balanced excitation under fault conditions.

Phase Current Balancing under Single-phase Lost

Under SPF, utilizing the identified fundamental and harmonics of the phase currents, the injected signal can be derived as follows:
Figure 5.4: Phase balancing under: (a) single-phase open fault in phase A, and (b) adjacent phase open fault in phase A and B.

\[ I_{s\lambda} = k_{1\lambda} \sin(\omega t - (2\lambda - 2)\pi/5 - \varphi_{1\lambda}) - \gamma_{3,1\lambda} \sin(3(\omega t - (2\lambda - 2)\pi/5 - \varphi_{1\lambda})) \]

(5.17)
where $\lambda$ is 2, 3, 4 and 5, $\varphi_{1\lambda}$ are the phase angle adjustment to maintain the balance in the system under SPF, and $\omega = 2\pi f t$.

Ideally, under SPF the zero sequence current for the fundamental component as well as for the harmonic currents are supposed to be zero as follows:

$$\sum_{\lambda=2}^{5} k_{1\lambda} \sin(\omega t - (2\lambda - 2)\pi/5 - \varphi_{1\lambda}) = 0$$

$$\sum_{\lambda=2}^{5} \gamma_{3,1\lambda} \sin(3(\omega t - (2\lambda - 2)\pi/5 - \varphi_{1\lambda})) = 0$$

(5.18)

To satisfy these balanced conditions, the modified magnitude and phases become as follows:

$$k_{12} = k_{13} = k_{14} = k_{15} = k_1$$

$$\gamma_{3,12} = \gamma_{3,13} = \gamma_{3,14} = \gamma_{3,15} = \gamma_{3,1}$$

$$\varphi_{14} = \varphi_{15} = \pi/6$$

$$\varphi_{12} = \varphi_{13} = -\pi/30$$

(5.19)

The adjustment in the phase current (fundamental and third components) is shown in the Fig. 5.4a.

Phase Current Balancing under Adjacent Two-phases Lost

Under TPAF, utilizing the identified fundamental and harmonics of the phase currents, the injected signal can be derived as follows:
\[ I_{s\lambda} = k_{2\lambda} \sin(\omega t - (2\lambda - 2)\pi/5 - \varphi_{2\lambda}) - \gamma_{3,2\lambda} \sin(3(\omega t - (2\lambda - 2)\pi/5) - \varphi_{2\lambda}) \]  
(5.20)

where \( \lambda \) is 3, 4 and 5, \( \varphi_{2\lambda} \) are the phase angle adjustment to maintain the balance in the system under TPAF.

Ideally, under TPAF the zero sequence current for the fundamental and harmonic currents are supposed to be zero as follows:

\[ \sum_{\lambda=3}^{5} k_{2\lambda} \sin(\omega t - (2\lambda - 2)\pi/5 - \varphi_{2\lambda}) = 0 \]
(5.21)

\[ \sum_{\lambda=3}^{5} \gamma_{3,2\lambda} \sin(3(\omega t - (2\lambda - 2)\pi/5 - \varphi_{2\lambda})) = 0 \]

To satisfy these balanced conditions, the modified magnitude and phases under TPAF become as follows:

\[ k_{23} = 1.62k_2, k_{24} = k_{25} = k_2 \]
\[ \gamma_{3,23} = 1.62\gamma_{3,2}, \gamma_{3,24} = \gamma_{3,25} = \gamma_{3,2} \]
(5.22)
\[ \varphi_{23} = -2\pi/5, \varphi_{24} = \varphi_{25} = 0, \]
\[ \varphi_{3,23} = -\varphi_{n,24} = -4\pi/5, \varphi_{3,25} = 0 \]

The adjustment in the phase currents (fundamental and third harmonics) are shown in the Fig. 5.4b.
5.3.4 Overall Control Strategy

The overall block diagram of the TRM technique is shown in Fig. 5.5. Under fault conditions, the remaining phase currents are initially stored to a fault detector which carries the information to a phase balancing block in the forward drive loop. Firstly, the remaining phase currents are transformed in the $d-q$ axes currents. As per the theoretical explanation, these currents contain third harmonic information. Then, these $d-q$ axes currents are forwarded in the current harmonic identification block. In this step, the amplitudes and phases of the third harmonic are measured. Using this information, the percentile harmonic magnitudes are chosen to get injected along with the fundamental signal. All these information (magnitude, phase, and order) together with the fault types, are fed to the phase balancing block or the main
controller. This controller generates the balanced excitation (both fundamental and third harmonic) which is utilized as reference phase currents. A current regulator (proportional-integral) block is utilized to control the phase currents. Finally, a five-phase voltage source inverter is utilized to produce the required input voltages.
CHAPTER VI
MAGNET TEMPERATURE ESTIMATION

This chapter proposes a dynamic frequency determination algorithm to be employed for magnet temperature estimation. In the existing methods for IPMSM and SPMSM [87, 88], various guidelines are proposed in determining the frequency of the injected signal for the magnet temperature estimation. Nevertheless, to estimate the magnet temperature, an absolute clean frequency determination is necessary. The proposed method isolates the fundamental and fault frequencies of the motor at a particular operating frequency to determine the available frequencies at which the signal can be injected. Then, the feedforward control method is developed to inject the chosen frequency with the fundamental signal. This method allows greater sensitivity, detectability, and accuracy for magnet temperature estimation.

6.1 Analysis of the Available Frequency Spectrum

For the motor drive application, the operating frequency spectrum can be constituted broadly in three categories as the fundamental operating frequency and the corresponding harmonics, the fault frequencies, and the noise frequencies. Mathematically, this spectral separation can be as follows:

\[ F_{\text{spectrum}} = F_{f,h} \cup F_{\text{fault}} \cup F_{\text{noise}} \]  

(6.1)
The fundamental operating frequencies and its corresponding harmonics are can be given by,

\[ F_{f,h} \in \{ f_s, 2f_s, 3f_s, ... \} \] (6.2)

where \( f_s \) is the fundamental operating frequency. To provide sufficient spectral separation from the dominant harmonics, the (6.2) can be further modified as follows:

\[ F_{f,h} \in \{ f_s \pm \delta f_s, 2(f_s \pm \delta f_s), 3(f_s \pm \delta f_s), ... \} \] (6.3)

where \( \delta f_s \) is called as the forbidden band and can be calculated under the assumptions of the cleaner frequency requirement.

6.2 Fault Frequencies

During the normal operation of the motor, faults developed either due to unintentional manufacturing defects or developed over time during the operation of the motor results in fault frequency components appearing in the motor current frequency spectrum. These faults are eccentricity faults and broken bar faults (in case of induction motor). In the case of eccentricity faults, the consequent fault frequencies are generated at a frequency given by [124],

\[ f_{\text{eccentricity}} = \left[ 1 \pm \left( \frac{2m - 1}{P} \right) \right] f_s \] (6.4)

where \( m \) is a positive integer corresponding and \( P \) is the number of pole pairs, and \( f_s \) is the fundamental excitation frequency. The fault frequency is observed not only
6.3 Noise

The remaining unclassified frequencies that are filtered after the isolation of the fundamental and the fault frequencies from the spectrum are termed as ‘noise’ frequencies. These frequencies represent groups of available frequencies at which no known external factor would generate a frequency component. The band of these noise frequencies is much larger than the band of the fundamental and harmonics. Therefore it covers wide frequency ranges. From this wide band of available noise frequencies, it is hard to choose a cleaner frequency for the temperature estimation. To process these frequencies, a random function generator tool is utilized. This tool arbitrarily generates a frequency and is checked with an SNR (signal to noise ratio) comparator. Sometimes the level of these noise signals is very high for several reasons including the unexpected motor vibration and unknown hardware malfunctions which further
makes the cleaner frequency identification more complex. For this reason, the SNR of these noise frequencies is calculated with an allowable threshold during the selection process. To ensure a cleaner frequency, it expected that frequency is well separated, and its magnitude is far less than the fundamental. If the SNR is higher than a preselected threshold than the frequency is passed to a sample frequency comparator otherwise the loop keep checking the frequencies. The later checklist is Nyquist sampling condition. The maximum bound comes from the Nyquist sampling theorem. The selected frequency which is to be injected for temperature estimation should be as follows:

\[ f_i < \frac{F_{\text{sam}}}{2} \quad (6.5) \]

where \( f_i \) is the selected frequency, \( F_{\text{sam}} \) is the sampling frequency.

6.4 Estimation Error Threshold

In conjunction of the previous conditions, an additional iterative loop that checks for convergence of magnet temperature estimation is added in the selection process which can potentially prevent the estimation errors in instantaneous implementations of the magnet temperature estimation. An overview of the flowchart of the proposed algorithm for frequency determination and it’s implementation is shown in the Fig. 6.1.
6.5 Magnet Temperature Estimation of a PMa-SynRM using Signal Injection Method

In this section, the method of the magnet temperature estimation using a signal injection is described mathematically. Furthermore, the control strategy considered for the temperature determination is thoroughly explained in the following sections.

6.5.1 Equivalent Model for the Signal Injection

Usually, the fundamental signal and their equivalent d-q axis components are considered in the control method. However, under additional signal injection in the machine, the mathematical construction of the additive components in the d-q axis model and their impact on the whole system require further analytical explanation.

For the additional signal injection, the voltage equations of the five-phase PMa-SynRM in the synchronous frame becomes as follows:

\[
\begin{bmatrix}
\Delta v_{df} \\
\Delta v_{qf}
\end{bmatrix} =
\begin{bmatrix}
R_{df} & 0 \\
0 & R_{qf}
\end{bmatrix}
\begin{bmatrix}
\Delta i_{df} \\
\Delta i_{qf}
\end{bmatrix} +
\frac{d}{dt}
\begin{bmatrix}
L_{df} & 0 \\
0 & L_{qf}
\end{bmatrix}
\begin{bmatrix}
\Delta i_{df} \\
\Delta i_{qf}
\end{bmatrix} +
\begin{bmatrix}
0 \\
\omega_f L_{qf}
\end{bmatrix}
\begin{bmatrix}
\Delta i_{df} \\
\Delta i_{qf}
\end{bmatrix} +
\begin{bmatrix}
\lambda_{PM} \\
0
\end{bmatrix}
\]

(6.6)

where, \(\Delta v_{df}\) and \(\Delta v_{qf}\), and \(\Delta i_{df}\) and \(\Delta i_{qf}\) are the d and q axes voltages and currents respectively for the additional frequency components. \(R_{df}\) and \(R_{qf}\), and \(L_{df}\) and \(L_{qf}\) are the d and q axes resistances and inductances respectively for the high frequency, \(\omega_f\) is the injected harmonic, \(\lambda_{PM}\) is the permanent magnet flux, and \(\frac{d}{dt} = j\omega\).
6.5.2 Control Strategy with the Signal Injection Method

The additional signal injection with the fundamental signal can be implemented in both current mode control which is also known as feedback mode or voltage mode control which is also known as feed-forward mode. Under feedback mode, the tuning of the PI (proportional integral) gains is cumbersome. For easier implementation and validation of the proposed method, feed-forward mode has been chosen to inject the additional signals to estimate the magnet temperature.

Feed-forward Control for the Signal Injection

Under feed-forward control, the input voltages are generated from the speed and torque command. Feed-forward control has well-known issues with steady-state error. To overcome the issue, proper parameters compensation (resistance and inductances) is required which vary with temperature. To inject the additional frequency for magnet temperature estimation, it is obvious to know the machine geometry including the magnet position inside the machine. In the present five-phase PMa-SynRM, the magnet position is in the q axis which is conventional for PMa-SynRM. Thus, it is presumed that signal associate in that q axis shall convey the temperature information.

Recalling from (6.6), under feed-forward control, the additional current components towards the d-axis can be set to zero which simplifies the model as follows:
\[
\begin{bmatrix}
\Delta v_{df} \\
\Delta v_{qf}
\end{bmatrix}
= \begin{bmatrix}
R_{df} & 0 \\
0 & R_{qf}
\end{bmatrix}
\begin{bmatrix}
\Delta i_{df} \\
\Delta i_{qf}
\end{bmatrix}
+ p
\begin{bmatrix}
L_{df} & 0 \\
0 & L_{qf}
\end{bmatrix}
\begin{bmatrix}
\Delta i_{df} \\
\Delta i_{qf}
\end{bmatrix}
+ \begin{bmatrix}
0 & -\omega_f L_{qf} \\
\omega_f L_{df} & 0
\end{bmatrix}
\begin{bmatrix}
\Delta i_{df} \\
\Delta i_{qf}
\end{bmatrix}
\] (6.7)

Parameter Compensation for the Signal Injection

Accurate parameter compensation is required to maintain the machine performances. For a particular operating condition, the machine parameters vary as follows:

\[R_s = f_R(T_c)\]
\[L_{dq} = f_{L_{dq}}(I_d, I_q)\] (6.8)
\[\lambda_{PM} = f_{\lambda_{PM}}(T_c, I_q)\]

Between these parameters, for a PMa-SynRM, the largest variation is seen for the d-q axis inductances which is shown in the Fig. 4.3.

Including the inductances, the resistance and permanent magnet flux variations can be stored in look-up tables (LUTs) which can be fed to the feed-forward controller as parameter compensation to estimate the magnet temperature properly.

6.5.3 Analysis on the Magnet Temperature for a Given Frequency

Proposing zero \(d\) axis current of the injected signal, the voltage signal from (6.7) becomes as follows:
\[
\begin{bmatrix}
\Delta v_{df} \\
\Delta v_{qf}
\end{bmatrix}
= \begin{bmatrix}
-\omega_f L_{qf} \Delta i_{qf} \\
R_{qf} \Delta i_{qf} + j\omega_f L_{qf} \Delta i_{qf}
\end{bmatrix}
\] (6.9)

Solving (6.9), the $R_{qf}$ at the injected frequency can be derived as follows,

\[
|R_{qf} + j\omega_f L_{qf}| = \left| \frac{\Delta v_{qf}}{\Delta i_{qf}} \right|
\] (6.10)

where $\omega_f L_{qf}$ is the inductive components of that high frequency. Particular attention should be given while selecting the high frequency. The much higher frequency may increase the inductive component significantly which is not expected in this situation.

In this dissertation, the inductance information is taken using finite element method.

The $R_{qf}$ consists of the stator resistance $R_{qs}$ ($R_s = R_{qs}$ under healthy condition) and the magnet resistance $R_{mag}$ which are each functions of temperature given by,

\[
R_{qf} = R_s(T_s) + R_{mag}(T_{mag})
\] (6.11)

where the stator resistance $R_s$ is obtained from the stator windings and is a function of the stator temperature $T_s$ and is given by,

\[
R_s(T_s) = R_{s0}(1 + \alpha_{Cu}(T_s - T_{s0}))
\] (6.12)

where $R_{s0}$ is the stator resistance at a known temperature $T_{s0}$, $\alpha_{Cu}$ is the temperature coefficient of resistivity of Copper. Similarly, the magnet temperature $R_{mag}$ is a function of the magnet temperature $T_{mag}$ and is given by,

\[
R_{mag}(T_{mag}) = R_{mag0}(1 + \alpha_{mag}(T_{mag} - T_{mag0}))
\] (6.13)
Substituting equations (6.12) and (6.13) back in equation (6.11) and rearranging the terms, we get,

$$T_{mag} = T_{mag0} + \frac{(R_{qf} - R_{sf0}(1 + \alpha_{Cu}(T_s - T_{s0}))) - R_{mag0}}{\alpha_{mag} R_{mag0}}$$

(6.14)

Therefore, by knowing the stator and magnet resistances $R_{sf0}$ and $R_{mag0}$ at the temperatures $T_{s0}$ and $T_{mag0}$ along with the stator temperature $T_s$, the magnet temperature can be determined from the determined $q$ axis resistance at the injected frequency $R_{qf}$ obtained from the commanded and measured signals.

In this section, the F-PMa-SynRM is controlled under feed-forward control method. The input voltages are generated from the speed and torque commands. The overall block diagram is shown in Fig. 6.2. A parameter compensation block is added to improve the machine performances therefore generating accurate magnet temperature information. The frequency determination algorithm (Fig. 6.1) is also
added in the control loop to inject most cleaner frequency to estimate the magnet temperature properly.
CHAPTER VII
OPEN PHASE FAULT DETECTION IN F-PMA-SYNRM

This chapter discusses the proposed fault detection method to identify the open phase lost in an F-PMa-SynRM. Firstly, the three-phase symmetrical components (SCs) theory has been mathematically extended to the five-phase system. Later, the five-phase SCs are utilized to identify the type of faults and the location of the faults. The proposed detection algorithm does not require any additional hardware to monitor the faults.

7.1 New Symmetrical Component Theory for F-PMa-SynRM under Faults

In this section, the SCs theory for a three-phase system has been extended to a five-phase system (without loss of generality).

The five-phase currents can be presented as follows:

\[ I_{sa} = I_{m1} \sin(2\pi ft) \]
\[ I_{sb} = I_{m2} \sin(2\pi ft - 2\pi/5) \]
\[ I_{sc} = I_{m3} \sin(2\pi ft - 4\pi/5) \]
\[ I_{sd} = I_{m4} \sin(2\pi ft - 6\pi/5) \]
\[ I_{se} = I_{m5} \sin(2\pi ft - 8\pi/5) \]

(7.1)
where \( I_{sa}, I_{sb}, I_{sc}, I_{sd}, I_{se} \) are the five phases currents, and \( I_{m1}, I_{m2}, I_{m3}, I_{m4}, I_{m5} \) are magnitudes of each phase currents, respectively.

These five current variables can be effectively analyzed for advanced fault detections in a five-phase machine. Fig.2.4 showed the possible open phase faults in a five-phase system including (i) SPF (phase A=0), (ii) TPAF (phases AE=0), and (iii) TPNF (phases BE=0). These faults can be effectively detected using a modified SCs theory. Detailed analysis is done in the following sections.

7.2 Derivation of Symmetrical Components for a Five-phase system

Fundamentally, the number of phase sequences in a three-phase system is three which are positive sequence (PS), negative sequence (NS), and zero sequence (ZS) components. In a five-phase system, the number of phase sequence is significantly increased due to larger phase combinations. For example, based on the phase rotation, a three-phase system has only two combinations (2x1). However, there are 24 (=4x3x2x1) possible combinations of phase sequences in a five-phase system.

To identify the proper phase sequences, each out of 24 sequences has been evaluated based on ZS component. If the ZS component becomes ideally zero, the sequence is selected as one of phase sequence in a five-phase system. This has led to the identification of four new sequences in five phase system which are positive sequence 1 (PS1) \((A \rightarrow B \rightarrow C \rightarrow D \rightarrow E)\), negative sequence 1 (NS1) \((A \rightarrow E \rightarrow D \rightarrow C \rightarrow B)\), positive sequence 2 (PS2) \((A \rightarrow C \rightarrow E \rightarrow B \rightarrow D)\), negative sequence 2 (NS2) \((A \rightarrow D \rightarrow B \rightarrow E \rightarrow C)\). These-phase sequences are depicted in
Figure 7.1: Proposed symmetrical components for five-phase system.

phasor diagram in Fig. 7.1. These four sequences along with ZS makes a set of five SCs for a five-phase system.

Base on identified five phase sequences, the currents of a five-phase system are decomposed as follows:

\[
\begin{align*}
I_{sa} &= \sum_{n=0}^{4} I_{an} \\
I_{sb} &= \sum_{n=0}^{4} I_{bn} = I_{a0} + I_{a1}\alpha^4 + I_{a2}\alpha^1 + I_{a3}\alpha^2 + I_{a4}\alpha^3 \\
I_{sc} &= \sum_{n=0}^{4} I_{cn} = I_{a0} + I_{a1}\alpha^3 + I_{a2}\alpha^2 + I_{a3}\alpha^4 + I_{a4}\alpha^1 \\
I_{sd} &= \sum_{n=0}^{4} I_{dn} = I_{a0} + I_{a1}\alpha^2 + I_{a2}\alpha^3 + I_{a3}\alpha^1 + I_{a4}\alpha^4 \\
I_{se} &= \sum_{n=0}^{4} I_{en} = I_{a0} + I_{a1}\alpha^1 + I_{a2}\alpha^4 + I_{a3}\alpha^3 + I_{a4}\alpha^2
\end{align*}
\]  \quad (7.2)

where \( I_{an}, I_{bn}, I_{cn}, I_{dn}, I_{en} \) (n=0,1,2,3 and 4) are the SCs of phase A, phase B, phase
C, phase D, and phase E, respectively, and for phase A, the SCs are given as $I_{a0}$ (ZS), $I_{a1}$ (PS1), $I_{a2}$ (NS1), $I_{a3}$ (PS2), $I_{a4}$ and (NS2), and $\alpha = 72^\circ$).

Using (2), the five SCs can be found utilizing the transformation matrix ($T$) as shown in (7.3).

\[
\begin{bmatrix}
I_{a0} \\
I_{a1} \\
I_{a2} \\
I_{a3} \\
I_{a4}
\end{bmatrix} = \frac{1}{5}
\begin{bmatrix}
1 & 1 & 1 & 1 & 1 \\
1 & \alpha_1 & \alpha_2 & \alpha_3 & \alpha_4 \\
1 & \alpha^4 & \alpha^3 & \alpha^2 & \alpha^1 \\
1 & \alpha^3 & \alpha^1 & \alpha^4 & \alpha^2 \\
1 & \alpha^2 & \alpha^4 & \alpha^1 & \alpha^3
\end{bmatrix}
\begin{bmatrix}
I_{sa} \\
I_{sb} \\
I_{sc} \\
I_{sd} \\
I_{se}
\end{bmatrix} = \begin{bmatrix} 0 \\
I_{m1} \sin(2\pi ft) \\
0 \\
0 \\
0 \end{bmatrix}
\]

(7.3)

In (7.3), the transformation matrix has been successfully derived for a five-phase system with the proposed SCs. In (7.3), only the PS1 component takes non-zero values, whereas all other components take as zero values under healthy condition which satisfies the SCs theory.

### 7.3 Symmetrical Components under Fault Conditions

This section provides the analysis of SCs under open phase faults (SPF, TPAF, and TPNF). The Magnitudes and phases of the SCs are analyzed under different fault conditions to identify its pattern.

When SPF occurs ($I_{sa} = 0$) the five SCs in (7.3) can be driven as follows:
\[ I_{a0} = K_{01} \sin(2\pi ft - \arctan(B_{01}/A_{01})) \]
\[ I_{a1} = K_{11} \sin(2\pi ft) \]
\[ I_{a2} = K_{21} \sin(2\pi ft - \arctan(B_{21}/A_{21})) \]
\[ I_{a3} = K_{31} \sin(2\pi ft - \arctan(B_{31}/A_{31})) \]
\[ I_{a4} = K_{41} \sin(2\pi ft - \arctan(B_{41}/A_{41})) \]

When TPAF occurs (\( I_{sa} = I_{sb} = 0 \)), the five SCs in (7.3) can be driven as follows:

\[ I_{a0} = K_{02} \sin(2\pi ft - \arctan(B_{02}/A_{02})) \]
\[ I_{a1} = K_{12} \sin(2\pi ft) \]
\[ I_{a2} = K_{22} \sin(2\pi ft - \arctan(B_{22}/A_{22})) \]
\[ I_{a3} = K_{32} \sin(2\pi ft - \arctan(B_{32}/A_{32})) \]
\[ I_{a4} = K_{42} \sin(2\pi ft - \arctan(B_{42}/A_{42})) \]

where,
\[ K_{n1} = \sum_{n=0, n \neq 1}^{4} I_{mn} \sqrt{\left(A_{n1}^2 + B_{n1}^2\right)} \frac{1}{C_{n1}} \text{ for } n = 1, C_{n1} = D_{n1} = 1 \]
\[ K_{n2} = \sum_{n=0, n \neq 1}^{4} I_{mn} \sqrt{\left(A_{n2}^2 + B_{n2}^2\right)} \frac{1}{D_{n1}} \text{ for } n \neq 1, C_{n1} \neq D_{n1} \neq 1 \]

\( K_{n1} \) and \( K_{n2} \) are amplitudes of SCs under SPF and TPAF, \( I_{mn} \) is the magnitude as in (7.1), \( \arctan(B_{n1}/A_{n1}) \) and \( \arctan(B_{n2}/A_{n2}) \) are the phase angles of the other SCs in reference to the PS1, \( B_{n1} \) and \( A_{n1} \) are the summation of \( \sin \sigma \beta \) and \( \cos \sigma \beta \), respectively (\( \sigma = 0, 1, 2..., \beta = 2\pi/5 \)).
Through (7.4) and (7.5), it has been comprehensively analyzed that, the magnitudes and phase angles of the SCs under deferent faults substantially changes with the mathematical pattern. This can be effectively utilized to identify the type of faults and location of the faults.

7.4 Analysis of the Symmetrical Components - Amplitude

In this section, the magnitude of SCs will be utilized for the detection of fault types. To comparatively analyze the signal magnitude between SCs, two signal ratio indexes \((r_1, r_2)\) have been proposed which is defined as follows:

\[
\begin{align*}
\frac{r_1}{K_3} &= \frac{K_3}{K_0}, \\
\frac{r_2}{K_4} &= \frac{K_2}{K_4}
\end{align*}
\]

(7.7)

where \(K_0, K_2, K_3,\) and \(K_4\) are the peak of ZS \((I_0)\), NS1 \((I_2)\), PS2 \((I_3)\) and NS2 \((I_4)\), respectively.

Under SPF and TPAF in (7.4) and (7.5), the changes of \(r_1\) and \(r_2\) have been presented in Fig. 7.2. In Fig. 7.2 the magnitude indexes \(r_1\) and \(r_2\) are equal to one under any SPF (phases A, B, C, D, E). The indexes, \(r_1\) and \(r_2\), become lower than one under any TPAF (phases AB, BC, CD, DE, and EA). Similarly, under any TPNF (phases AC, AD, BD, BE, CE) the indexes become higher than one. It can be seen that the indexes \(r_1\) and \(r_2\) clearly follow a pattern which is providing the information of the types of faults as follows:
Figure 7.2: Proposed signal ratio (Magnitude index) under all possible open phase faults.

\[
\begin{align*}
    r_1, r_2 < 1 &; \quad \text{Two-phase adjacent fault} \\
    r_1, r_2 = 1 &; \quad \text{Single-phase fault} \\
    r_1, r_2 > 1 &; \quad \text{Two-phase non-adjacent fault}
\end{align*}
\]  

7.5 Analysis of the Symmetrical Components - Phase

In this section, the phase of SCs will be utilized for detecting the fault locations. The phase angle changes of SCs under faults also have valuable information on the location of faults. Fig. 7.3 shows the MATLAB simulation results of phase angle changes under different fault conditions. In Fig. 7.3, leading phase (w.r.t. PS1) is colored as red and lagging phase (w.r.t. PS1) is colored as blue. Other than, in-phase or out of phase (w.r.t. PS1) is colored as green. Fig. 7.3 shows that, the change of
phases of those sequences show unique pattern under different fault condition (SPF, TPAF, and TPNF).

7.6 Overall Detection Scheme

The overall block diagram of the detection procedure is shown in Fig. 7.4. The motor is run with field orient control strategy at maximum torque per ampere condition. Under fault conditions, the five-phase currents are initially fed to a low pass filter to remove higher order harmonics. Only the fundamental current components are
considered during the detection procedure. The filtered phase signals are multiplied by the five-phase transformation matrix \( T \) as shown in (7.3) to identify five SCs. The magnitudes and phases of the five SCs have been measured. Utilizing this magnitude and phase information, the magnitude indexes and phase pattern have been identified (7.7). Finally, the type and location of the open phase fault have been effectively identified by using (7.8) and Fig. 7.3. This fault information can be sent to the motor controller to activate fault tolerant control strategy [125].
CHAPTER VIII
SIMULATION RESULTS

This chapter presents the performances validation of the proposed methods through the simulation results. These simulations have been done through the finite element method and the MATLAB/SIMULINK. For this purpose, a finite element analysis (FEA) model of the F-PMa-SynRM is developed. The simulation results are provided orderly as follows: 1) maximization of reluctance torque under open phase faults, 2) minimization of torque ripple under open phase faults, 3) fault detection in F-PMa-SynRM.

8.1 Finite Element Model utilized in the Simulation

The FEA model is shown in Fig. 8.1. Fig. 8.1a shows the FEA model of the stator. Fig. 8.1b shows the FEA model of the rotor.

In the FEA model of the F-PMa-SynRM, the fractional slotting configuration is utilized which is shown in the Fig. 8.2. It shows, there are fifteen slots with four pole configuration which makes it fractionally configured \((15/4 = 3.75)\) F-PMa-SynRM.

The five-phase PMa-SynRM model (F-PMa-SynRM) has been developed based on the specification given in Table. 8.1. It is designed to operate at sixty
Figure 8.1: FEA model of the F-PMa-SynRM: (a) Stator, and (b) rotor.

Figure 8.2: Slotting configuration of the F-PMa-SynRM

hertz with $3 \text{ kW}$ output power. The rated current and torque are 15.17 ampere and 15.2 Nm, respectively. Each phase has identical resistance as 0.3 ohm with twenty
two conductors. The detail design and optimization procedure of the F-PMa-SynRM is explained in [17].

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specifications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of slot/poles</td>
<td>15/4</td>
</tr>
<tr>
<td>Rated current (rms) (A)</td>
<td>15.17</td>
</tr>
<tr>
<td>Rated voltage (rms) (V)</td>
<td>67</td>
</tr>
<tr>
<td>Power (kW)</td>
<td>3</td>
</tr>
<tr>
<td>Rated speed (rpm)</td>
<td>1800</td>
</tr>
<tr>
<td>Rated Torque (Nm)</td>
<td>15</td>
</tr>
<tr>
<td>Phase resistance (ohm)</td>
<td>0.3</td>
</tr>
<tr>
<td>Rated d-axis inductance (mH)</td>
<td>10</td>
</tr>
<tr>
<td>Rated q-axis inductance (mH)</td>
<td>2.8</td>
</tr>
<tr>
<td>Phases</td>
<td>5</td>
</tr>
</tbody>
</table>
8.2 Results for Maximization of Reluctance Torque under Open Phase Faults

In this section, the finite element simulation results are provided for maximization of the reluctance torque under faults.

8.2.1 Torque under Balanced Phase Currents under Faults- FEA

The torque with the balanced phase currents (calculated in chapter 4) has been found with finite element method in Fig. 8.3. Fig. 8.3a shows the torque variation under SPF (phase $A = 0$). The calculated phase currents (magnitude and phase angles) have been feed to the five-phase PMa-SynRM, and the torque is found as 0.33 p.u. ($= 4.95 \text{Nm, actual value } = \text{p.u. value } \times \text{ rated value}$). Fig. 8.3b shows the torque under TPAF (phases $AE = 0$). The calculated phase currents (magnitude and phase angles) have been fed to the machine, and the torque is found as 0.3 p.u. Similarly, under TPNF (phases $BE = 0$), utilizing the calculated phase currents the torque is found as 0.24 p.u.

8.2.2 Optimal Phase Advance for Maximum Toque under Different Faults- FEA

In this section, the simulation is done to calculate the torque under different open phase faults by considering the adjustment of the phase advance ($\gamma_f$) which has been discussed in chapter 4. Fig. 8.4 shows that the torque can be maximized by introducing the optimal phase advance under different faults.

The results for SPF, TPAF, and TPNF conditions are given respectively in Fig. 8.4a, Fig. 8.4b and Fig. 8.4c. Under SPF, Fig. 8.4a shows the maximum
Figure 8.3: Torque under balanced phase currents considering limited magnitude:
(a) single-phase open, (b) two adjacent phase open, and (c) two non-adjacent phase open.

torques are within $\sim 0.25\text{-}0.55 \text{ p.u.}$ for a phase advance change in $\sim 42\text{-}55$ degree under different loading situations. Under TPAF, Fig. 8.4b shows the maximum torques are within $\sim 0.3\text{-}0.51 \text{ p.u.}$ for a phase advance change in $\sim 39\text{-}50$ degree under different loading situations. Under TPNF, Fig. 8.4c shows the maximum torques are within $\sim 0.28\text{-}0.47 \text{ p.u.}$ for a phase advance change in $\sim 40\text{-}50$ degree under different loading situations.
8.3 Results for Minimization of Torque Ripple under Open Phase Faults

This section provides the finite element simulation results of the harmonic phenomenon under faults and the proposed harmonic injection method in the phase currents under SPF and TPAF conditions. To guarantee the safety during the test under faults, the machine has been operated at low speed with different load conditions. For this reason, the operating speed is considered as 9 Hz during the simulation.
Figure 8.5: Flux density: (a) under healthy condition, and (b) under TPAF.

8.3.1 Harmonic Phenomenon under Open Phase Faults

The harmonic phenomenon under faults is presented in the Fig. 8.5-Fig. 8.8. Fig. 8.5a shows the flux density under healthy condition. It is observed, under the healthy
Figure 8.6: Flux linkage: (a) under healthy condition, and (b) under TPAF.

condition, the field vectors travel the rotor core uniformly generating a uniform back EMF in the phases which are shown in Fig. 8.8a. Fig. 8.5b shows the flux density under TPAF. It is observed that under the faulty condition, the flux density in the rotor core is not uniform generating non-uniform back EMF in the phases. This back-EMF results in different current harmonics in the phases. Fig. 8.6a and Fig. 8.6b show the flux linkages under healthy and TPAF conditions. The harmonics in the phase C flux linkage is shown in the Fig. 8.7. It is observed, the third harmonic
becomes stronger during the fault. The back EMF under SPF is shown in the Fig. 8.8b which shows similar harmonic effects during a fault.

8.3.2 Harmonic Injection in the Phase Currents under Open Phase Faults

Fig. 8.9 shows the phase currents with the injected harmonics under SPF and TPAF conditions. Fig. 8.9c shows the spectral analysis of the balanced phase current (phase B) with the third harmonic which is 46\% (calculated from (5.15) to (5.19)) of the fundamental is injected under SPF condition. Also, the phase currents are adjusted as $\varphi_{12} = \varphi_{13} = -6^\circ$ and $\varphi_{14} = \varphi_{15} = 30^\circ$ as per (5.19) which is shown in Fig. 8.9a. Similarly, a balanced third harmonic is injected with a magnitude of 20\% of the fundamental current (calculated from (5.15) to (5.22)) under TPAF condition which is shown in Fig. 8.9d. Also as per (5.22), the phase currents are adjusted as $\varphi_{23} = -72^\circ$ and $\varphi_{3,23} = -\varphi_n,24 = -4\pi/5$ which is shown in Fig. 8.9b.
Figure 8.8: Induced EMF: (a) under healthy condition, and (b) under SPF.

8.3.3 Torque with the Proposed Method under Open Phase Faults

The results of this harmonic injection in the phase current are shown in the torque results in Fig. 8.10a and Fig. 8.10b. Fig. 8.10a shows the torque under healthy and SPF condition (phase A=0). It is observed that under SPF condition, the ripple in the torque increases to 22.6%. By injecting those harmonic along with fundamental signals in Fig. 8.9a, the torque ripple is reduced to 11.3%. Fig. 8.10b shows torque under two-phase adjacent open fault (phase A=E=0). The result shows that torque
Figure 8.9: Balanced phase current with harmonic injection: (a) phase currents (B, C, D, E) under SPF at phase A, (b) phase currents (C, D, E) under TPAF in phase A and B, (c) FFT of phase B current under SPF at phase A, and (d) FFT of phase D current under TPAF in phase A and B.

Ripple under the fault increases to 25.1%. By injecting harmonic with the fundamental in the phase current in Fig. 8.9b, the torque ripple is reduced to 10.6%.
8.3.4 Saturation Effects under Open Phase Faults

Under faults, the machine has been simulated at rated condition to observed the saturation effect. Fig. 8.11 shows the flux density distribution under SPF and TPAF. In Fig. 8.11a, the flux density is shown under SPF with and without the proposed method. It is shown that the saturation effects in the stator and rotor sides are less with the proposed method. A similar effect is observed for TPAF in the Fig. 8.11b. This reduced saturation effect does hold the initial assumption of not getting the direct and quadrature axes inductances in the saturation region.

Figure 8.10: Torque: (a) Healthy and single-phase open fault in phase A, and (b) adjacent phase open fault in phase A and B.
8.4 Results for Fault Detection in F-PMa-SynRM

Finite element analysis (FEA) has been performed to prove the performance of the proposed method. The five-phase machine in Table 8.1, has been utilized under the
tests. The simulation has been performed under different fault conditions, including SPF, TPAF, and TPNF. Detailed finite element analysis (FEA) has been done to the motor in Table 8.1 to observe the pattern identified in Fig. 7.2 and Fig. 7.3 under non-linear operating conditions of the machine.

8.4.1 Phase Currents under Different Open Phase Faults

Fig. 8.12 shows the phase currents (A, B and D) under TPNF (C and E). It can be seen that the phase B current becomes substantially distorted causing THD as 37%. The fundamental component which is shown in Fig. 8.12 is obtained through filtering to evaluate the comparative analysis. The per unit (p.u.) values of the fundamental currents of the phases A, B, and D are 0.79, 0.49, and 0.90, respectively. The Table

<table>
<thead>
<tr>
<th>Operating Conditions</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
<th>E</th>
</tr>
</thead>
<tbody>
<tr>
<td>Healthy Condition</td>
<td>0.8</td>
<td>0.8</td>
<td>0.8</td>
<td>0.8</td>
<td>0.8</td>
</tr>
<tr>
<td>E Open</td>
<td>0.74</td>
<td>0.55</td>
<td>0.88</td>
<td>0.88</td>
<td>0</td>
</tr>
<tr>
<td>D and E Open</td>
<td>0.89</td>
<td>0.54</td>
<td>0.89</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>C and E open</td>
<td>0.79</td>
<td>0.49</td>
<td>0</td>
<td>0.90</td>
<td>0</td>
</tr>
</tbody>
</table>
8.2 shows the summary of p.u. values of the fundamental current under healthy and other possible fault conditions.

The comparison of the torque and torque ripple under different faults is given in Fig. 8.13. Fig. 8.13 shows the average torques are 2 Nm, 1.88 Nm, 1.85 Nm and 1.65 Nm and the torque ripples are 10%, 26%, 29%, and 31% under healthy, SPF, TPAF, and TPNF, respectively.

8.4.2 Analysis of the Symmetrical Components - Amplitude

Using the proposed transformation matrix \((T)\), five SCs are shown in Fig. 8.14 and Fig. 8.16 for 100% and 30% rated conditions, respectively. Fig. 8.14a shows the SCs under healthy conditions. It is observed in Fig. 8.14a that all the SCs (ZS, NS1, PS2,
NS2) become zero except PS1.

Fig. 8.14b shows the SCs under SPF (phase E=0) condition. The magnitudes of ZS, NS1, PS2, and NS2 are 0.3, 0.5, 0.30, 0.29 and 0.31, respectively.

Fig. 8.14c shows the SCs under TPAF (phase D and E). The magnitude of the ZS, NS1, PS2 and NS2 are 0.25, 0.23, 0.17, and 0.36, respectively.

Fig. 8.14d shows the SCs under TPNF (Phase C and E). The magnitude of the ZS, NS1, PS2 and NS2 are 0.19, 0.37, 0.4, and 0.18, respectively.

Fig. 8.15 summarizes the magnitude index ($r_1$ and $r_2$) under all possible open phase faults, including SPF, TPAF, and TPNF. Fig. 8.15 shows that, under SPF (phase A or B or C or D or E), the magnitude indexes, $r_1$ and $r_2$, remain close to one...
Figure 8.14: SCs at 100% rated: (a) Healthy condition, (b) phase E=0, (c) phase DE=0, (d) phase CE=0.

(= 1). Under TPAF (phases AB or BC or CD or DE or EA), the magnitude indexes, $r_1$ and $r_2$, remain lower than one ($1 <$). Under TPNF (phases AC or AD or BD or BE or CE), the magnitude indexes, $r_1$ and $r_2$, remain higher than one ($1 >$).

Fig. 8.16a shows the SCs under SPF under 30% load. The magnitude of the ZS, NS1, PS2 and NS2 are 0.10, 0.086, 0.098, and 0.09, respectively. From these values, the magnitude indexes are calculated as $r_1=.98$ and $r_2=.96$. Fig. 8.16b shows the SCs under TPAF. Under this fault, the magnitude of the ZS, NS1, PS2, and
NS2 are 0.12, 0.025, 0.037, and 0.09, respectively. From these values, the magnitude indexes are calculated as $r_1=0.33$ and $r_2=0.28$. Similarly, Fig. 8.16c shows the SCs under TPNF and the magnitude of the ZS, NS1, PS2, and NS2 are 0.04, 0.075, 0.10, and 0.04, respectively. From these values, the magnitude indexes are calculated as $r_1=2.52$ and $r_2=1.89$.

8.4.3 Analysis of the Symmetrical Components - Phase

Fig. 8.17 summarizes phase angles under SPF (phase A or B or C or D or E), TPAF (phases AB or BC or CD or DE or EA), and TPNF (phases AC, AD, BD, BE, CE).

In Fig. 8.17, leading phase with respect to positive sequence 1 is colored as red, lagging phase with respect to PS1 is colored as light blue, and in-phase or out of phase with respect to PS1 is colored as green. The phase information has been analyzed to figure out the exclusive combination of the phases to specify the location.
of faults. It can be observed that change of phase shows unique pattern under each fault condition (SPF, TPAF, and TPNF). For example, when SPF occurs in phase E, change of phase shows unique pattern which differs from other patterns for SPF at phase A, B, C, D. Similarly, distinctive patterns have been found for TPAF and TPNF which eventually provide the information about the fault locations in all conditions as described in Fig. 7.3.
8.4.4 Effectiveness of the proposed Method under Unbalanced Resistances fault

The performance of the proposed algorithm has been verified under different types of fault conditions. For example, the algorithm has been evaluated under unbalanced resistance fault in phases. The results are provided in Fig. 8.18. In Fig. 8.18a, it has been shown that the magnitude indexes are $r_1=0.51$ and $r_2=0.19$ which are less than one under unbalanced resistances fault in adjacent phases (phase A and B) with a phase angle pattern as discussed in chapter 7. In Fig. 8.18b, it is shown that...
Figure 8.18: Unbalanced resistance in: (a) adjacent phases: Phase current (left), SCs under (right) (b) non-adjacent phases: Phase current (left), SCs under (right).

the magnitude indexes are $r_1=2.87$ and $r_2=2.03$ which are higher than one under unbalanced resistances fault in non-adjacent phases (phase A and C) with a phase angle pattern as discussed in chapter 7.
CHAPTER IX
EXPERIMENTAL RESULTS

This chapter presents detailed hardware description that has been developed as part of this dissertation. The detail information on the five-phase motor controller, F-PMa-SynRM, and the five hp dynamo testing bed are provided sequentially. After that, this chapter provides the performances validation through experimental tests of the proposed algorithms as follows 1) maximization of reluctance torque, 2) minimization of torque ripple, 3) magnet temperature estimation, and 4) the fault detection method.

9.1 Five-phase Controller and F-PMa-SynRM

In this section, detail description of the five-phase motor controller and the F-PMa-SynRM is provided respectively.

9.1.1 Five-phase motor controller

The five-phase motor controller can work with an open loop or closed loop control methodologies. During closed-loop control method, this controller can handle one inner loop control (current control) and an outer loop control (speed). For the inner loop control, the controller utilizes the phase current information through the analog
to digital conversion (embedded in F28335). For the outer loop control, the controller utilizes the rotor information provided by the position sensor (incremental encoder).

This motor controller consists of several components. The major components are as follows:

1. Main control unit
   
   (a) Texas instrument digital signal processor (DSP), F28335
(b) Current sensor (connected to the ADCs of F28335)

(c) Position sensor (rotor information provided by incremental encoder)

(d) Buffer IC (control the PWM output to the gate driver)

(e) JTEG Emulator (to carry command from user computer to the DSP)

(f) Signals conditioning circuits (to scale down to 3.3 V for F28335)

(g) Op Amp modules (develop logic for hardware protection)

(h) Gate drivers

2. Power inverter

   (a) Insulated gate bipolar transistor (IGBT)

   (b) DC link capacitor

3. Three DC power supplies (two 6 V and one 15 V)

The five-phase motor control board is built as a laboratory prototype in the Advanced Energy Conversion Lab (AECL), UA. The DC link voltage of the power inverter is chosen to work up to 200 V. The IGBTs are chosen to withstand up to 55 peak Amps. Overall, the inverter is rated for five kilowatts power output. The five-phase control board is shown in the Fig. 9.1.

The detail parts information of the main control unit, power inverter and the gate drives are given in the Table. 9.1, Table. 9.2 and Table. 9.3, respectively.
Table 9.1: Main control unit

<table>
<thead>
<tr>
<th>Item</th>
<th>Model</th>
<th>Part No.</th>
<th>Rating</th>
<th>no of pcs</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistor</td>
<td>2k/0.25 Watt,1%</td>
<td>5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>3k/0.25 Watt,1%</td>
<td>5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>9k/0.25 Watt,1%</td>
<td>1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>0.47k/0.25 Watt,1%</td>
<td>2</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>4.7k/0.25 Watt,1%</td>
<td>1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>5.8k/0.25 Watt,1%</td>
<td>1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>1k/0.25 Watt,1%</td>
<td>10</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Resistor</td>
<td>10k/0.25 Watt,1%</td>
<td>4</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Capacitor</td>
<td>0.1uF/25V</td>
<td>26</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Capacitor</td>
<td>1nF/25V</td>
<td>10</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Capacitor</td>
<td>10uF/25V</td>
<td>2</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Logic gate</td>
<td>74HC4075</td>
<td>771-74HC4075N</td>
<td>2</td>
<td></td>
</tr>
<tr>
<td>High Speed CMOS Logic drv</td>
<td>74HC541</td>
<td></td>
<td>2</td>
<td></td>
</tr>
<tr>
<td>Op Amp</td>
<td>LM2901N</td>
<td>3</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Transistor</td>
<td>BCR112-SOT23 BCR 112 E6327CT-ND</td>
<td>5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Gate drive</td>
<td>SKHI 61R</td>
<td>2</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Current sensor</td>
<td>CAS-25NP</td>
<td>5</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 9.2: Power inverter

<table>
<thead>
<tr>
<th>Item</th>
<th>Rating</th>
<th>no of pcs</th>
</tr>
</thead>
<tbody>
<tr>
<td>Capacitor</td>
<td>10nF/1kV</td>
<td>10</td>
</tr>
<tr>
<td>Capacitor</td>
<td>470uF/200V</td>
<td>3</td>
</tr>
</tbody>
</table>
Table 9.3: Gate driver

<table>
<thead>
<tr>
<th>Item</th>
<th>Rating</th>
<th>no of pcs</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistor</td>
<td>22.1/0.25 Watt,1%</td>
<td>10</td>
</tr>
<tr>
<td>Resistor</td>
<td>8.2k/0.25 Watt,1%</td>
<td>10</td>
</tr>
<tr>
<td>Capacitor</td>
<td>680pF/25V</td>
<td>10</td>
</tr>
<tr>
<td>Capacitor</td>
<td>100nF/25V</td>
<td>2</td>
</tr>
<tr>
<td>Capacitor</td>
<td>10uF/25V</td>
<td>1</td>
</tr>
</tbody>
</table>

9.1.2 Five-phase PMa-SynRM

In this section, the description of the F-PMa-SynRM is given. The F-PMa-SynRM is designed, built and reported earlier in [17]. The specification of the fabricated F-PMa-SynRM is considered as similar to the FEA model and is given in 8.1 (chapter 8). Fig. 9.2 shows the F-PMa-SynRM. Fig. 9.2a shows the five-phase stator. As discussed in chapter 8, it has fifteen slots. Each phase contains twenty two conductors. Fig. 9.2b shows the rotor of the F-PMa-SynRM. It is designed with four poles, eight magnets, and flux barriers. The whole F-PMa-SynRM is shown in Fig. 9.2c. This F-PMa-SynRM is run under field-oriented control and other control methods (healthy and fault conditions) that are discussed in this dissertation.

9.1.3 Overall Hardware Setup

Finally, utilizing the five-phase inverter, a five hp dynamo based experimental set up has been developed to test the F-PMa-SynRM under different loading conditions. An
Figure 9.2: F-PMa-SynRM: (a) stator, (b) rotor, and (c) whole motor.

overview of the experimental set up is shown in the Fig. 9.3. It has major components as follows:
1. DC link voltage source

2. Five-phase inverter

3. DC generator

4. DC electronic load

5. F-PMa-SynRM

6. User interfacing computer
Table 9.4: Sensor Data

<table>
<thead>
<tr>
<th>LOCATION</th>
<th>SENSOR</th>
<th>TEMPERATURE TOLERANCE</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>RANGE</td>
<td>VALUE</td>
</tr>
<tr>
<td>Conductors</td>
<td>Type K</td>
<td>−200°C / + 1250°C</td>
<td>±0.75%</td>
</tr>
<tr>
<td>Thermocouple</td>
<td>Magnets</td>
<td>−73°C / + 260°C</td>
<td>±0.06%</td>
</tr>
</tbody>
</table>

7. Different monitoring devices

All the control methods which are proposed in this dissertation, have been implemented and validated using this set up.

9.2 Hardware for Magnet Temperature

For the verification of the magnet temperature estimation algorithm, additional hardware set up has been developed. Measurement of the stator temperature is simpler than measuring the rotor temperature. For the measurement of the stator temperature $T_s$, type K thermocouples were employed and installed within the slot lining of the conductor coil sides as shown in the Fig. 9.4. The specifications of the employed Type K thermocouples are provided in the Table 9.4. An hardware system which can deliver the magnet temperature data wirelessly is discussed in the following section.
Figure 9.4: Stator temperature measurement setup.

Figure 9.5: Rotating PCB for magnet temperature measurement.
9.2.1 Development of a Circular Disk to Transfer Temperature Data Wirelessly

In order to verify the accuracy of the determined magnet temperature, temperature sensors were mounted on the magnets and connected to an innovative circular Printed Circuit Board mounted on the rotor to determine real time magnet temperature. The obtained data from the sensors were then transmitted wirelessly using Bluetooth. An overview of the design of the PCB employed for this purpose is shown in the Fig. 9.5.

A total of five temperature sensors were employed and mounted on each of the magnets of the motor. The temperature sensors employed for the magnet temperature estimation were PT-100 temperature dependent resistors (RTDs). When compared to thermocouples, these sensors are less susceptible to noise and anticipating the noisy environment within the rotor near the magnets during the operation of the motor, were utilized instead. In order to ensure that they were mounted on the
magnets, the RTD sensors were fixed on to the magnet surface using adhesive leads and mounted on the lip of the magnet structure to ensure mechanical stability as shown in the Fig. 9.6a. It should be noted that in this measurement process, the axial temperature variation along the magnet was considered negligible considering the small stack length of the motor.

An additional design consideration was the radial symmetry of the PCB design in the mounting of the various integrated circuit (IC) components. As the PCB was being mounted on the rotor shaft, the radial symmetry of the device mounting was maintained to avoid introducing any eccentricity due to an apparent uneven load distribution in the rotor during its operation. In order to achieve this, the sensors were divided into two equal halves and the required halves of the sensor’s components and ICs placed at the corresponding diametrically opposite positions on the PCB. However, as only a single micro-controller and Bluetooth device were present in the design, suitable commercial modules were considered during the design to maintain the radial symmetry while they were being mounted. The Fig. 9.6b shows the PCB mounted on the rotor shaft.

9.3 Results for Maximization of Reluctance Torque under Open Phase Faults

In this section, the experimental testing has been performed to verify the effectiveness of the proposed FTC method. The 5hp dynamo system with the five-phase PMa-SynRM in Fig. 9.3 has been utilized.
Figure 9.7: Phase currents and torque: (a) Normal condition, (b) under two-phase non-adjacent open fault, (c) with modification under two-phase non-adjacent open fault, (d) average torque (before and after phase balances).

The proposed FTC algorithm of the five-phase motor is developed in code composer studio V: 5.30, and implemented in the digital signal processor (TI DSP F28335).
9.3.1 Torque and Phase Currents under Balanced Excitation under Faults- Experimental

Fig. 9.7a shows the phase currents (0.5 p.u. peak) under healthy condition of the motor (phases B, C, D, and E). Fig. 9.7b shows the phase currents (phases A, C, and D) under TPNF which are distorted much. Fig. 9.7c shows the phase currents under TPNF when the balanced phase currents are provided in the machine. Under different faults, the change in the torque is observed under balanced excitation in the Fig. 9.7d.

9.3.2 Optimal Phase Advance for Maximum Torque under Different Faults- Experimental

Fig. 9.8a shows the maximum torque variation with the phase advances under SPF. It is observed that the maximum torques are \(\sim 0.45-0.56\) p.u. for a phase advance change in \(\sim 42-60\) degree which is closely similar to the FEA results shown in Fig. 8.4a. Fig. 9.8b and 9.8c show the maximum torque variation with the phase advances under TPAF and TPNF conditions respectively. Under TPAF, Fig. 9.8b shows the maximum torques are \(\sim 0.43-0.51\) p.u. for a phase advance change in \(\sim 45-50\) degree. Under TPNF, Fig. 9.8c shows the maximum torques are \(\sim 0.33-0.49\) p.u. for a phase advance change in \(\sim 43-49\) degree.

9.3.3 Torque Ripple under Different Faults- Experimental

To observe the torque ripple, the five-phase PMa-SynRM has been run at low speed (90-130 rpm). Fig. 9.9a shows the torque ripple under normal and SPF conditions.
Figure 9.8: Optimal phase advance for maximum torque: (a) single-phase open, (b) two adjacent phases open, and (c) two non-adjacent phases open.

It is shown that the torque ripple increases to 12% with a reduction of 25% in the nominal average torque. However, the torque ripple becomes 3.6% with the proposed method while maintaining 92.2% of average torque. Fig. 9.9b shows the ripple under TPAF. It is shown that the torque ripple increases to 38% under TPAF with a 48% reduction in average torque. However, the torque ripple becomes 9.6% with 82.8% of nominal torque. Fig. 9.9c shows the torque ripple increases to 59.2% with a reduction of 58.6% in the nominal torque. With the proposed method, the torque
Figure 9.9: Torque ripple: (a) single-phase open, (b) two adjacent phases open, (c) two non-adjacent phases open, and (d) torque ripple variation with d-axis current. ripple becomes 10.5% with 78.6% nominal torque. Fig. 9.9d shows the torque ripple variation with the d-axis current under healthy and SPF conditions.

9.3.4 Transient Performance of the Fault Tolerant Controller

Fig. 9.10 shows the transient response (torque) of the five-phase system under different phase faults. Under healthy condition, the torque response is shown in Fig. 9.10a. In this condition, the overshoot and settling time is 5.5% and 1.67s, respectively. The torque responses under SPF and TPAF are shown in Fig. 9.10b, and
Figure 9.10: Torque responses: (a) Healthy condition, (b) Single-phase open fault, and (c) two-phase adjacent open fault.

Fig.9.10c, respectively. The overshoots in these conditions are 17% and 10.1%, and the settling times are observed as 1.7s and 0.93s, respectively.

9.3.5 Comparative Results

The Table. 9.5 shows the comparison of the optimal phase advance ($\gamma$) between analytical, FEA and experimental values under different fault conditions. The analytical values of the phase advance ($\gamma$) is calculated by utilizing a numeric solver.
### Table 9.5: Phase Advance Comparison

<table>
<thead>
<tr>
<th>Phase advance (rad)</th>
<th>$\gamma$ (Analytical)</th>
<th>$\gamma$ (FEA)</th>
<th>$\gamma$ (Exp.)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SPF</td>
<td>0.783</td>
<td>$\sim 0.83$</td>
<td>$\sim 0.85$</td>
</tr>
<tr>
<td>TPAF</td>
<td>0.736</td>
<td>$\sim 0.80$</td>
<td>$\sim 0.83$</td>
</tr>
<tr>
<td>TPNF</td>
<td>0.713</td>
<td>$\sim 0.72$</td>
<td>$0.82$</td>
</tr>
</tbody>
</table>

### Table 9.6: Comparison of Torque Performances

<table>
<thead>
<tr>
<th>Operating Conditions</th>
<th>Torque performances</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Average torque</td>
</tr>
<tr>
<td>Healthy</td>
<td>100%</td>
</tr>
<tr>
<td>Single phase open</td>
<td>92.2%</td>
</tr>
<tr>
<td>Adjacent phases open fault</td>
<td>82.8%</td>
</tr>
<tr>
<td>Non-adjacent phases open fault</td>
<td>78.6%</td>
</tr>
</tbody>
</table>

The results in the table are observed consisting between the analytical, FEA and experimental with a maximum error of 13.8% (TPNF) or less.
A comparative analysis of the torque has been given in Table 9.6. The comparison has been made for the average torque (in percentage of the nominal value) and the torque ripple (%) under different open phase faults. It has been observed that the proposed method which has been implemented in a PMa-SynRM, can provide the maximum amount of average torque as 92.2%, 82.8%, and 78.6% of the nominal torque under SPF, TPAF, and TPNF, respectively.

9.4 Results for Minimization of Torque Ripple under Open Phase Faults

In this section, the experimental testing has been conducted to verify the effectiveness of the proposed torque ripple minimization method. The 5 hp dynamo system which is shown in Fig. 9.3, has been operated under different fault conditions. The proposed TRM algorithm of the five-phase motor is developed (code composer studio V: 5.30) and implemented in the digital signal processor (TI DSP F28335).

9.4.1 Back EMF Harmonics under Open Phase Faults

To predict the magnitude of the harmonics in the remaining healthy phase currents under open phase faults, the back EMF harmonics are analyzed experimentally. Fig. 9.11 shows that the harmonics strength in the back EMF under different conditions. Fig. 9.11a shows that the magnitude of the third harmonic under the healthy condition is 15% of the fundamental component whereas it becomes 32% under SPF and 42% under TPAF condition in Fig. 9.11b and Fig. 9.11c, respectively. Additionally, the fifth harmonic is not present under healthy condition whereas it becomes 11%
of the fundamental under SPF and 14% under TPAF. It is clear from Fig. 9.11 that the phase current harmonics do not follow any patterns which endorses the necessity of active harmonic suppression in real time.

9.4.2 Harmonic Injection in the Phase Currents under SPF

Fig. 9.12 shows the phase current under SPF condition. Fig. 9.12a shows the phase currents when no modification is done. The harmonic identifier which discussed in

Figure 9.11: Back Electromotive Force (no modification) : (a) Healthy, (b) single-phase open, and (c) adjacent phase open.
Fig. 9.13 shows the normalized amplitudes of these harmonics. It is easily perceivable from Fig. 9.13 that the third harmonic is stronger in phase C, D and E (more than 50% of the fundamental signals).

To reduce the torque ripple a balanced excitation with a third harmonic injection has been applied to the machine which is shown in the Fig. 9.12b. The amplitudes of the balanced third harmonics are shown in the Fig. 9.14. The injected third harmonic in the stator phase currents are maintained nearly 48% of the fundamental with a balanced phase angles (5.19).

![Figure 9.12](image)

Figure 9.12: Phase current under single phase fault: (a) with no modification, and (b) with proposed method.
9.4.3 Harmonic Injection in the Phase Currents under TPAF

Fig. 9.15 shows the phase current under TPAF condition. Fig. 9.15a shows the phase currents when no modification is done to the system. The harmonic identifier which is discussed in chapter 5, is utilized to identify the amplitude of the harmonics under this condition. Fig. 9.16 shows the normalized amplitudes of these harmonics. It is
observed that the third harmonics in the stator phase currents are more than 100% of the fundamental in the phase C and D.

To reduce the effect of these strong harmonics, a balanced excitation with a fundamental and third harmonic has been applied to the machine as discussed in (5.22). The balanced excitation is shown in the Fig. 9.15b. The amplitudes of the
Figure 9.15: Phase current under two-adjacent phase fault: (a) with no modification, and (b) with proposed method.

Table 9.7: Comparison with FEA and experimental test.

<table>
<thead>
<tr>
<th>Fault type</th>
<th>Torque ripple</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>FEA</td>
</tr>
<tr>
<td>Single phase open</td>
<td>11.3%</td>
</tr>
<tr>
<td>Two-phase adjacent open</td>
<td>10.6%</td>
</tr>
<tr>
<td></td>
<td>Experimental test</td>
</tr>
<tr>
<td></td>
<td>11%</td>
</tr>
<tr>
<td></td>
<td>12%</td>
</tr>
</tbody>
</table>

balanced third harmonics are shown in the Fig. 9.17.

9.4.4 Torque and Speed with the Proposed Method under Open Phase Faults

To test the machine under fault, it has been operated at 270 rpm which is shown in Fig. 9.18. In Fig. 9.18, the speed shows the constant value under normal condition and with the proposed method. During the fault, the speed reveals fluctuating nature.
Figure 9.16: Harmonic under two adjacent fault (no modification): (a) phase C, (b) phase D, and (c) phase E.

Fig. 9.19 shows the experimental torque results under SPF and TPAF conditions with the transition of the torque at the time of fault occurrence and the time of phase adjustment with harmonic injection. Fig. 9.19a shows the torque ripple increases during the SPF and it reduces to 11% when the proposed method is applied. Fig. 9.19b shows the torque ripple is increased under TPAF. However, with the proposed
Figure 9.17: Harmonic under TPAF with proposed method: (a) phase C, (b) phase D, and (c) phase E.

method the torque ripple is reduced to 12%.

9.4.5 Comparison Analysis

The summary of the torque ripple and mean normalized torque under various tests is shown in Fig. 9.20. It is observed in Fig. 9.20, under SPF the torque ripple has been improved from 22% to 11% while maintaining 98% mean with the proposed
method. Under TPAF, the torque ripple has been improved from 23% to 12% while maintaining 88% mean torque with the proposed method. A comparison between the FEA and test results is given in Table 9.7. This Table 9.7 shows the results between the predicted in FEA and experimental tests are matched with a minimal difference. The efficiency analysis (50% load) during the faults and the proposed
method is shown in the Fig. 9.21. It is observed, the efficiency dropped to 58% under TPAF which is improved to 75% while applying the proposed method. Similarly, under SPF, the efficiency is improved to 91% from 84%.
9.5 Results of Magnet Temperature Estimation

In this section, the experimental validation of the proposed temperature estimation is provided. In order to verify the proposed algorithm with the frequency determination, the motor was run at multiple operating frequencies and the frequency selection algorithm was employed to dynamically determine the frequency at which the signal needed to be injected for the estimation of the magnet temperature.

The operating frequencies are selected as both low (6 Hz and 9 Hz) and high frequencies (35 Hz and 58.5 Hz). At each condition, a high frequency is selected by utilizing the proposed algorithm in Fig. 6.1. The algorithm generated the cleaner high frequencies which would be utilized to inject and estimate the magnet temperature as given in the Table 9.8.

<table>
<thead>
<tr>
<th>OPERATING FREQUENCY (Hz)</th>
<th>INJECTED FREQUENCY (Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6 Hz</td>
<td>310 Hz</td>
</tr>
<tr>
<td>9 Hz</td>
<td>426 Hz</td>
</tr>
<tr>
<td>35 Hz</td>
<td>1501 Hz</td>
</tr>
<tr>
<td>58.5 Hz</td>
<td>2341 Hz</td>
</tr>
</tbody>
</table>
9.5.1 Direct Measurement of the Magnet Temperature

First of all, the data obtained from the thermocouples (stator conductors) were transduced using a data acquisition device. The output of the thermocouple for a fixed current, $I_{RMS}$ of 5A and with different speed is as shown in the Fig. 9.22.

The five RTD sensors data of the magnet temperature is taken wirelessly as given in the Fig. 9.23 under 58.5 Hz operation.

9.5.2 Estimation Through Injection of Selected Frequencies in Phase Current Signals

In this section, the proposed analytical estimation method through cleaned frequency injection has been implemented experimentally.
Phase Current Signals

The current waveforms considering the operating frequency at 6 Hz injected with 310 Hz and 58.5 Hz injected with 2341 Hz is shown in Fig. 9.24. It is clearly observed that these phase currents become distorted under the frequency injection. However, the total harmonic distortion does not change much under the process which is explained in the following section through the phase current spectrum analysis.

Phase Current Spectrum Analysis

In order to verify the efficacy of the frequency selection algorithm, the corresponding normalized frequency spectrum output of the measured current signals (phase A)
Figure 9.24: Current signals: (a) 6 Hz operating frequency with 310 Hz injected signal and (b) 58.5 Hz operating frequency with 2341 Hz injected signal.

It is observed from Fig. 9.25 that the 3rd and 7th harmonics maintain 28% and 9% of the fundamental signal. However, the magnitude of these low order harmonics do not vary before and during the selected signal injection which proves that the total
harmonic distortion (THD) remain unchanged for them during the clean frequency injection. However, as the magnitude of the selected frequency shows up during the injection method it increases the THD very little (less than 1%). It is also clearly observed that the selected frequency at 310 Hz shows prominent nature comparing the nearby noise signals and higher order harmonics proving the performances of the frequency determination algorithm and meet the requirements with regards to spectral separation. Additionally, the resulting measured frequency component is free from the contributions of the fundamental frequency and its harmonics, as well as fault frequencies and its components at the particular operating condition and only corresponds to the corresponding commanded signal.
Fig. 9.26 shows the phase current (phase A) spectrum at the operating frequency at 35 Hz injected with 1501 Hz. At this speed condition the low order harmonics show less dominance comparing the 6 Hz speed operation (in Fig. 9.25). At this condition the 3rd harmonic becomes 10% of the fundamental and the 7th harmonic becomes 2% of the fundamental. The magnitudes of these low order harmonics do not change with the injection of the selected frequency which ensures the preservation of the THD under frequency injection for them. However, the THD increases for the selected high frequency very little (less than 1%). Fig. 9.26 also shows the injected frequency at 1501 Hz holds the stronger magnitude comparing the nearby noises and harmonics proving accurate means for the temperature estimation.
For further verification of the proposed frequency determination algorithm, the machine was run near rated speed at 58.5 Hz with 2341 Hz injection. Fig. 9.27 shows the phase current spectrum at this condition where the 3\textsuperscript{rd} and the 7\textsuperscript{th} harmonic become 12\% and 3\% of the fundamental signals, respectively, providing almost same THD before and during the frequency injection in the system. Additionally, the injected signal at 2341 Hz meets the initial expected spectral separation ensuring the accurate means for the temperature estimation.

9.5.3 Torque under Proposed Frequency Injection Method

It is also imperative to check the output torque whether the average and ripple in the torque still holds under satisfaction level. A plot of the torque output for the
operating condition at 6 Hz under normal operating condition and after injection of the 310 Hz signal is shown in the Fig. 9.28 which clearly indicates a relatively low impact of the injected signal on the output torque of the motor.

9.5.4 Estimated Magnet Temperature using the Frequency Determination Algorithm

Utilizing (??), the magnet temperature has been calculated from the injected high frequencies in Table 9.8. For 6 Hz, 9 Hz, 35 Hz and 58.5 Hz the magnet temperature is estimated as 29.7, 29.7, 29.5, 29.4 °C receptively.
9.5.5 Comparison between Direct Measurement and Estimation of the Magnet Temperature using the Frequency Determination Algorithm

From the data determined it is clear that the proposed algorithm provides an easy and efficient method to determine the frequency to be utilized for magnet temperature estimation using the signal injection method. A comparison of the magnet temperatures between the direct measurement (wirelessly), FEA results, and estimation through the proposed algorithm is given in Table 9.9. The results obtained clearly show that the proposed method of implementation of the signal injection method in the case of a PMa-SynRM is valid and provides relatively agreeable level of accuracy.

Table 9.9: Comparison Results

<table>
<thead>
<tr>
<th>OPERATING FREQUENCY (Hz)</th>
<th>INJECTED FREQUENCY (Hz)</th>
<th>MEASURED MAGNET TEMPERATURE (IN °C)</th>
<th>ESTIMATED MAGNET TEMPERATURE (IN °C)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6 Hz</td>
<td>310 Hz</td>
<td>28.75</td>
<td>29.7</td>
</tr>
<tr>
<td>9 Hz</td>
<td>426 Hz</td>
<td>28.857</td>
<td>29.7</td>
</tr>
<tr>
<td>35 Hz</td>
<td>1512 Hz</td>
<td>29.1</td>
<td>29.5</td>
</tr>
<tr>
<td>58.5 Hz</td>
<td>2880 Hz</td>
<td>29.25</td>
<td>29.4</td>
</tr>
</tbody>
</table>
9.6 Results for Fault Detection in F-PMa-SynRM

In this section, the experimental testing has been performed to verify the effectiveness of the proposed fault diagnosis method. The same 5 hp dynamo system (Fig. 9.3) has been utilized. Field orient control of five-phase motor has been developed for the control of the machine. To guarantee the safety during the testing a faulty machine, and to validate the proposed method the machine has been run at low speed at different load conditions.

Figure 9.29: Currents under phase CE=0, (a) phase A, (b) phase B, (c) phase D.
9.6.1 Phase Currents under TPNF Condition

Fig. 9.29 shows the motor currents under TPNF (phase C and E) when the machine is loaded with 30% load. In Fig. 9.29, the original currents in other healthy phases (phase A, B and D) are shown with the fundamental signals. THD in phase A, B, and D are calculated as 44%, 46.2%, and 17.7%, respectively. The p.u. values of the fundamental currents are 0.39, 0.24, and 0.42 p.u. respectively. Table 9.10 shows the p.u. values of the fundamental currents under different open phase fault conditions.

Under healthy condition, the efficiency of the five-phase motor has been reported as 91% at 30% load in [126]. However, with the same load, under different fault conditions the efficiencies are shown in Fig. 9.30. In Fig. 9.30, it is observed that the efficiencies are \(~83\%\), \(~60\%\), and \(~55\%\) under SPF, TPAF, and TPNF re-
spectively. The torque and torque ripple under different fault conditions have been shown in Fig. 9.31. It is observed that the average torques are 2.15 Nm, 1.6 Nm, 1.4 Nm, and 1.3 Nm, and the torque ripple are 5.03%, 21.43%, 23.81%, and 27.50% under healthy, SPF, TPAF, and TPNF, respectively.

9.6.2 Analysis of the Symmetrical Components - Amplitude

In this section, the SCs under different open phase faults have been calculated. The fundamental currents measured in section IV(A) are converted to the SCs by utilizing the proposed transformation Matrix in (7.3). The calculated SCs are depicted in Fig. 9.32, when the machine is loaded with 30% load. Fig. 9.32a shows the SCs under healthy condition. It is observed in Fig.9.32a that the magnitude of PS1 is a non-zero value, whereas the rest of the SCs (ZS, NS1, PS2 and NS2) are close to zero which are well matched with FEA results in Fig. 8.14a and theoretical analysis in section II(D).
Table 9.10: Per Unit Values of Fundamental Current under Faults

<table>
<thead>
<tr>
<th>Operating Conditions</th>
<th>p.u. values (Phases)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Healthy Condition</td>
<td>0.35 0.35 0.35 0.35 0.35</td>
</tr>
<tr>
<td>Phase E Open</td>
<td>0.33 0.27 0.39 0.39 0</td>
</tr>
<tr>
<td>Phases D and E Open</td>
<td>0.46 0.26 0.50 0 0</td>
</tr>
<tr>
<td>Phases C and E open</td>
<td>0.39 0.24 0 0.42 0</td>
</tr>
</tbody>
</table>

Fig. 9.32b shows the SCs under the SPF condition. It has been observed that the magnitudes of the ZS, NS1, PS2, and NS2 are .062, 0.05, 0.06, and 0.05 p.u. and the magnitude indexes, \( r_1 \) and \( r_2 \) are close to 1 as it is shown in Fig. 8.14b.

Fig. 9.32c shows the SCs under the TPAF (D and E) condition. In this fault, the magnitudes of the ZS, NS1, PS2, and NS2 are 0.02, 0.09, 0.014, and 0.11 p.u. and the magnitude indexes, \( r_1 \) and \( r_2 \) are less than one (< 1) as it is shown in Fig. 8.14c.

Fig. 9.32d shows the SCs for the TPNF (C and E). The magnitudes of the ZS, NS1, PS2, and NS2 are 0.026, 0.10, 0.12, and 0.034 p.u. and the magnitudes indexes, \( r_1 \) and \( r_2 \) are higher than one (> 1) which are also shown in the Fig. 8.14d.
Figure 9.32: SCs at 30% load: (a) Healthy condition, (b) phase E=0, (c) phase DE=0, (d) phase CE=0.

9.6.3 Analysis of the Symmetrical Components - Phase

In this section, the phases of the proposed SCs have been analyzed to identify the unique patterns under different open phase faults. Fig. 9.32b (phase E=0) shows that, in reference to the PS1, the phases of the ZS, NS1, PS2, and NS2 are leading-lagging-lagging-leading which are well matched with the FEA results as in Fig. 8.17.

Fig. 9.32c shows the phases of the ZS, NS1, PS2, and NS2 are lagging-leading-lagging-leading which provide the information of the fault-location under TPAF which
Table 9.11: Magnitude Indexes under Different Faults

<table>
<thead>
<tr>
<th>Fault types</th>
<th>FEA (% load)</th>
<th>Experimental (% load)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>100% 30%</td>
<td>100% 30%</td>
</tr>
<tr>
<td></td>
<td>30% 13%</td>
<td>30% 13%</td>
</tr>
<tr>
<td></td>
<td>13%</td>
<td>13%</td>
</tr>
<tr>
<td></td>
<td>$r_1$</td>
<td>$r_2$</td>
</tr>
<tr>
<td></td>
<td>$r_1$</td>
<td>$r_2$</td>
</tr>
<tr>
<td>Single-phase open</td>
<td>0.97 0.98</td>
<td>0.95 0.96</td>
</tr>
<tr>
<td>Adjacent phase open</td>
<td>0.68 0.33</td>
<td>0.62 0.28</td>
</tr>
<tr>
<td>Non-adjacent phase open</td>
<td>2.11 2.52</td>
<td>2.05 1.89</td>
</tr>
</tbody>
</table>

are also well matched to the FEA results as in Fig. 8.17.

Fig. 9.32d shows the phases of the ZS, NS1, PS2, and NS2 are leading-lagging-leading-lagging which provide the information of the fault-location under TPNF condition which are also well matched to the FEA results as in Fig. 8.17.

Similar test has been conducted when the machine has been loaded with 13% rated load. Fig. 9.33a, Fig. 9.33b, and Fig. 9.33c show the SCs under SPF, TPAF, and TPNF, respectively.
Table 9.12: Phase Information of SCs under Faults (w.r.t. positive sequence 1)

<table>
<thead>
<tr>
<th>Fault types</th>
<th>FEA (100% and 30% load)</th>
<th>Experimental (13% and 30% load)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Phase</td>
<td>Phase</td>
</tr>
<tr>
<td>Single-phase open</td>
<td>ZS NS1 PS2 NS2</td>
<td>ZS NS1 PS2 NS2</td>
</tr>
<tr>
<td></td>
<td>+ - - +</td>
<td>+ - - +</td>
</tr>
<tr>
<td>Adjacent phase open</td>
<td>- + - +</td>
<td>- + - +</td>
</tr>
<tr>
<td>Non-adjacent phase open</td>
<td>+ - + +</td>
<td>+ - + +</td>
</tr>
</tbody>
</table>

9.6.4 Summary of the Proposed Amplitude Indexes and Phase of the Five-phase SCs

In this section, the summary of the overall analysis (finite element analysis and experimental results) has been tabulated in Table 9.11 and Table 9.12. In Table 9.11, the magnitude indexes, $r_1$ and $r_2$ are presented for similar type of faults which has
Figure 9.33: SCs at 13% load: (a) phase A=0, (b) phase AE=0, (c) phase AD=0.

been found using FEA and experimental results (two load points). The results show that, under SPF, they remain close to one. Under TPAF, $r_1$ and $r_2$ become less than
one (< 1) and under TPNF, they become higher than one (> 1). In the Table 9.12, the phase angles (ZS, NS1, PS2, and NS2) have been tabulated for similar type of faults which has been found in FEA and experimental results.
CHAPTER X

CONCLUSIONS

This chapter summarizes the dissertation, discusses its findings and contributions and provides directions for future research.

10.1 Summary of Present Works

The dissertation presents analysis and control of the five-phase permanent magnet assisted synchronous reluctance motor (F-PMa-SynRM) drive under different open phase faults. F-PMa-SynRMs are excellent choice while requiring less torque ripple and higher fault tolerance. They allow reduced current per phase without increasing the voltage, higher torque density, wider speed range, and better stability. Due to an additional number of phases, the control features of those machines are different from the conventional three-phase drives.

In chapter 4, various open phase fault conditions have been analyzed mathematically to provide maximum and sustainable torque control. As in F-PMa-SynRMs, the saturation effects are critical, the amplitudes of phase currents have been reasonably limited in the algorithm derivation. With these limited magnitudes, the phase angles have been modified through MATLAB and FEA simulations to provide balanced excitations under different fault conditions. Also, to provide the maximum
amount of reluctance torque in the F-PMa-SynRM, an optimal phase advance method has been proposed which provides maximum torque control with fewer saturation effects under fault conditions.

In chapter 5, a torque ripple minimization method has been developed to perform under different open phase faults in an F-PMa-SynRM. The detailed analytical theory has been established to improve the torque ripple by suppressing the dominant phase current harmonics. The proposed method has been formulated with three steps including the current harmonic identification, harmonic injection, and phase balancing of both fundamental and third harmonic under faults. A correlation-based signal identifier has been introduced in the feedback control loop to provide more straightforward and effective implementation of the harmonic identification.

In chapter 6, a novel approach to determine a suitable frequency to estimate the magnet temperature in an F-PMa-SynRM is developed. The proposed frequency determination algorithm employed constraints based on the isolation of the fundamental, fault frequencies, the corresponding harmonics and the signal to noise ratio. A detail analytical method has been carried out to accurately estimate the magnet temperature by utilizing the selected high frequencies.

In chapter 7, the symmetrical components theory has been developed for the fault detection of a five-phase system. This theory has been demonstrated to efficiently explain the open phase faulty conditions in a five-phase PMa-SynRM. The magnitude and phase pattern of the proposed SCs has been optimally utilized to find the types of faults. Under the analysis, three major faults in a five-phase system have
been extensively followed. Additionally, the robustness of the proposed method has
also been evaluated under unbalanced resistances fault in a five-phase system.

To check the performance of these proposed methods, an FEA model of
the F-PMa-SynRM has been developed and utilized to generate simulation results.
For experimental validation, a real 3 kW F-PMa-SynRM is built as a laboratory
prototype. A five-phase inverter system is also developed and coupled with a 3.7 kW
dynamo system. The inverter is controlled digitally by using the TI DSP F28335. To
validate the magnet temperature estimation algorithm, an innovative wireless data
transmission PCB was designed and mounted on the rotor shaft.

10.2 Key Findings and Contributions

The key findings and contributions of those proposed works are given as below

1. Maximization of the reluctance torque.

   (a) The analytical phase advances under different faults are found very close
to the FEA and experimental values.

   (b) The maximum deviation is observed for a two-phase non-adjacent open
fault condition (TPNF) between the FEA and experimental values that
showed 13.8% error.

   (c) Using proposed FTC method, 92.2% torque was obtained during the SPF
condition while comparing existing methods which reported maximum 80% of
the nominal torque.
(d) Using proposed FTC method, 82.8% torque was obtained during the TPAF condition while comparing existing methods which reported maximum 24.1% of the nominal torque.

(e) Using proposed FTC method, 78.6% torque was obtained during the TPNF condition while comparing existing methods which reported maximum 57.3% of the nominal torque.

2. Minimization of the torque ripple.

(a) The results obtained from the FEA and experiment match closely with a deviation of 0.3% for SPF and 1.4% for TPAF condition.

(b) The torque ripple improves from 22% during SPF to 11% while applying the proposed method.

(c) The torque ripple improves from 23% during TPAF to 12% while applying the proposed method.


(a) Four different speeds of operation were performed to test the algorithm. The operating speeds were 6 Hz, 9 Hz, 35 Hz, and 58 Hz.

(b) In all operating conditions, the results indicate that the proposed method shows a close agreement between the direct measurement and the estimated values.

(c) The biggest difference was observed during 6 Hz operation and the deviation between the estimated and directly measured temperature was 3.3%. 

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4. Open-phase fault detection.

(a) The proposed fault detection algorithm is validated through two different experiment tests that considered 13% and 30% of the load.

(b) The detection method is easier to implement than the complex motor current signature analysis. The detection results are found 100% consistent between the FEA simulation and experimental tests.

(c) The detection method is also validated for the partial unbalanced condition of the phases.

(d) Moreover, the proposed method does not require any additional hardware.

From the results obtained, it is evident that the proposed FTC method can provide maximum torque control and TRM method can reduce the torque ripple significantly which suggest that they can be promising control techniques for F-PMa-SynRM in industries. It is also demonstrated that the proposed magnet temperature estimation and the fault detection methods can perform accurately and also can be strong candidates in motor drive applications.

10.3 Future Works

The future works of this dissertation can extended as follows:

1. This dissertation focused only on the maximum torque per ampere control during different open phase fault condition. Therefore, there is the scope of contributing to the flux weakening fault tolerant control of the F-PMa-SynRM.
2. During the FTC, there was less importance given to the motor efficiency. Further improvement in the efficiency can be done by developing hybridized FTC control which shall focus both on maximum torque realization and efficiency.

3. Instead of considering complete phase failure, the FTC under partial phase unbalanced can be developed.

4. Also, the proposed FTC did not consider model parameter changes. In practical, these model parameters may change for uneven temperature rise and localized saturation. Model predictive control under faults can be developed and hybridized with the proposed FTC to reduce model parameters errors.

5. In the proposed TRM method, the thermal effects due to the harmonic injection is not observed. Therefore, it can be a strong point to further analyze the TRM method.

6. Under fault conditions, if the machine is run for a long-term, there might be possibilities to demagnetize the magnet. Further improvement of the TRM method can be done by observing such demagnetization problems.

7. The noise, vibration, and harshness test can be done for further validation of the proposed TRM method.

8. The estimation method is validated for only small rises in temperature in the magnet, such as \(~ 30\) deg C. Higher temperature testing can be done to check the robustness of the proposed method.
9. The fault detection method was not fully automated. The major challenge was to design the filter under a fault condition, and it was not possible to calculate the filter gains online. For convenient operation, the system can be improved to work independently.

10. The fault detection method has not been validated for higher speed operation. At high-speed operation, there might arise new issues which require the existing method to be improved.

11. As the F-PMa-SynRM is a highly non-linear machine, advanced control methods such as a state feedback control method can be developed to increase the stability of the system.


