Design and Analysis of Fractional-Slot Concentrated-Winding Multiphase Fault-Tolerant Permanent Magnet Synchronous Machines

DISSERATION
submitted in partial fulfillment of the requirements for the degree of Doctor of Philosophy in Applied Engineering by

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under the supervision of
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To my family and friends
and most specially to my grandmother Menchu.

“And yet it moves...”
quote attributed to Galileo Galilei
Acknowledgments

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me unconditionally. This thesis is specially dedicated to you.

In these lines I am probably forgetting many people who would deserve some words of thanks. To those that should be in this list and are missing, I sincerely apologize, but it’s impossible to remember everyone. Therefore, to all of you, named or not...

Thank you very much!
Summary

In the last decades, the use of permanent magnet machine drives has experienced a sustained growth owing to their high efficiency and power density figures and due to their inherent suitability for direct-driven applications. However, and despite being highly reliable, the fact that the excitation field in a permanent magnet machine cannot be turned off at will has made engineers reluctant to employ these drives in safety critical applications in the past.

Various techniques have been proposed in the related literature to grant fault-tolerance to a permanent magnet machine drive. This thesis starts by reviewing previous work on the matter and by analyzing the different fault-tolerant approaches. After the various methods are briefly discussed, a comparison between the distinct techniques is established, from which the approach of splitting the drive in multiple independent phases emerges as one the most promising design procedures. This requires that the drive is designed to provide the maximum possible magnetic, electrical, thermal and physical isolation between phases. In order to limit the high magnitude currents arising from a short-circuit fault, a further requirement is that the permanent magnet machine is designed to have a high enough phase self-inductance. The previous requisites are naturally met in permanent magnet machines making use of fractional-slot concentrated-windings. Additionally, multiphase systems have shown to provide a number of advantages over the traditional three-phase systems; specially regarding fault-tolerance and the attainable level of performance after a fault.

A next step in the development of the present thesis has been to review the design principles that allow to select the most appropriate winding arrangements for fault-tolerant multiphase fractional-slot concentrated-winding permanent magnet machines. From the research conducted, it has been found out that the traditionally proposed rules to select the most adequate configurations are restricted to odd phase number machines or to specific winding configurations. In order to fill this gap, an analytical procedure has been established to evaluate the merits of different winding configurations in terms of magnetic isolation and regardless of the geometry of the machine.

A design methodology incorporating the previous winding selection criteria has been proposed for fault-tolerant permanent magnet machines. Based on this methodology, a five-phase machine prototype has been designed and manufactured. The design process for the prototype, including the analysis of the required specifications and design constraints, is thoroughly discussed. A number of experimental tests have demonstrated the intrinsic fault-tolerant capability of the prototype machine and the suitability of the proposed design methodology.

Next, an analytical drive model suitable for fault analysis has been developed. The model has served as a tool to predict the behavior of the designed machine under different fault conditions and to test post-fault remedial strategies. Specifically, the post-fault operation under winding open-circuit faults, terminal short-circuit faults and transistor open and short-circuit faults has been investigated.

For the previous fault scenarios, modified control strategies that allow to improve the post-fault performance of AC machine drives have been proposed. In particular, a unified approach to compute suitable current references for winding open-circuit and terminal short-circuit faults has been derived. The method, aimed at minimizing the stator copper losses while preserving the main harmonic of the air-gap magnetomotive force, is general and valid for any phase number drive and different supply conditions.

Experimental tests on the designed prototype have demonstrated the adequacy of the proposed modified control strategies in reducing the parasitic effects arising from the different fault conditions. Furthermore,
by adopting the proposed remedial actions, it has been possible to operate the machine drive under fault scenarios for which the system previously became unstable.
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List of Publications

The following publications have been originated from the present thesis:


List of Symbols

The symbols are divided among Roman symbols and Greek symbols. The more common subscripts adopted in the present document are reported as well.

Common subscripts

- $A, B, C...$: Machine phases
- $agap$: Air-gap component
- $b$: Base quantity
- $D, L, R, U$: Down, left, right and up regions
- $DC$: DC value (constant)
- $d-q$: Axes of the synchronous reference frame
- $ef$: Effective
- $m$: Permanent magnet component
- $max$: Maximum value
- $mean$: Mean or average value
- $min$: Minimum value
- $n$: Subscript index
- $nom$: Nominal or rated value
- $ph$: Phase variables
- $RMS$: RMS value
- $r, s$: Rotor, stator regions
- $sc$: Short-circuit term
- $slot$: Slot component
- $slot$-$opening$: Slot-opening component
- $t, y$: Teeth, yoke regions
- $x, y, z$: Cartesian components/coordinates
- $\alpha - \beta$: Axes of the fixed reference frame (Clarke transformation)
- $\theta, r, z$: Cylindrical components/coordinates
# Roman Symbols

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<td>$A$</td>
<td>Electric loading or linear current density</td>
</tr>
<tr>
<td>$A$</td>
<td>Magnetic vector potential</td>
</tr>
<tr>
<td>$B$</td>
<td>Magnetic flux density</td>
</tr>
<tr>
<td>$B_{agap}$</td>
<td>Radial air-gap flux density, peak value</td>
</tr>
<tr>
<td>$B_{r,m}$</td>
<td>Magnet remanence</td>
</tr>
<tr>
<td>$B_1$</td>
<td>Stator tooth flux density, peak value</td>
</tr>
<tr>
<td>$B_{y,r}$</td>
<td>Rotor yoke flux density, peak value</td>
</tr>
<tr>
<td>$B_{y,s}$</td>
<td>Stator yoke flux density, peak value</td>
</tr>
<tr>
<td>$b$</td>
<td>Slot width</td>
</tr>
<tr>
<td>$b_0$</td>
<td>Slot-opening width</td>
</tr>
<tr>
<td>$b_1$</td>
<td>Slot bottom width</td>
</tr>
<tr>
<td>$b_2$</td>
<td>Slot top width</td>
</tr>
<tr>
<td>$D_{agap}$</td>
<td>Air-gap mean diameter</td>
</tr>
<tr>
<td>$D_{cond,Cu}$</td>
<td>Conductor copper diameter</td>
</tr>
<tr>
<td>$D_{cond,w}$</td>
<td>Conductor wire diameter</td>
</tr>
<tr>
<td>$D_{ext,PM}$</td>
<td>Permanent magnet outer diameter</td>
</tr>
<tr>
<td>$D_{ext,r}$</td>
<td>Rotor iron outer diameter</td>
</tr>
<tr>
<td>$D_{ext,s}$</td>
<td>Stator outer diameter</td>
</tr>
<tr>
<td>$D_{int,s}$</td>
<td>Stator bore diameter</td>
</tr>
<tr>
<td>$D_{shaft}$</td>
<td>Rotor shaft diameter</td>
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<tr>
<td>$e$</td>
<td>Instantaneous back electromotive force</td>
</tr>
<tr>
<td>$f$</td>
<td>Fundamental electrical frequency</td>
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<tr>
<td>$g$</td>
<td>Air-gap thickness</td>
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<tr>
<td>$H$</td>
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<td>$H_c$</td>
<td>Magnet coercitivity</td>
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<td>$h$</td>
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<td>Wedge height</td>
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<td>$h_m$</td>
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<td>$h_{r,tip}$</td>
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<td>$h_{y,s}$</td>
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<td>$k_{Fe}$</td>
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<td>$k_f$</td>
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<td>$k_w$</td>
<td>Winding factor</td>
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<td>$L_0$</td>
<td>$0$-axis inductance</td>
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<td>$L_b$</td>
<td>Base inductance</td>
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<tr>
<td>$L_{q_3}$</td>
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<td>$l_{tot}$</td>
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m  Number of phases  
N  Mechanical speed (rpm)  
N_{cpg}  Number of coils per group and phase  
N_{grp}  Number of coil groups  
N_{ph}  Number of coils per phase  
N_s  Number of series turns per phase  
N_{scp}  Number of series coils per phase  
n_t  Number of winding layers  
P  Electromagnetic power  
P_{bear}  Bearing loss  
P_{Cu}  Copper loss  
P_{in}  Input power  
P_{Fe,t}  Stator teeth loss  
P_{Fe,y,r}  Rotor iron loss  
P_{Fe,y,s}  Stator yoke loss  
P_{PM}  Magnet loss  
P_{shaft}  Output shaft power  
P_{wind}  Windage loss  
p  Number of pole pairs  
Q  Number of stator slots  
q  Number of slots per pole and phase  
R_{ph}  Stator phase resistance  
R_r  Rotor outer radius  
R_s  Stator inner radius  
T  Temperature  
T_e  Electromagnetic torque  
T_a  Ambient temperature  
T_{bear}  Bearing temperature  
T_{end-wdg}  End-winding temperature  
T_{hous}  Housing temperature  
T_{PM}  Magnet temperature  
T_{shaft}  Rotor shaft temperature  
T_t  Stator teeth temperature  
T_w  Winding temperature  
T_{y,r}  Rotor yoke or back-iron temperature  
T_{y,s}  Stator yoke temperature  
t  Electrical periodicity of the machine, \(gcd(p, Q)\)  
t  Time instant  
U  RMS phase voltage  
U_{DC}  DC bus voltage  
W  Magnetic energy  
w  Magnetic energy density  
w_t  Stator tooth width  
y_q  Coil throw or coil span (expressed in number of slots)  
Z  Number of turns per coil  
Z_{paral}  Number of parallel conductors  
Z_{sc}  Short-circuit impedance
Greek symbols

\( \alpha_0 \) Stator slot opening angle
\( \alpha_{B_r} \) Magnet remanence temperature coefficient
\( \alpha_{H_c} \) Magnet coercitivity temperature coefficient
\( \alpha_m \) Electrical angle between phase windings
\( \alpha_{PM} \) Magnet arc width referred to the pole-pitch
\( \alpha_{pole} \) Rotor pole angle
\( \alpha_s \) Electrical angle between the phasors of two adjacent slots
\( \alpha_{slot} \) Stator slot angle
\( \beta \) Slot leakage related dimensionless parameter
\( \gamma \) Slot leakage related dimensionless parameter
\( \delta \) Electrical angle between the phase voltage and the induced voltage
\( \delta \) Slot leakage related dimensionless parameter
\( \epsilon \) Slot leakage related dimensionless parameter
\( \zeta_n \) Air-gap inductance related dimensionless parameter
\( \eta \) Efficiency
\( \theta_e \) Electrical angle for phase variables
\( \Lambda_m \) RMS value of the flux linkage due to the permanent magnets
\( \lambda \) Instantaneous flux linkage
\( \lambda \) Permeance coefficient
\( \lambda_m \) Flux linkage due to the permanent magnets
\( \mu_0 \) Permeability of free space
\( \mu_{r,m} \) Magnet relative permeability
\( \nu \) Harmonic order
\( \xi \) Relationship between air-gap inductance coefficients (figure of merit)
\( \zeta \) Relationship between slot inductance coefficients (figure of merit)
\( \tau_{PM} \) Magnet-pitch
\( \tau_p \) Pole-pitch
\( \phi \) Electrical angle between the phase voltage and the current
\( \phi_{sc} \) Electrical angle between the current and the voltage in a short-circuited phase
\( \chi \) Torque derating factor
\( \psi \) Electrical angle between the phase current and the induced voltage
\( \psi \) Relationship between inductance coefficients (figure of merit)
\( \omega_e \) Fundamental electrical angular frequency
## List of Acronyms

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<tr>
<td>AC</td>
<td>Alternating current</td>
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<tr>
<td>BLDC</td>
<td>Brushless DC</td>
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<td>CAD</td>
<td>Computer-aided design</td>
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<tr>
<td>CFD</td>
<td>Computational fluid dynamics</td>
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<td>DC</td>
<td>Direct current</td>
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<tr>
<td>DSP</td>
<td>Digital signal processor</td>
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<tr>
<td>DTC</td>
<td>Direct torque control</td>
</tr>
<tr>
<td>EMF</td>
<td>Electromotive force</td>
</tr>
<tr>
<td>EV</td>
<td>Electric vehicle</td>
</tr>
<tr>
<td>FEA</td>
<td>Finite element analysis</td>
</tr>
<tr>
<td>FEM</td>
<td>Finite element method</td>
</tr>
<tr>
<td>FSCW</td>
<td>Fractional-slot concentrated-winding</td>
</tr>
<tr>
<td>IGBT</td>
<td>Insulated gate bipolar transistor</td>
</tr>
<tr>
<td>IPM</td>
<td>Interior permanent magnet</td>
</tr>
<tr>
<td>ISA</td>
<td>Integrated starter alternator</td>
</tr>
<tr>
<td>MMF</td>
<td>Magnetomotive force</td>
</tr>
<tr>
<td>MTPA</td>
<td>Maximum torque per ampere</td>
</tr>
<tr>
<td>OC</td>
<td>Open-circuit</td>
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<tr>
<td>PM</td>
<td>Permanent magnet</td>
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<tr>
<td>PMSM</td>
<td>Permanent magnet synchronous machine</td>
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<td>PWM</td>
<td>Pulse-width modulation</td>
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<td>SC</td>
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<tr>
<td>SCR</td>
<td>Silicon controlled rectifier</td>
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<tr>
<td>SPM</td>
<td>Surface-mounted permanent magnet</td>
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<tr>
<td>THD</td>
<td>Total harmonic distortion</td>
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<tr>
<td>TRIAC</td>
<td>Triode for alternating current</td>
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<tr>
<td>VSI</td>
<td>Voltage-source inverter</td>
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| gcd     | Greatest common divisor                        |
| lcm     | Least common multiple                          |
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Chapter 1

Introduction

1.1 Background

Electrical machine drives are a mature technology present in most industrial and domestic activities. It is estimated that, of all the electricity produced worldwide, two thirds are destined to power electric motors; mostly asynchronous motors used in industrial applications.

One of the aspects regarding the use of electrical machine drives that most interest presents today and that it is being widely investigated is the issue of faults. In addition to causing a loss of performance (reduced efficiency, increased noise and vibrations), faults in machine drives can compromise the operational availability of the system, imply a significant loss of revenue due to unplanned shutdowns and pose a serious risk to the safety of users, drive operators and other associated equipment.

To date, significant efforts have been made in the prevention, detection and diagnosis of faults in these systems. A quite recent research line that is receiving considerable attention is the design of machine drives that are fault-tolerant. This becomes of the utmost importance in safety critical applications, in which the failure of the system cannot be tolerated. For non safety critical systems, such as those found in production industry, fault-tolerance can also bring benefits regarding increased running times and longer maintenance cycles. In general, systems with very high demands on reliability or safety are likely to benefit from the adoption of fault-tolerant designs.

Permanent magnet machine drives have been traditionally regarded as having a poor fault-tolerant capability compared to other drive types. However, the numerous advantages they offer in terms of increased power density, high efficiency and robust integration makes it important to investigate on how to increase the fault-tolerant capability of these drive systems.

The aim of this thesis is to develop the necessary tools for the design and analysis of fault-tolerant permanent magnet machines. In particular, fault-tolerant designs making use of fractional-slot concentrated-windings and multiphase winding systems are considered.

1.2 Framework

The content of this dissertation is framed within the research activities on electrical machine drives conducted by the Industrial Power Systems Laboratory group of the Department of Electrical, Electronics and Automation Engineering of Tecnun-University of Navarra.

Investigations related to electric drives have been carried out within the research group for more than 15 years; in which much work related to the design, manufacturing, control, characterization and simulation of electrical machines of different topologies has been conducted. The present thesis falls within a new line of research in the group regarding the design and analysis of fault-tolerant machine drives. In
particular, this investigation is oriented towards fractional-slot concentrated-winding multiphase fault-tolerant permanent magnet synchronous machines.

The present work has been funded by the Department of Education, Universities and Research of the Basque Government through the Researcher Formation Program scholarship; reference number BFI-2010-63.

1.3 Aim of the thesis

The aim of this thesis is to develop the necessary tools for the design and analysis of fault-tolerant permanent magnet machines. Specifically, fault-tolerant designs making use of fractional-slot concentrated-windings and multiphase winding systems are considered.

To achieve this dual objective a series of partial objectives are set, as listed below:

- Analysis of the different permanent magnet machine and inverter topologies suitable for fault-tolerant applications.
- Research on criteria to select the most adequate winding configurations for fault-tolerant fractional-slot concentrated-winding machines.
- Development of a design methodology for fault-tolerant permanent-magnet synchronous machines.
- Development of a machine drive model suitable for fault analysis.
- Analysis on remedial strategies to improve the system’s performance after a fault.
- Design and manufacture of a fault-tolerant permanent magnet machine prototype to validate the developed design methodology and test the proposed post-fault remedial strategies.

Although out of the scope of this thesis, it is expected that, in a near future, the developed prototype will allow to test different fault detection and diagnosis techniques as well.

The achievement of these partial objectives will lead to the establishment of comprehensive methodology for the design and analysis of fault-tolerant permanent magnet machine drives.

1.4 Structure of the document

In the following, the structure of this document is outlined. The arrangement of the chapters matches quite exactly the sequence followed during the realization of this thesis.

After this brief introductory chapter, in Chapter 2 the state of the art in fault-tolerant permanent magnet machine drives is reviewed. The most common applications and the main failure mechanisms for these systems are listed and the different fault-tolerant design approaches are discussed.

In Chapter 3 different criteria to select the most appropriate winding combinations for fault-tolerant applications are discussed and own selection criteria based on an analytical calculation of the machine inductances are presented.

In Chapter 4 a design methodology for fault-tolerant permanent magnet machines is proposed and the design of a five-phase machine prototype is described.

A literature review on permanent magnet machine modeling is presented in Chapter 5. Additionally, the development of a five-phase permanent magnet machine drive model suitable for fault analysis is described. The proposed model is used to investigate a number of fault scenarios.
1.4. STRUCTURE OF THE DOCUMENT

Chapter 6 investigates the fault-tolerant operation of the design machine prototype under various fault conditions. For each fault scenario, the implementation of one or several modified control strategies is discussed and the dynamic fault response of the considered drive is evaluated.

In Chapter 7 the measurements and tests conducted on the prototype machine are discussed.

Finally, in Chapter 8 the main conclusions of the investigation, as well as the contributions made to the state of the art in fault-tolerant permanent magnet machines are enumerated. Some future research lines that have not been possible to cope with in this investigation are suggested as well.

As a complement to the study, a number of appendices are attached at the end of this document. In Appendix A a graphical description of the most common multiphase winding systems employed in AC machine drives is given. The convention for defining the electrical angle and the various reference frames used throughout this dissertation is explained in Appendix B. Next, in Appendix C, analytical expressions for the calculation of the slot leakage inductances in rectangular slot machines with either one, two or four layers are obtained. Likewise, formulas for the calculation of the air-gap inductances in radial flux machines are derived in Appendix D. The winding diagram for the designed prototype machine is shown in Appendix E. In Appendix F a general procedure to obtain suitable current references to operate under different fault conditions is outlined. Finally, in Appendix G the formulas needed to implement a current control in the $\alpha - \beta$ reference frame are derived.
Chapter 2

State of the Art in Fault-Tolerant PMSM Drives

The present chapter covers the state of the art in fault-tolerant permanent-magnet synchronous machine (PMSM) drives. First, the most common applications of PMSM drives and the main failure mechanisms of these systems are listed. Then, the general principles of fault-tolerance are reviewed and the different approaches to grant fault tolerance to a PMSM drive are discussed. Since multiphase modular designs, based on a maximum isolation between drive phases, have proven to be among the best choice in order to achieve a sufficient degree of fault-tolerance, the main requirements of this approach, together with the characteristics of the winding topologies commonly employed in these designs and the properties of multiphase systems, are more profoundly examined.

2.1 Present day use of PMSM drives

Modern days permanent magnet synchronous machine (PMSM) drives have found use in a wide variety of high performance applications, ranging from low power appliances to designs in the multi-megawatt rating [1]. Particularly, brushless DC (BLDC) drives with trapezoidal back-electromotive force (back-EMF) waveform machines are more prominent in the lower end of power [2], in which spindle drives, disk drives, auxiliary power units and flywheel energy storage systems are traditional examples.

Operated as motors, the main applications of PMSM drives are found in the following sectors [3–6]:

- Industrial: presses, pumps, manufacturing, robotics, machine tool applications, other servo drive systems.
- Transportation: naval propulsion, automotive applications, wheelchairs.
- Household appliances: washing machines.
- Elevators.
- Aerospace.
- Military.
- Medical applications.
- Nuclear power plants.
- HVAC: heating, ventilating and air conditioning.

In some applications making use of direct-drive configurations, PM machines have become the dominant choice [5]. They have found particular development in hybrid electric (HEV) and electric vehicles (EV); as almost all existing commercial HEVs and EVs make use of NdFeB high energy density permanent magnets (PMs) [7]. Examples of commercial vehicles using PMSM drives for traction are the Toyota Prius,
the Honda Insight, the Chevrolet Volt, the Nissan Leaf and several other [8]. Operated as generators, PMSMs are primarily used in wind power applications [5, 9]. They are also becoming popular in high speed applications, such as in turbochargers and gas turbines [10].

The key advantages that have favored PMSMs to become the preferred choice over other types of electrical machines in many applications are their [3, 6]:

- High power density.
- High torque to inertia ratio.
- High efficiency.
- Simple control.
- High reliability (if compared to brushed machines).

Additionally, they allow an optimal and robust integration of the machine into the drive train system [1]. According to [11], PMSM servo motor drives may be up to a 25% smaller in size than other designs making use of another machine type for the same application.

2.2 Faults in PMSM drives

Even if highly reliable, PMSM drives can be subject to a number of faults; both in the machine and in the power converter. Faults may lead to a performance degradation and even to the failure of the system. In the following section, the different possible fault scenarios a PMSM drive can suffer, together with some fault statistics are presented.

2.2.1 Definition of faults

In the more general sense, a fault can be defined as the unpermitted deviation from its acceptable behavior of a characteristic variable of a system [12]. It is therefore a state that may lead to the instability, malfunction or, if allowed to further develop, to the failure of the system [13]. As a consequence of the failure, loss of production and income, damage to the equipment and danger to the system operators may occur [14].

2.2.2 Possible fault scenarios in PMSM drives

Owing to the complexity of AC drives, many of their elements are susceptible to suffer faults. The possible fault scenarios in a PMSM drive can be broadly classified among electrical and mechanical faults. Potential electrical faults in a PM machine are [15–17]:

- Stator winding open-circuit faults.
- Stator winding short-circuit faults; including inter-turn short-circuit faults, coil to coil faults, line to line and line to ground faults.
- Stator core faults.
- Permanent magnet faults.
- etc.

while the most common inverter and integrated drive related faults are [18, 19]:

- Transistor open-circuit faults (power switch permanently opened).
- Transistor short-circuit faults (power switch permanently closed).
- DC bus capacitor faults.
- Power supply failure or AC line fault in mains supplied rectifiers.
2.2. FAULTS IN PMSM DRIVES

- Sensor failure.
- Control equipment failure.

Finally, the following mechanical faults can occur in a PMSM drive [20,21]:

- Bearing faults.
- Static and dynamic eccentricity conditions (due to a bent shaft or a shaft misalignment).
- Gearbox faults in non direct-driven machines.
- etc.

Additionally, fault conditions arising from the combination of the above faults can be considered as well. One of the characteristics of faults in electrical machines is that they tend to be cumulative, usually one fault inducing another. For instance, a fault in a machine bearing may cause an unbalance in the magnetic field of the machine, leading to an increase in machine losses and, hence, to a stator winding temperature increase. The higher the temperature, the lower the lifetime of the winding isolation that may end up failing and prompting a short-circuit fault between adjacent turns. Inter-turn short-circuits rapidly evolve into major faults, such as winding open-circuit faults, phase to phase short-circuits or phase to ground faults [22].

It is, therefore, necessary that the faults in an AC drive are rapidly detected and that the required remedial actions take place in order to minimize the parasitic effects arising from the faults. In the following, a brief description of the causes and consequences of the most common faults in PMSM drives is given.

Winding open-circuit faults

Winding open-circuit faults are among the most common faults in AC motor drives [23, 24]. They are caused as a consequence of the internal interruption of the motor windings, mechanical faults at the machine terminals, mechanical stresses on the connectors that link the motor and the inverter or by an electrical failure on an inverter phase leg [25,26]. In a standard 3-phase drive, an open-circuit fault leads to the inability to produce a constant magnitude rotating magnetic field which, in turn, causes large torque pulsations, mainly at twice the frequency of the supply signal [27].

Winding short-circuit faults

It is believed that the majority of stator winding faults begin as insulation breakdown failures that lead to a short-circuit between several turns of a phase winding [22]. This causes a large circulating current to flow inside the shorted turns that generates a high amount of heat to be locally released within a short time [28]. If undetected, inter-turn short-circuit faults rapidly propagate to neighboring turns, culminating in major faults such as coil to coil, phase to phase, phase to ground or open-circuit faults. Ultimately this causes a failure of the motor, unless the proper remedial actions are taken.

The physical mechanisms that provoke the aging of the insulation and can finally lead to an insulation breakdown are [20,29]: thermal overloading and cycling due to an improper heat removal, high currents during motor starting, short-circuits, etc.; chemical reactions due to contamination by hydrocarbons, moisture, dirt, etc.; electrical discharges and voltage spikes in variable speed drives; mechanical stresses; etc.

Inter-turn short-circuit faults can lead to the breakdown of the system in a matter of a few seconds or minutes [30]. In [31], it was shown that it takes less than 2 seconds for a single inter-turn fault to evolve into a major fault for a typical 15 kW induction motor. If the short-circuit fault involves a failure of the phase to ground insulation, large currents that may irreversibly damage the core of the machine are produced, which will most probably end up in a failure of the system [32].

Other winding related faults that may occur are winding short-circuit faults involving the whole phase winding. These may happen as a consequence of the closing of all the high side or low side switches of the inverter, a DC bus short-circuit, a short-circuit between machine terminals in the terminal box...
and physical damage on the cables connecting the motor and the inverter [33, 34]. The first cause, the closing of all the high side or low side inverter switches may be due to an error in the control logic. However, it is more likely that it is produced in order to protect the drive against other more severe fault conditions [34].

Stator core faults

Stator core faults are rare if compared to stator winding faults. According to [35], they may account to 1% of all machine faults in low and medium voltage motors. Although stator core faults may be a minor concern in small machines, their repair cost is higher than that of stator winding faults, as it usually requires that the whole core is replaced [14]. The main causes of stator core faults are [36,37]: electrical arcs and stator core melting due to winding short-circuit faults, core end-region heating resulting from the axial flux in the end-windings, interlaminar insulation damage, vibrations due to the dilatation and relaxation of the lamination stack, manufacturing defects in the laminations, etc.

Permanent magnet faults

The most common permanent magnet faults involve the demagnetization, the cracking or the breaking of the magnet material. The demagnetization of the PMs used in synchronous machines can occur as a combination of thermal, electromagnetical, mechanical and environmental stresses [38].

Elevated temperature levels during a long period of time cause changes in the metallurgical structure of the material, which loses its magnetic properties and its ability to be remagnetized. Also, if the magnet material exceeds its Curie temperature, the magnetization is lost, although the material may be remagnetized as long as the metallurgical structure remains unaltered. For AlNiCo, SmCo and NdFeB magnets, the Curie temperature is higher than the temperature at which significant metallurgical changes occur, while ceramic magnets exhibit the opposite behavior [38]. Chemicals and humid environments favor the corrosion and oxidation of rare-earth magnet materials, which also cause changes in the metallurgical structure. The structurally altered parts exhibit lower remanence and coercitivity levels and, hence, are more prone to demagnetization.

Regarding electromagnetical stresses, it is well known that PMs can be demagnetized by external magnetic fields applied in a direction opposite to that of the material magnetization. The magnetization curve of a magnet material exhibits a knee point below which the magnet is irreversible demagnetized. This point usually falls in the second or third quadrant of the material’s $B-H$ characteristic. The composition of the magnet and its temperature determine the position of the knee point.

Other possible causes of PM demagnetization include the decrease in magnet strength due to domain relaxation, physical damage due to vibrations, shocks and mechanical forces, etc. [38]. The consequences of the degradation of the magnets in a PMSM are a reduction in the flux density values and, therefore, an increase in the machine currents for the same load. This leads to a reduction in the machine’s efficiency, a temperature increase, a decrease of the insulation lifetime, etc.

Transistor open-circuit faults

One of the most common inverter faults involves the permanent opening of a power semiconductor switch due to damage to the transistor itself or the control logic commanding the gate signals [39]. Some causes of open transistor faults are [40,41]: the lifting of the semiconductor bonding wires due to thermal cycling, rupture of the semiconductor material due to short-circuit faults or problems in the gate control signal. The authors in [25] cite damages to the transistor gate or in the opto-couple driver device due to voltage bursts and faults in the control logic commanding the gate signals by electrostatic discharge disturbances as probable mechanisms for open transistor faults.

As a consequence of the permanent opening of a power switch, the current in the corresponding inverter leg is allowed to flow only during part of electrical cycle, either through the other transistor or through the associated free-wheeling diode, leading to a DC component in the phase currents and pulsations in the torque waveform [42]. The DC currents also generate unequal stresses in the high side and low side
transistors and, hence, can lead to secondary inverter faults. According to [43], as they are not critical faults, transistor open-circuit faults may not trigger the inverter shutdown and may remain undetected for a long time.

**Transistor short-circuit faults**

This kind of fault usually occurs as a consequence of a single power semiconductor switch being permanently damaged. The damage may be due to a short-circuit fault in the associated free-wheeling diode, provoked by an overcurrent or an excessive reverse voltage across the diode. The presence of impurities in the fabrication process may also trigger the short-circuiting of the power semiconductor switch. Finally, erroneous control signals caused by driver malfunction, auxiliary power supply failure, electromagnetic disturbance or control software errors may trigger this kind of fault as well [25,33,42].

The short-circuiting of an inverter transistor is a serious fault condition that demands immediate remedial action from the drive’s protection circuit. In order to prevent a DC bus shoot-through, at least the other switch in the same leg has to be opened within a few $\mu$s [42]. As a consequence of the fault, high magnitude DC current components that are mainly limited by the stator resistance appear in all the phases [25]. The fault poses a risk in terms of current withstand capability of the remaining healthy transistors, winding overheating and PM demagnetization. An extremely high and oscillating torque is produced as well that can also jeopardize the integrity of the bearings, the shaft, the coupled load, etc.

**DC bus capacitor faults**

Electrolytic capacitors used in power inverters slowly degrade as a consequence of thermal aging, electrolyte vaporization, etc. [44]. According to [18], these elements may have shorter lifetimes than that of the power semiconductor switches. In addition to growing faults, total breakdown faults of the DC capacitors may occur as well. In the latter case, the inverter is usually shut down due to the insufficient smoothing of the DC bus voltage [18].

**Power supply failure**

This fault may happen as a consequence of the DC bus of the variable frequency drive being disconnected from its power supply. The PMSM is supplied just by the DC link capacitor, that is quickly discharged as the electric energy is transferred from the capacitor to the electrical machine [25]. The capacitor voltage and the delivered torque steadily decrease until the rotor stops. Although it is rare that this fault causes subsequent damage to the equipment, in safety critical applications, in which the drive must continue operating uninterruptedly, this situation may not be bearable and a redundancy approach may be necessary.

**Sensor failure**

Modern PMSM drives are equipped with a number of sensors, such as current sensors measuring the branch currents, DC bus voltage sensors, position and/or speed sensors, etc. All of these can be subject to failures, leading to an incorrect estimation of the measured variables and, hence, to control instabilities, reduced performance, etc.

The easiest way of avoiding the negative consequences arising from a sensor fault is to have a sensor redundancy. Strategies that allow AC drives to operate with faulted sensors with some minor performance degradation have been proposed in the related literature as well [45,46]. In most cases, the variable to be measured by the faulted sensor is estimated by a dedicated observer instead.
CHAPTER 2. STATE OF THE ART IN FAULT-TOLERANT PMSM DRIVES

Mechanical faults

The most common mechanical faults in electrical machines are bearing related [20]. Most electrical machines employ rolling bearings as mechanical element supporting the shaft. Rolling bearings are constituted by an inner and outer race through which a set of balls or rolling elements runs. Even under normal operating conditions with a constant and balanced load, periodical stresses are applied onto the bearings. These stresses erode the bearings and lead to fatigue failures of the elements [29]. Other common causes for bearing faults are [29, 47, 48]: pollution, corrosion, poor lubrication, improper mounting, overloading, unbalanced loads, excessive vibrations, induced eddy currents, etc. Among the aforementioned fault mechanisms, the lack of a proper lubrication is the most critical factor. It is estimated that around 40% of the premature failures in motor bearings are caused by an improper lubrication [49].

Bearing faults, that usually manifest themselves as asymmetry or eccentricity faults, may lead to the following parasitic effects: magnetic unbalance, additional noise and vibration, increased losses and efficiency reduction, damage onto the shaft, the machine and/or the coupled elements, and finally, to the inability to operate the machine. Condition monitoring of the bearings, together with a predictive maintenance plan and periodical substitution of the rolling elements, helps reduce bearing related faults and the derived negative consequences.

Other possible mechanical faults include the wear of the gearbox elements in non direct-driven machines, misaligned shaft faults, etc.

2.2.3 Fault statistics of AC drives

In the following section some statistics on the reliability of AC drives are presented. They are taken from a number of surveys found in the related literature. The studies are divided into fault statistics in electrical machines and faults in power inverters. General failure rates for different components in 3-phase drives used in aircraft applications are given in Table 2.1 [24].

<table>
<thead>
<tr>
<th>Cause of failure</th>
<th>Failure rate/phase (per hour)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Winding open-circuits</td>
<td>$1.3 \cdot 10^{-6}$</td>
</tr>
<tr>
<td>Connections open-circuits</td>
<td>$1.0 \cdot 10^{-6}$</td>
</tr>
<tr>
<td>Other open-circuit faults</td>
<td>$0.4 \cdot 10^{-6}$</td>
</tr>
<tr>
<td>Winding short-circuits (between phases)</td>
<td>$6.7 \cdot 10^{-6}$</td>
</tr>
<tr>
<td>Connections short-circuits</td>
<td>$1.0 \cdot 10^{-6}$</td>
</tr>
<tr>
<td>Other short-circuit faults</td>
<td>$0.4 \cdot 10^{-6}$</td>
</tr>
<tr>
<td>Total electrical failures</td>
<td>$6.6 \cdot 10^{-5}$</td>
</tr>
<tr>
<td>Power supply</td>
<td>$5.4 \cdot 10^{-5}$</td>
</tr>
<tr>
<td>Power electronic controller</td>
<td>$8.5 \cdot 10^{-5}$</td>
</tr>
<tr>
<td>Control signal</td>
<td>$1.3 \cdot 10^{-5}$</td>
</tr>
<tr>
<td>DSP failure</td>
<td>$1.0 \cdot 10^{-5}$</td>
</tr>
<tr>
<td>Total electronic failures</td>
<td>$1.5 \cdot 10^{-4}$</td>
</tr>
</tbody>
</table>

Table 2.1: Motor and inverter failure rates in 3-phase drives [24]

Hence, according to the data in [24], inverter faults are more prevalent than machine faults and winding open-circuit faults more likely than short-circuit faults between phases. It must be noted that the cited reference does not provide information on the type of machine and voltage and power ratings. However, it allows to grasp an overall picture regarding faults in AC drives.

Faults statistics of electrical machines

Thought the years, various surveys on the reliability of electrical machines have been conducted. In these surveys, a large number of electrical machine operators have been asked on the applications, operating factors, failure types, failure frequency, causes, impacts, etc. The results from some of these surveys are
presented in Figure 2.1. As it can be appreciated, the majority of electrical machine failures arise from faults in the bearings and the stator windings.

![Fault statistic distribution in electrical machines](image)

Figure 2.1: Fault statistic distribution in electrical machines

The survey in [35], carried out by the General Electric Co. in 1982, covered about 4800 low and medium voltage powerhouse motors, with ratings greater than 100 hp. Most of the analyzed motors (around 97%) were 3-phase squirrel cage induction motors operating in different branches of industry. Among the stator related faults (37%), 23% were due to failures of the ground insulation and 4% due to turn insulation faults. Rotor related faults were mostly due to cage faults in induction machines (7%). The study in [35], was extended in 1986 by including 1527 new motors in the previous analysis [52, 53], leading to a total amount of 6312 motors, from which 1474 failures on 1052 failed motors were reported.

A second set of studies was conducted by the Motor Reliability Working Group of the Power Systems Reliability Subcommittee of the IEEE Industry Applications Society in 1985 [50, 54, 55]. The purpose of the study was to identify failure data and the effects of preventive maintenance on various classes and applications of industrial and commercial motors. The focus was on large motors with ratings greater than 200 hp. The largest group of analyzed motors were induction and synchronous machines, that showed similar failure rates for the same voltage class. Among many trends, it was found that, in slow speed motors (up to 720 rpm), the main cause of machine failure were failures in the stator windings.

In 1995, a survey on cage induction motors operating in the offshore oil industries, petrochemical industries, gas terminals and refineries was conducted [51]. Specifically, 2596 induction motors with ratings in the 10 – 1000 kW range were analyzed. The survey presented the faults sorted according to supply and motor data, driving conditions, electrical protection, maintenance, etc. A subset of 483 high-voltage induction motors above 100 kW and up to 8000 kW were analyzed by the same authors in [13]. Similar fault statistic distributions were reported, although stator winding faults amounted to be around a 25% of all the machine faults.

By comparing the results from the different surveys, it can be deduced that the fault statistics depend largely on factors such as the specific application, the type of machine, the power and voltage ratings, the speed, the environment, the type of enclosure, etc. Most of the analyzed machines are large induction and synchronous motors. Additionally, the collected results refer to failed machines; meaning that faults that
do not involve a total failure of the system are not reflected in the studies. However, the cited surveys allow to get a general idea of the most common faults in electrical machines.

It is important to note that the cited references are quite old. Modern day AC machines are primarily supplied by variable frequency inverters that put higher stresses on the winding insulation. Most probably, the fault distribution picture looks quite differently in modern VSI supplied small and medium power PMSMs.

**Faults statistics of power converters**

A number of surveys on the reliability of variable speed drives have been carried out over the years as well. For instance, a survey from 1995 [56] found that faults in the control circuit of the drive may account to 53% of the total faults; 38% and 7% of the faults corresponding to the power part and external auxiliaries, respectively. The survey reviewed variable frequency drives in industry with power ratings ranging from 1 to 200 kW and different applications and environments. The survey analyzed mainly frequency drives with an open loop control and a constant \( V/f \) ratio. According to [18], the high control circuit fault rate in the survey may be due to the lower reliability of the early days PWM control. Results from other newer surveys are presented in Figure 2.2.

![Fault statistic distribution in power converters](image)

(a) Survey results from [57], 2007

(b) Survey results from [58], 2011

Figure 2.2: Fault statistic distribution in power converters

The study in [57] reviewed 200 products from 80 companies; mainly printed-circuit-board-based micro-electronic circuits. The other survey [58], aimed at evaluating the reliability of power electronic converters in the general, was subjective and carried out among a number of semiconductor manufacturers, integrators and users. The industry branches incorporated in the study included aerospace, automation, motor drive and utility power applications.

Both studies indicate that the power semiconductor devices and the DC bus capacitors are among the most fragile components. The survey in [58] concluded that main stresses that lead to component failures were related to the environment, transients and heavy loads.

### 2.3 Fault-tolerance in AC drives

#### 2.3.1 Introduction to fault-tolerance

**Definition of fault-tolerance**

The term fault-tolerance can be defined in a number of ways [20, 59]. The basic idea of fault-tolerance is that a fault in an individual component or sub-system does not cause an overall system malfunction [60]. Sometimes, depending on the type of fault, a degradation on the system performance may be accepted as long as the fault does not cause a complete system failure. Depending on the application, different number of faults may be required to be tolerated before the system malfunctions.
The term fault-tolerance is common in safety critical applications, that are defined as those whose failure may cause death or serious injury to people, loss or severe damage to equipment or environmental harm [16]. The concept is also usually related to measures being taken to limit the impact of different faults on the overall performance of a system [20].

Need of fault-tolerance

Fault-tolerance is required in systems in which the continuous availability of the system is needed and in which the consequences of a malfunction are more expensive than the cost of preventing such a malfunction [60]. In other applications, legislation or safety requirements can demand the use of fault-tolerant systems. In general, systems with very high demands on reliability or safety are likely to benefit from the adoption of fault-tolerant designs. These includes life support apparatus, vehicles, military applications, X-by-wire systems, continuous production industrial processes, etc. [20].

Degrees of fault-tolerance

The characteristic of fault-tolerance is not an absolute. No system can be made truly tolerant to any possible combination of faults [60]. There will always be a combination of faults that will lead to a malfunction of the system. The required degree of fault-tolerance is dependent on the application and must be defined during the system’s design stage. Usually, granting fault-tolerance to a system adds complexity and cost. Therefore, only the most likely and/or severe faults should be considered [20].

Principles of fault-tolerance

The general principles of fault-tolerance have been covered in [20, 60]. These are:

- Redundancy.
- Partitioning.
- Fault detection and fault diagnosis.
- Reconfiguration.
- Repair and maintenance.

Redundancy  The concept of redundancy involves the addition of extra elements or subsystems that are able to take over in case one component fails. The redundant elements may be operating during normal operation sharing the workload or they may be put into service when a fault occurs. This increases reliability and availability at the price of generally adding cost to the system [60]. Another price to be paid for the higher availability is a higher rate of repair due to the higher component count, which may be incorrectly perceived as a reduced reliability.

The authors in [61] distinguish two levels of redundancy: full redundancy and half redundancy. The first one involves that the system operates normally in case of any single component failure. The second one occurs when a performance degradation is acceptable in case of a fault. The level of redundancy is usually a trade-off between availability, complexity and cost.

Partitioning  A fault-tolerant system requires that a failure in one component does not cause the failure of additional elements in the system. There must be an isolation between elements in order for the fault to be contained within the failed unit. This is usually accomplished by partitioning the system in independent subsystems [60].
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Fault detection and fault diagnosis  In order for a system to be fault-tolerant, any possible fault must be detected and diagnosed; that is, its type, position, hazard level and cause determined in order to take the appropriate remedial actions [20]. If a failure is not detected and repaired, the system is no longer redundant and an upcoming fault may cause the system’s malfunction.

Reconfiguration  Most commonly, the isolation of a faulty element or the switching from a faulted element to a healthy redundant one requires the system to be reconfigured. This can include the reconfiguration of both the hardware and software of the system. The set of measures that are usually taken in order to limit the impact of the different faults on the overall system falls in this category [20].

Repair and maintenance  Of course, the primary objective of a fault-tolerant system is to allow the system operation until the faulted element can be repaired or substituted [62]. In the long run, repair must be conducted on the system in order to reestablish its full capability [20]. The mitigation of the fault by the previously discussed actions can only be a short/medium time solution. Maintenance actions that return the system to its designed operating conditions are needed.

2.3.2 Fault tolerance applied to PMSM drives

When applied to AC drives, the term fault-tolerance has usually two distinct acceptations depending on the application [63,64]. The first acceptation is that the drive must be able to withstand a temporary fault (e.g. a short-circuit fault) without being damaged. The second meaning is that the drive must be capable of operating in fault-conditions with a minimum level of performance. Since operation under fault conditions represents an abnormal state, efficiency should be a secondary concern as long as the uninterrupted operation of the drive can be maintained [19]. The first definition requires that the different drive elements have a high degree of isolation between them; while the second one is usually accomplished by having multiple independent phases or redundant drive elements [65].

Among the various types of electrical machines, PMSMs are among the most critical ones when it comes to granting fault-tolerance to an AC drive. The presence of magnets that cannot be turned off at will makes it difficult to de-excite the machine. This has made engineers hesitant to employ PMSMs in safety critical applications in the past and has motivated substantial work to be carried out on both the machine and the power converter side in order to increase the fault-tolerance of PMSM drives [22].

2.3.3 Examples of fault-tolerant PMSM drives

Several examples of fault-tolerant PMSM drives covering a wide range of applications have been reported in the related literature:

- **Aerospace**: fuel pump [62,66–69], flap and slat actuators [70–73], cabin pressure control servo-drive actuator [12], drives more-electric-aircraft in general [74–81].
- **Automotive**: traction [82–87], X-by-wire systems (braking, steering, etc.) [65,88,89].
- **Generation**: Wind turbine [90–93].
- **Marine**: Propulsion [27,61,94,95].

2.4 Methods for granting fault-tolerance to a PMSM drive

Several methods have been proposed in the related literature to grant fault-tolerance to a PMSM drive. Generally speaking, these methods seek the fault-tolerance by partitioning the drive in various independent elements and exploiting the redundancy; be it the redundancy of the entire drive system, redundancy at the power converter level or in the electrical machine. In the following section, the different methods are briefly discussed and compared in terms of cost, reliability, performance under fault conditions, etc.
2.4. METHODS FOR GRANTING FAULT-TOLERANCE TO A PMSM DRIVE

2.4.1 Methods based on redundant machine and inverter structures

One of the simplest techniques to grant fault tolerance to a PMSM drive system is to employ redundant machine and inverter structures. These configurations have been considered by a number of authors [20, 65]. For instance, [61] discusses different redundant drive configurations for electric marine propulsion applications. Broadly speaking they can be classified among the following:

- Drive trains in parallel.
- Direct drive multiple machines on the same shaft.
- Drive trains connected through a gearbox/belt drive.
- Partial redundancy of drive elements.

Drive trains in parallel

Different parallel drive trains can be used when the mechanical load can be supplied by independent axles. This is the case for ship propulsion systems, for which the power is commonly shared among different motor drives [61]. These structures are highly redundant and allow the system to operate in case of failure of any of the individual drive trains. As a drawback, the system’s cost, weight and volume are doubled. An example of such a configuration is depicted in Figure 2.3.

![Figure 2.3: Redundant configuration with drive trains in parallel (adapted from [61])](image1)

Direct drive multiple machines on the same shaft

When the mechanical load cannot be supplied by independent drive trains, multiple machines can be mounted onto the same shaft in order to achieve the desired degree of redundancy. This configuration is exemplified in Figure 2.4.

![Figure 2.4: Redundant configuration with multiple machines on the same shaft (adapted from [61])](image2)

Such a configuration has been proposed, for instance, in [16, 74], in the context of the all-electric-aircraft. It must be noted, though, that, in the previous reference, each machine drive segment is designed to be fault-tolerant itself. The proposed system employs redundant position sensors, two independent inverters and controllers and a back-up controller for supervisory and fault-detection actions. The same strategy is investigated in [88] and [20] for electrical power steering applications; although the latter reference considers the addition of a clutch in order to disconnect both machines in case of a mechanical fault.

Compared to the configuration with parallel drive trains, the present strategy reduces the redundancy cost in terms of independent shafts, but there is an increased complexity and cost of the drive circuit due to the multiple and independent controllers [16]. Since, for a PMSM drive under short-circuit conditions, a braking torque due to the fault is generated, it must be ensured that the braking torque is low enough to allow the movement of the common shaft by the remaining healthy drive segment [88].
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Drive trains connected through a gearbox/belt drive

The concept in this case is similar to the previous configuration, although the machine drive segments are connected in parallel, instead of in series. Hence, the same considerations regarding the braking torque under short-circuit conditions must be kept in mind [65]. The use of a gearbox introduces a penalty in terms of additional weight, volume and cost and there is a reduction in the system’s efficiency. The overall reliability of the system may be compromised as well due to faults in the mechanical coupling element. An example of such a configuration is depicted in Figure 2.5.

Figure 2.5: Redundant configuration with drive trains coupled through gearbox (adapted from [61])

This configuration has been discussed, for instance, in [12] for the cabin outflow valves in a plane. Instead of a gearbox, a belt drive with two clutches, one per motor, can be used [20]. The use of clutches allows to isolate a faulty motor entirely from the load and grants fault-tolerance to mechanical faults. Additionally, a clutch can be used to eliminate the parasitic braking torque arising from a short-circuit fault by disengaging the faulted drive segment.

Partial redundancy of drive elements

One last choice is to have a partial redundancy of the system by doubling just one of the components. Figure 2.6 represents an example of such a configuration, in which one electrical machine is supplied by two inverters [65]. Under healthy conditions, the machine is powered by a single inverter while the other remains unconnected; where as, in case of fault, the inverters are swapped and the drive continues to operate normally. Although in this configuration just a single component is doubled, the system cannot be considered truly fault-tolerant, as a failure of the non-redundant element leads to a total system breakdown. Additionally, unlike in previous configurations, the machine and inverters must be rated to deliver full power.

Figure 2.6: Partial redundancy of drive system (adapted from [65])

2.4.2 Methods based on modified inverter topologies

A second group of methods aimed at providing fault-tolerance to a PMSM drive system are based on a modification of the standard inverter structure, so that only certain power converter elements are made redundant. The most common structures have been thoroughly discussed in [19]. These include the use of:

- Switch-redundant topologies.
- An additional leg connected to the neutral point.
- An additional redundant leg.
- Full H-bridge/cascaded inverters.
2.4. METHODS FOR GRANTING FAULT-TOLERANCE TO A PMSM DRIVE

Some other options, such as the adoption of a delta connection, use of direct AC/AC matrix converters, employ current servo amplifiers, etc. have been proposed in the related literature as well. Generally speaking, the cited methods require a reconfiguration of both the hardware and the software of the system once a fault has occurred. In the following subsections, the different methods are briefly explained.

Switch-redundant topologies

These structures have been introduced in [96] and [97] to accommodate for winding open-circuit and power switch short-circuit faults, respectively. A combined structure, described in [19], is shown in Figure 2.7.

![Figure 2.7: Switch-redundant inverter topology (adapted from [19])](image)

The modified inverter configuration makes use of 4 Triodes for Alternating Current (TRIACs) or back-to-back connected Silicon Controlled Rectifiers (SCRs) that allow to reconfigure the drive after a fault. This way, fault-tolerance against winding open-circuit and power switch open and short-circuit faults is gained [19]. Under a winding open-circuit fault, the TRIAC connecting the neutral point to the midpoint of the DC link is closed and the currents in the healthy phases are scaled by $\sqrt{3}$ and shifted by $30^\circ$ in order to preserve the main harmonic of the armature air-gap MMF. The drive is able to deliver rated torque, but, as only half the DC bus voltage can be applied to the remaining healthy phases, only half the rated power is available [19]. Additionally, some issues regarding minimum capacitance size and DC midpoint balancing problems may arise [96].

In case of a power switch open or short-circuit fault, the following mitigation procedure is applied: the corresponding switch in the same leg is turned off and the TRIAC connecting the faulted phase branch to the midpoint of the DC link is closed. In both cases, the machine phase associated to the faulted transistor is isolated from the faulted inverter pole and connected to the midpoint of the DC link (in case of a switch short-circuit faults, additional fuses not represented in Figure 2.4 are needed). The resulting state is that of a four-switch 3-phase inverter. The drive, operating under these conditions, is able to deliver full torque, but the field-weakening mode is entered at half the speed compared to the pre-fault state [19].

Additional leg connected to the neutral point

The use of an additional inverter leg connected to the neutral point of the machine’s star connection has been proposed in [98, 99]. A scheme for such a modified inverter structure is shown in Figure 2.8. Under healthy operating conditions, the extra inverter leg can remain inactive or be used in order to gain an additional degree of freedom and improve the drive’s performance [98]. After a phase fault, the corresponding machine terminal and the inverter pole are disconnected one from another and the fourth leg starts to operate, supplying the neutral point of the star connection and granting, thus, fault-tolerance against winding open-circuit and power switch faults. In order to disconnect the fault, bidirectional switches before the machine terminals [99] or additional SCRs, fuses and dedicated capacitors can be employed [98]. In the latter case, the drive is also made tolerant to the short-circuit fault of both power switches in the same leg.
The current references to be applied after a fault are the same as those for a winding open-circuit fault and a switch-redundant topology. In this case, the drive is able to deliver $\frac{1}{\sqrt{3}}$ of the rated power after a fault [19].

**Additional redundant leg**

Another possible modification to the standard 3-phase half-bridge inverter structure is to add a redundant inverter leg that can be connected so as to substitute any of the other legs in case of fault [98, 100]. The disconnection of the faulted phase can be attained by the procedures cited above [98, 99]. The present modified drive structure is able to deliver full rated power under any single power switch fault, at the cost of a higher semiconductor device count. A scheme for such a modified inverter structure is shown in Figure 2.9.

An alternative structure to the previous configuration is proposed in [101], in which the inverter redundancy arising from having the power split among various machine drives is used to eliminate the additional inverter leg. The proposed drive structure is shown in Figure 2.10. Under healthy conditions, the system behaves as two independent 3-phase PMSM drives; whereas, under a power switch fault, the system can be reconfigured as a five-leg topology, in which each PMSM has two individual legs and one common leg shared with the other PMSM. In order the system to be able to deliver full rated torque at fault conditions, the power switches must be overrated by two. Alternatively, a strategy based on electrically displacing the phase current angles in the two PMSMs is proposed in [101]. This way, full
rated torque can be delivered without leg current overrating, but the speed is reduced to around 50% of the pre-fault state value.

![Diagram of redundant PMSM and VSI configuration proposed by [101]](image)

**Figure 2.10: Redundant PMSM and VSI configuration proposed by [101]**

**Full H-bridge/cascaded inverters**

The use of single H-bridge inverters to supply multiple independent phase windings was first proposed in [102] for an induction machine. This very same inverter structure has been latter considered in the fault-tolerant operation of PMSM drives [62, 66]. An scheme for a full H-bridge/cascaded inverter configuration is shown in Figure 2.11, where only a single phase drive segment is depicted.

![Diagram of single phase winding supplied from an H-bridge inverter (adapted from [19])] (image)

**Figure 2.11: Single phase winding supplied from an H-bridge inverter (adapted from [19])**

Just by itself, the full H-bridge inverter configuration can only grant fault-tolerance against open-circuit faults (winding and power switch). By adding a TRIAC or another dedicated device in series with each phase winding, isolation of the faulted phase is possible and the fault-tolerant capability is improved [19]. Since there is no electrical connection between machine phases, a zero sequence current is allowed to flow through the machine and the rotating armature MMF can be preserved after a single phase fault. Regarding power switch open-circuit faults, operation under fault conditions is limited to the speed at which the peak line to line back-EMF is lower than the DC bus voltage [19].

Due to the increased device count, it is argued in [19] that, by using this power converter topology, inverter switching losses increase during normal operation. However, according to [16], since the semiconductor devices must withstand the phase voltage rather than the line voltage, the power device voltage ratings are reduced and switching losses may even decrease; which, in turn, reduces the heat sink requirements,
Adoption of a delta connection

Instead of employing the more common star connection for a 3-phase AC system, the adoption of a delta connection has been proposed in [21,103] for an induction motor drive. In case of a single phase winding open-circuit fault, the remaining healthy phase currents can still be independently controlled and a low ripple torque can be produced, provided the proper control reconfiguration is adopted. While the method is simple and requires no additional hardware, the system in this case is only made fault-tolerant to single phase winding open-circuit faults.

Direct AC/AC matrix converters

The use of conventional VSIs requires bulky, heavy and unreliable capacitors for the DC link, that may lead to power quality management issues [78]. As an alternative, fault-tolerant inverter structures based on modified AC/AC matrix converters have been proposed in the related literature as well. Matrix converters offer several advantages, such as bidirectional power flow capability, compact size, sinusoidal input and output currents, unity power factor operation ability, long lifetime under harsh environments, etc. [104, 105]. On the other hand, they have limited voltage transfer ratios, the number of power semiconductor devices is high and they require complex control algorithms and expensive solutions for overvoltage protection [106].

The modified structures for fault-tolerant direct AC/AC matrix converter topologies are analogous to the ones proposed for VSIs. In particular, the following configurations have been proposed:

- Switch-redundant matrix converters with the neutral point of the machine connected to the neutral point of the grid supply in case of fault [106,107].
- Switch-redundant matrix converters in which the faulted phase machine terminal is connected to the neutral point of the grid supply [108].
- Matrix converters with an additional phase leg connected to machine’s neutral point [104,107]. In particular, the authors in [105] present two different types of fault-tolerant AC/AC matrix converters with a redundant phase leg: namely a direct matrix converter and an indirect one.

In general terms, most of the work related to fault-tolerant direct AC/AC matrix converters has been focused on granting fault-tolerance against winding open-circuit and open switch faults, although the switch-redundant topology in which the machine terminals are connected to the neutral point of the grid supply has been shown to be tolerant to switch short-circuit faults in [109], for instance.

Use of current servo amplifiers

Instead of using a VSI or a matrix converter, the use of current servo amplifiers is proposed in [3] to grant fault-tolerance to a PMSM drive. The three current servo amplifiers [110] control each phase current independently, making it possible to introduce optimal phase currents that eliminate the torque ripple and minimize the stator copper losses, even under a single-phase open-circuit fault. Although the proposed fault-tolerant control architecture can be applied to both BLDC and PMSMs, it requires that the neutral point of the machine’s star connection is grounded.

Fault-tolerant multilevel inverters

Since in multilevel inverters the power is divided among more power semiconductor devices, it has been suggested that these topologies may be inherently more tolerant to internal faults than conventional
three-phase two-level inverters [111]. However, the higher number of transistors involves an increased failure rate and fault-tolerant topologies have been proposed for multilevel inverters as well [111].

For recent reviews on fault-tolerant two-level and multilevel 3-phase VSI topologies, the interested reader is referred to [112,113].

### 2.4.3 Methods based on modular machine designs

Finally, the last group of methods that allow a PMSM drive system to become tolerant to a number of faults are based on having modular machine designs. Two distinct approaches are usually considered; in the first one, each machine phase is regarded as a single module and the drive is designed so that every phase operates independently. This method demands that each inverter and machine phase is isolated from all the other phases [66]. The second procedure is to have various sets of isolated 3-phase windings that are independently supplied. These last structures are usually referred to as split-phase and multiple-stator designs.

An schematic view of these two fault-tolerant approaches is shown in Figure 2.12.

![Diagram of fault-tolerant approaches based on modular machine designs](image)

Figure 2.12: Fault-tolerant approaches based on modular machine designs (adapted from [24])

#### Single phase modular configurations

In order for each drive phase to operate independently and behave as a single module, it is required that it is magnetically, electrically, thermally and physically isolated from the rest of phases [62]. These isolation requirements must be fulfilled both at the machine and at the power converter side. In this way, the failure of a phase has a minimal impact on the other phases, which can continue to deliver their share of the total power. To maintain a high level of performance after a fault, usually multiphase drive systems with more than 3-phases are used [69]. Additionally, in order to improve the drive’s response to the loss of a phase, modified control strategies are commonly applied [114].

The principles of this method applied to PMSM drives and the penalties arising from it have been covered in [66,67]. In general terms, single phase modular configurations make use of fractional-slot concentrated-winding machines supplied by full H-bridge inverters. If a drop in the output torque after a fault is acceptable, this method involves no penalties in terms of increased machine size, weight or volume. However, the use of single phase H-bridge inverters implies a higher device count and a modest increase in the volt-amperes rating of the power converter.

The details of this method are further discussed in a latter section.

#### Split-phase/multiple-stator designs

In these drive configurations, the machines are designed with multiple sets of isolated 3-phase windings, supplied each by an independent 3-phase power converter. In case of a phase fault, the corresponding 3-phase winding is isolated and the drive continues to operate with a reduced power. Compared to
CHAPTER 2. STATE OF THE ART IN FAULT-TOLERANT PMSM DRIVES

the previous method, the use of multiple-stator designs has the advantage that an adequate post-fault operation can be achieved with no software reconfiguration. However, the reduction in power due to the loss of a single phase is larger. Examples of these topologies can be found in ship propulsion applications, wind power generation, EV components, etc. [61,115,116].

In the related literature, several examples of PMSM drives with two [117–119], three [80], four [120] and more [87] 3-phase windings can be readily found. The multiple stator windings can be arranged so that the EMF phasor distribution of each 3-phase set is the same or so that there is an angular displacement between the EMF phasors belonging to different sets [65]. In the first case, the windings are supplied by currents in phase, while, in the second case, it is common to have the winding sets shifted by an electrical angle of $\frac{\pi}{3n}$ and, hence, supplied by currents with a delay of $\frac{1}{6n}$ of a period, where $n$ is the number of 3-phase sets. When distributed windings are used, the first topology leads to coil arrangements with a lower contact between winding sets and a lower mutual coupling between 3-phase windings, where as the second arrangement usually leads to overlapped windings [65]. By having the windings sets displaced $\frac{\pi}{3n}$ electrical degrees, the system can be supplied as a 3n-phase system, leading to an increase in the winding factor and a reduction of the air-gap MMF harmonic content. Thus, lower rotor losses and a lower torque ripple can be obtained in the latter case [65,121].

2.4.4 Comparison among the different fault-tolerant approaches

A comparative among some of the previously discussed methods is collected in Table 2.2. The methods are compared in terms of power converter cost, machine cost, number of power semiconductor devices and power capability after a single phase open-circuit fault (switch open or winding open-circuit). A standard 3-phase, half-bridge VSI supplied, non fault-tolerant PMSM is taken as a reference. The cost of the power converter is calculated according to the total volt-amperes rating of the inverter, defined as the product of the number of semiconductor devices times their peak voltage and current capabilities [122]. In the calculations, it is assumed that the cost of a TRIAC is that of a power semiconductor switch (most commonly a IGBT or MOSFET) [19]. The cost of fuses, gate drive circuits and other additional electrical and mechanical elements needed in the different fault-tolerant topologies is ignored. The complexity of the power converter is measured by the number of semiconductor devices. In the calculation of the power capability after a single phase fault, it is assumed that the healthy phase currents are accordingly shifted in order to preserve the armature MMF waveform [96,123]. However, no increase in the healthy current magnitudes or optimal current reshaping [23] to compensate for the fault is considered. Enough provision in the DC bus voltage is assumed so that the shifting of the healthy phase currents at rated speed and torque conditions is possible.

<table>
<thead>
<tr>
<th>Topology</th>
<th>Inverter cost</th>
<th>Machine cost</th>
<th>Semiconductor device count</th>
<th>Power after a single phase OC fault*</th>
</tr>
</thead>
<tbody>
<tr>
<td>Standard 3-phase PMSM drive</td>
<td>100%</td>
<td>100%</td>
<td>6</td>
<td>0%</td>
</tr>
<tr>
<td>Redundant machine and inverter (Fig. 2.4)</td>
<td>2 x 50%</td>
<td>2 x 50%</td>
<td>12</td>
<td>50%</td>
</tr>
<tr>
<td>Switch-redundant inverter (Fig. 2.7)</td>
<td>179%</td>
<td>100%</td>
<td>10</td>
<td>29%</td>
</tr>
<tr>
<td>Additional leg connected to neutral (Fig. 2.8)</td>
<td>183%</td>
<td>100%</td>
<td>11</td>
<td>33%</td>
</tr>
<tr>
<td>Additional redundant leg (Fig. 2.9)</td>
<td>233%</td>
<td>100%</td>
<td>14</td>
<td>100%</td>
</tr>
<tr>
<td>Five-phase H-bridge supplied</td>
<td>115%</td>
<td>100%</td>
<td>20</td>
<td>72%</td>
</tr>
<tr>
<td>Dual three-phase (Fig. 2.12b)</td>
<td>2 x 50%</td>
<td>100%</td>
<td>12</td>
<td>50%</td>
</tr>
</tbody>
</table>

* with no magnitude increase in the healthy phase currents

Table 2.2: Comparison among different fault-tolerant approaches

The first choice, the redundant machine and inverter configuration, allows the PMSM drive to operate with half power after a single phase fault just by switching off the faulted machine. By adding a clutch in-between the machines, mechanical faults can be tolerated in the rear machine as well, at the expense of additional weight and cost and a reduced reliability due to faults in the mechanical element. Even if the performance figures of this configuration and those of the dual three-phase machine look the same, the latter should be preferred due to the lower cost of having a single machine instead of two smaller ones. The switch-redundant inverter topology and the modified inverter structure with an additional leg connected to the neutral point impose the lowest penalties in terms of required number of elements and,
2.4. METHODS FOR GRANTING FAULT-TOLERANCE TO A PMSM DRIVE

hence, reduced reliability. However, the post-fault power capability of these configurations is low and the inverter cost increase substantial. A better post-fault operation with a lower inverter cost is achieved with the single phase modular configuration consisting of a five-phase PMSM supplied by a full H-bridge inverter. The number of required power semiconductor devices, though, is the highest one. It must be noted that the inverter cost associated to the single phase modular configuration shown in Table 2.2 does not consider the use of TRIACs placed in series with the machine phases. Thus, such a drive configuration cannot operate with switch short-circuit faults. If TRIACs are added to the configuration, the inverter cost becomes a 144% of that of the standard 3-phase VSI and the number of power semiconductor devices increases to 25. Finally, the modified inverter topology with a redundant inverter leg allows full power to be maintained after a switch open-circuit fault, but the power converter cost is highest. Furthermore, this topology cannot withstand a winding open-circuit fault on its own, requiring an additional switching element connected to the machine’s neutral point in order to allow some degree of controlability to remain after the fault.

From the results of the previous table, it can be concluded that the single phase modular and the dual 3-phase configurations are among the best choices in terms of overall system cost and fault-tolerant capability. Furthermore, among the discussed topologies, the five-phase modular design is the only one able to sustain multiple electrical faults. In the rest of configurations, a fault in a single machine or inverter element renders the PMSM drive to a state in which the fault-tolerant capability is no longer present. Multiple-stator designs can also tolerate more than one electrical fault in different modules when the number of 3-phase winding sets is higher than two. A 5-phase PMSM drive requires custom laminations and a non-standard power converter design. Thus, the overall system cost is higher than in the case of a multiple-stator design, but a better performance after a fault can be achieved [124].

It should be noted that the discussed methods are not rigid and that different fault-tolerant approaches can be combined. For instance, the authors in [16, 74] propose a motor drive system composed of two 3-phase motors mounted on the same shaft in which each machine is made fault-tolerant by adopting a single phase modular design. The proposed design employs redundant position sensors, independent controllers and a back-up controller for fault-detection and supervisory actions. According to [61], the use of dual-stator machine designs and redundant drive structures is common in ship propulsion applications. Hence, redundancy at both the system level and at the component level is achieved. Several authors have also considered multiple-stator designs with a reduced magnetic, physical and thermal coupling between the different 3-phase winding sets; features that are normally sought in single phase modular machine designs [79, 87, 117, 120, 125].

2.4.5 Fault-tolerance against winding short-circuit faults

Up to this point, only tolerance against winding open-circuit faults or faults in the power semiconductor switches has been discussed. In order for a PMSM drive to be able to sustain winding short-circuit faults, additional measures must be taken. One common approach is to design the PM machine with a high enough phase inductance so as to limit the short-circuit currents to a given threshold [66]. As a consequence, the power factor at healthy conditions is low. All the single phase modular designs found in the related literature, without exception, use this method. Examples of multiple-stator designs with a unity phase self-inductance can be readily found as well [79, 87, 117, 125]. In both cases, it must be ensured that the magnetic coupling between machine phases is reduced in order for the current in the faulted phase not to affect the flux linked by the remaining healthy phases [126]. The braking torque arising from a winding short-circuit fault is another aspect that must be taken into consideration; especially at low speeds [88].

In [64], several other methods that can be employed in order to limit the fault currents in short-circuit conditions are discussed. These include the use of:

- Memory motors.
- Machines with the magnets located in the stator and different kinds of actuators in order to reduce the PM flux linkage.
- Auxiliary windings.
CHAPTER 2. STATE OF THE ART IN FAULT-TOLERANT PMSM DRIVES

- Mechanical displacement/disengagement elements.
- Magnetic shields and shunts.
- Bi-state magnetic materials.
- Magnetic fluids.

These methods are aimed at weakening the PM magnetic field and, thus, to the limit they can be used to mitigate the short-circuit currents. Most of them require additional components, which, in turn, translates into additional complexity, weight, cost and reliability issues. How fast can the different methods detect a fault and trigger the corresponding fault mitigation mechanism is also in question [64]. In terms of simplicity, cost and reliability, the use of machine designs with a high enough phase self-inductance is the most promising approach and is, generally speaking, the preferred choice.

2.4.6 Research proposal

As it has been previously discussed, the characteristic of fault-tolerance is not an absolute and no system can be made truly tolerant to any possible combination of faults. The required degree of fault-tolerance and, hence, the design approach to be considered, depends on the particular application at hand. Since granting fault-tolerance to a system adds complexity and cost, only the most likely and/or severe faults should be considered.

The present research work has been oriented towards single phase modular PM machine designs. These drive systems are able to sustain multiple faults and allow to retain a good post-fault performance at a reasonable cost. In the present thesis, only machine winding and power converter related electrical faults have been considered. The incidence of other kind of faults, such as those in the DC bus capacitor and in the bearings, has been assumed to be largely preventable with adequate condition monitoring and preventive maintenance techniques. Failures in the power supply and in the sensors can be tackled by employing redundant configurations. Additionally, control strategies that allow a PMSM drive to operate with faulted sensors have been proposed in the related literature as well [45, 46].

In the following sections, the key aspects related to the use of fault-tolerant single phase modular PMSM drives are further addressed. First, the drive requirements and the derived penalties related to this method are reviewed. Then, the characteristics of the winding topologies commonly employed in these designs are studied and, finally, the properties of multiphase AC drive designs are investigated. With this review, the main aspects of the state of the art in fault-tolerant PMSM drives are covered.

2.5 Requirements for fault-tolerant drives with phase isolation

The requirements for fault-tolerant PMSM drive designs based on a maximum isolation between phases have been covered in [66, 67]. The main goal is to split the drive in multiple independent phases that can be regarded as single modules. In the following section, the main requirements for this method, together with the derived penalties in terms of drive’s size and, hence, cost are reviewed.

2.5.1 Maximum isolation between phases

In order for the PMSM drive phases to behave as independent modules, there must be the minimum possible magnetic, electrical and thermal interaction between them. Physical isolation between phases is also an important condition. These isolation requirements must be fulfilled both at the machine and at the power converter side [67].

Magnetic isolation

Magnetic coupling is the phenomenon by which the flux linked by one phase is affected by the current flowing through another one. Typically, it is described by the mutual inductance between phases, $L_{ij}$:
2.5. REQUIREMENTS FOR FAULT-TOLENT DRIVES WITH PHASE ISOLATION

That is, the flux linked by one phase $\lambda_i$ depends on the flux created by the magnets $\lambda_{i,m}$, the current flowing through the phase itself $L_{ii}i_i$ and the currents flowing through the rest of phases $\sum_{j \neq i}^m L_{ij}i_j$. The mutual inductance values between phases are a function of a number of factors, including the geometry of the machine, the number of conductors, the saturation of the magnetic circuit, etc. A reduced mutual inductance between phases is essential in fault-tolerant machines in order to prevent the fault of one phase to affect the remaining healthy phases; specially in short-circuit conditions. When the mutual inductance between phases is reduced, the flux produced by the healthy phase currents is not linked by the faulted phase and the short-circuit current becomes limited. Likewise, in case of fault, the current in the faulted phase does not affect the flux linked by the healthy phases, that can continue to operate normally [126].

Various techniques can be applied to reduce the mutual inductance between phases. The easiest one is to employ certain winding configurations that inherently lead to a negligible mutual coupling between phases [75]. In case of surface mounted PM machines, the air-gap component of the mutual inductance between phases can be greatly reduced by employing deep magnets together with a nonmagnetic retaining sleeve [67]. In order to reduce the slot leakage component of the mutual inductance, each slot must contain only coils belonging to a single phase.

Electrical isolation

In a standard half-bridge inverter supplied star-connected PMSM, a power switch or winding short-circuit fault may cause the neutral point of the star connection to rise to the DC link voltage, so that the ability to deliver a net torque is lost [66]. By employing a full H-bridge inverter, each machine phase becomes electrically independent and an additional degree of freedom for machine supply is gained; arising from the fact that the zero sequence current is not forced to be null anymore. This additional degree of freedom can be used to improve the drive’s performance in fault conditions [126]. Particularly, a higher output torque can be achieved for the same voltage limit. This fact is easiest understood in the case of a standard 3-phase PMSM drive. If the machine is supplied by a half-bridge inverter, the loss of one phase leads the machine to a state akin to that of a single-phase machine, making it unable to generate a rotating magnet field. On the contrary, if a full H-bridge inverter is considered, the remaining two healthy phases can be independently controlled and a rotating magnetic field and, hence, a net torque can be generated [20]. According to [67], although the use of a full H-bridge inverter doubles the number of power semiconductor switches, it only slightly increases the volt-amperes rating of the inverter, as each power switch needs to withstand only the phase voltage instead of the line voltage in a star-connected system.

Although having an electrical isolation between machine phases is highly desirable, this is not an absolute requirement for multiphase fault-tolerant drives. In [123], it has been demonstrated that a 5-phase machine supplied from a half-bridge inverter is capable of delivering a significant amount of torque with a low torque ripple, even after the loss of one phase. In the case of an open-circuit fault involving the loss of two adjacent phases, the torque capability is highly reduced due to the inability to control the zero-sequence current. Generally speaking, vast improvements in the post-fault operation of a PMSM drive can be obtained by supplying each machine phase independently.

Thermal isolation

A good thermal isolation between phases is required in order to prevent the propagation of a fault from the faulted phase to the remaining healthy phases. Inter-turn short-circuit faults are characterized by giving rise to elevated fault currents and to a substantial amount of heat to be released within the faulted winding. If no proper thermal isolation between phases exists, the locally generated heat will propagate to neighboring phases, jeopardizing the integrity of the healthy phase windings isolation. In order to attain an effective thermal isolation between phases, it is necessary that the phase windings are
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physically separated and that a good thermal path to the cooling surface (most usually the stator outer surface) is provided [67,126].

Physical isolation

In addition to providing for a good thermal isolation, having a physical isolation between phases helps prevent the occurrence of short-circuit faults involving the loss of two phases [66]. This is highly desirable in fault-tolerant drives, as the loss of two phases seriously impairs the drive’s torque capability.

2.5.2 Unity phase self-inductance

Besides the need for the various machine and inverter phases to be isolated and operate independently, a further requirement for fault-tolerant drives is that any fault can be thermally withstood over a long period of time. This implies that, in case of a single phase short-circuit fault, the fault current must be limited to, at least, the rated phase current value [66]. In case of a terminal short-circuit fault, the short-circuit current in a PMSM is mainly limited by the $d$-axis inductance; specially at high speeds, when the winding reactance largely dominates over the phase resistance. Therefore, by designing the machine with a 1 pu $d$-axis inductance, the steady-state short-circuit current can be effectively limited [67]. This requirement is usually achieved by designing the slot geometry so that a high slot leakage inductance is obtained [126]. Some phase, pole and slot winding combinations also help obtain high phase inductance values, as it will be discussed on a latter section.

2.5.3 Other requirements

Other common requirements to be met in fault-tolerant PMSM drive designs are that:

- The loss of one phase does not cause too high a torque ripple or excessive disturbances to the DC link current in the power converter [75]. This demands that multiphase drives, in which the number of phases is higher than three, are used. The adoption of multiphase systems leads to several other advantages, as it will latter discussed.

- In case the drive is required to deliver the full rated torque under fault conditions, the machine and the inverter must be overrated accordingly. The necessary overrating depends on the type of fault, the number of phases, the inverter topology (half-bridge or full H-bridge inverter), the adopted post-fault control strategy, etc. Ideally, a $m$-phase fault-tolerant drive should be overrated by a factor equal to $F = \frac{m}{(m-1)}$ in order to compensate for the loss of power under a single phase fault [66,67].

- The thermal and physical requirements stated for the electrical machine must be met in the power converter as well. Specifically, sufficient thermal and physical isolation between power semiconductor switches is required in order for a fault in one switch to not propagate to neighboring switches.

- The power converter must be equipped with the necessary means to quickly detect and respond to several fault conditions; including, winding open-circuit faults, short-circuit faults, power transistor faults, etc. This is necessary in order to prevent that a faulted phase draws an excessive current from the DC link or that significant braking torque is applied to the shaft [62]. In the related literature, several modified control strategies for PMSM drives to operate under fault conditions have been proposed. These remedial strategies are discussed in detail in a latter chapter.

- Related to the previous requirement, certain properties are demanded from the power semiconductor switches [62]. Firstly, the power semiconductor devices must be able to sustain a short-circuit fault for a sufficient time until the fault is detected and the proper remedial action is taken. Additionally, in the presence of a short-circuit fault, the control must be able to turn off a closed switch. It is also highly desirable that the used power semiconductor devices have appropriate gate drive circuits capable of providing information on the state of the switch.
2.5. REQUIREMENTS FOR FAULT-TOLERANT DRIVES WITH PHASE ISOLATION

2.5.4 Penalties for fault-tolerant drives with phase isolation

A number of penalties arise when a PM drive is designed to be fault-tolerant. These stem from the unusual drive design with a per unit \( d \)-axis inductance and the use of individual H-bridges to supply each phase independently. The authors in [62], establish a comparison between a fault-tolerant PMSM drive and a traditional 3-phase machine with a negligible phase self-inductance and supplied by a half-bridge inverter. The key points are described below.

It must be noted that the penalties arising from having a 1 pu phase inductance are not related to the method of designing the PMSM with the minimum possible coupling between phases. In this case, the drawbacks stem from the requirement that the drive must be able to thermally sustain a short-circuit fault over a long period of time. The same penalties would apply in case a split-phase machine is designed to endure a short-circuit fault.

Additional machine material stemming from the unity phase inductance

Having a high phase inductance results in an increased stator iron flux, if compared to a machine with negligible inductance. This demands that the stator yoke is increased by a factor of \( \sqrt{2} \). An additional volume of iron comes from the large tooth tips usually employed to increase the stator leakage inductance and achieve the per unit inductance requirement. For the particular design described in [62], the increase in stator iron mass due to the aforementioned requirement is about a 15%.

Additional power semiconductor switches

By using individual H-bridges to supply each phase separately, the number of required power switches doubles. This increases the cost and volume of the gate drive electronics proportionally with the number of switches, but only moderately increases the total cost of the power semiconductor devices, as it usually depends on the volt-ampere rating of the inverter [62]. The increase in the power device count arising from the use of a full H-bridge inverter may not be a serious penalty in high power applications, in which multilevel converters are commonly used due to the power limitations on single level converters. By adopting a modular design with multiple phases, the electric power is distributed among a higher number of phases, the power rating of each semiconductor device is reduced and single level converters may still be used [121,127]. In the end, the number of required power semiconductor switches will depend on the power requirements of each particular application.

Additional power converter material

As previously commented, the amount of power semiconductor material is proportional to the total volt-ampere rating of the inverter [62]. This is usually defined as the product of the number of switching devices, the DC link voltage and the peak current capability [122]. For a 3-phase system, the volt-ampere rating of a full H-bridge inverter is \( 2/\sqrt{3} \) times that of a half-bridge inverter, as the number of power switches doubles, but each switch must withstand only the phase voltage instead of the line voltage [62]. Moreover, when a full H-bridge inverter is used, the volt-ampere rating of the inverter does not increase with a growing number of phases.

When the high phase inductance comes into picture, the previous volt-ampere rating figure must be increased due to the large reactive power and the low power factor associated with the 1 pu inductance. A non-salient PMSM, in which no reluctance torque is available, will operate at a power factor close to \( 1/\sqrt{2} \). In this case, the previous volt-ampere rating must be increased by a factor of \( \sqrt{2} \), resulting in a \( 2\sqrt{2}/\sqrt{3} = 1.633 \) increase over the standard 3-phase half-bridge inverter [62]. In order to reduce the volt-ampere requirement for the drive, the current angle with respect to the back-EMF can be advanced by \( \psi = 30^\circ \). In a non-salient PMSM drive, this increases the current necessary to deliver the rated torque by a 15%, but reduces the required voltage by a 29% (if the effect of the phase resistance is neglected), leading to a total reduction in the inverter volt-ampere rating of 18% [62]. In this case, the required volt-ampere rating for a fault-tolerant 3-phase PMSM drive is just a 34% higher than that of a standard 3-phase half-bridge supplied PM machine.
Additional machine and inverter material stemming from redundancy

The last penalty related to having a fault-tolerant drive is related to the required degree of overrating in case the drive must deliver the rated torque at fault conditions. In case a \( m \)-phase drive must produce the rated power when one phase is lost, both the machine and inverter material must be increased by a factor equal to \( \frac{m}{m-1} \).

2.6 Fractional-slot concentrated-winding machines

Regarding the windings commonly employed in the stator of AC machines, a wide variety of possibilities exist. Most usually the winding types are classified depending on the number of slots per pole and per phase, \( q = Q/(2pm) \). Integral-slot windings are those in which \( q \) is an integer number, while fractional windings usually refer to those windings for which \( q \) is a fractional, higher than unity number. Traditionally, integral-slot windings with \( q > 1 \) have been termed distributed-windings, whereas windings having one slot per pole per phase have been named concentrated-windings \((q = 1)\). Fractional-slot Concentrated-Windings (FSCW) refer to a subclass of AC windings for which \( q \) is a fractional, less than unity number. An alternative classification can be established among overlapping and non-overlapping windings, depending on whether the coil ends in the machine overlap or not. Machines with \( q \geq 1 \) lead to the end-windings to overlap, while the coils of FSCW machines are concentrated around and single tooth and any overlap with other phase end-windings is eliminated. Owing to this reason, FSCWs are sometimes termed concentrated non-overlapping windings or tooth-coil windings. Finally, the winding types can be further classified according to the number of layers \( n_l \), for which single-layer and double-layer windings are the most common choice. For instance, Figure 2.13 shows different kinds of windings for a 3-phase four-pole machine.

The first three windings in Figure 2.13 have \( q \geq 1 \) and, thus, lead to an overlapping of the end-windings of the different phases. As it can be appreciated, in the fractional-slot concentrated-winding the coils are wound around a single tooth and the overlapping between coils is eliminated. Depending on the coil configuration, two main classes of FSCWs are distinguished: single-layer windings, in which only alternate teeth are wound, and double-layer windings, in which every teeth carries a coil. The difference between both configurations is depicted in Figure 2.14.

Over the last decade, FSCW machines have experimented a huge development [128]. Specifically, they have proven to be specially well suited for fault-tolerant applications [66, 69]. The particular traits of FSCW PM machines have been extensively discussed in [126,129–132], for instance.

In the following section, why FSCWs are the natural choice for fault-tolerant applications seeking the maximum isolation between phases is explained and the main characteristics of FSCW machines are reviewed.

2.6.1 Use of FSCWs in fault-tolerant machines

The requirements for fault-tolerant PMSM drive designs based on a maximum isolation between phases have been discussed in the previous section. These include a unity \( d \)-axis inductance and the lowest possible coupling between machine phases [66,67]. These requisites are naturally met in FSCW machines as discussed below.

Magnetic isolation

Fractional-slot concentrated-winding machines exhibit an intrinsically low mutual inductance between phases compared to distributed-winding machines [126]. It has been shown in [75, 133], for instance, that some phase, pole and slot combinations lead to a negligible mutual inductance between phases. Furthermore, when single-layer FSCWs are considered, each slot carries only conductors belonging to a single phase and the slot leakage component of the mutual inductance between phases is theoretically
2.6. FRACTIONAL-SLOT CONCENTRATED-WINDING MACHINES

(a) Single-layer integral-slot distributed-winding ($Q = 24$, $p = 2$, $q = 2$)

(b) Single-layer integral-slot concentrated-winding ($Q = 12$, $p = 2$, $q = 1$)

(c) Double-layer fractional winding ($Q = 18$, $p = 2$, $q = \frac{3}{2}$)

(d) Double-layer fractional-slot concentrated-winding ($Q = 6$, $p = 2$, $q = \frac{1}{2}$)

Figure 2.13: Different four-pole stator winding possibilities

(a) Single-layer winding ($n_l = 1$)

(b) Double-layer winding ($n_l = 2$)

Figure 2.14: FSCW configurations ($Q = 6$, $p = 2$, $q = \frac{1}{2}$)

eliminated [62]. In this case, only some magnetic coupling through the air-gap is present, that can be highly reduced by employing thick surface-mounted magnets [67].
Physical isolation

As the coils in FSCWs are wound around a single tooth, no overlapping between phase coils exists and the contact between machine phases is reduced. This reduces the chances for a short-circuit fault between distinct phases to occur [126]. In single-layer or alternate teeth wound machines, the physical isolation between phase windings is maximized by the additional teeth between coils. This arrangement makes phase to phase short-circuit faults specially unlikely [67]. In double-layer winding machines, a separator between coil sides may be required in order to guarantee the isolation between phase windings [132].

Thermal isolation

The physical isolation between coils leads to a good thermal isolation between phases. In case a single-layer FSCW is employed, this isolation is maximized and, as long as a good thermal path to the outer surface of the stator is provided, a fault in one phase will have a minimum impact on the operation of the rest of phases [67].

Unity phase self-inductance

The main components of the self-inductance in a PMSM are related to the air-gap flux and the leakage flux across the slot [134, 135]. Fractional-slot concentrated-windings inherently lead to high slot leakage inductances and high air-gap harmonic leakage components due to the rich harmonic content of the armature MMF distribution [126, 131]. In case of single-layer FSCWs, the increased MMF harmonic content compared to double-layer windings and the higher number of conductors belonging to the same phase in each slot leads to the highest phase self-inductance values.

Modularity

A further advantage of FSCW machines for fault-tolerant designs is that the winding layout, with coils wound around a single tooth, leads to highly modular structures [132]. Modular designs allow individual segments to be separately assembled and removed, making the repair of a faulted stator winding easier.

2.6.2 Design of fractional-slot concentrated-windings

Since in a FSCW the number of slots per pole and per phase is not restricted to be an integer number, there are many possible combinations of phases, poles and slots that lead to a winding of this kind; both single-layer and double-layer. Different methods exist that allow to design a FSCW. For instance, [136] describes an algorithm to design 3-phase FSCWs that is expanded in [137] to cater for four, five and six phase machines. In the latter case, reduced forms of the winding systems are considered ($\alpha_m = \pi/m$, see Appendix A). An equivalent method, based on a graphical representation of the EMF phasors, is the so called method of the “star of slots”. The theory behind this method, together with the design of FSCWs with an odd number of phases, is thoroughly explained in [126, 133, 138]. The star of slots allows to analyze the harmonic content of both the back-EMF waveform and the air-gap armature MMF distribution by a graphical representation of the phasors corresponding to each harmonic. Thus, it is useful in the computation of the winding factors for the different harmonics. This is the method used in the present thesis.

2.6.3 Comparison between FSCW and overlapping winding machines

In addition to significant advantages regarding the fault-tolerant behavior of PM machines, the adoption of FSCWs leads to several benefits over the use of the more traditional overlapping windings. In particular, the most noteworthy advantages of employing a FSCW are:
2.6. FRACTIONAL-SLOT CONCENTRATED-WINDING MACHINES

- Shorter end-windings and, thus, lower machine length and lower cost associated to the end-regions [130].
- Higher slot fill factor, if coupled with modular designs [139]. In traditional machines making use of overlapping windings, the individual coils are usually manually inserted into the slots through the slot openings, which leads to an aleatory distribution of the conductors inside the slot. By using single tooth segments or soft magnetic composites together with pre-wound windings, quite high slot fill factors can be achieved [140]. In particular, [139] cites a fill factor of 78% for a FSCW machine with powdered iron cores and prepressed windings. If joint-lapped cores and an automated winding process is used, high slot fill factors can be obtained as well [140]. In this last reference, the different methods available to manufacture FSCW machines are thoroughly discussed.
- Reduced copper losses, as a consequence of the shorter end-windings and the higher copper fill factor [141]. This may involve an increase in efficiency for designs in which stator copper losses are the main source of machine losses.
- Higher power density in machines with a reduction in copper losses. In [11], various examples on how the use of FSCWs can lead to a reduction in the operating temperature of the machine and, hence, to an increase in the power density are shown. Additionally, FSCWs are most commonly employed in high pole number machine designs, for which the multi-pole magnetic circuit with small stator yokes can lead to a significant reduction in the machine’s weight and volume [142]. The high slot fill factor and the small stator yoke also provide for a good thermal path between the coils and the machine’s outer surface, which can result in a reduction in winding temperature and, in turn, to an increase in power capability.
- Easier manufacture and reduced manufacturing costs in modular designs with an automated winding process [129,133].
- Reduced cogging torque and torque ripple if suitable winding combinations are chosen [143].
- Inherently large flux-weakening range due to the high slot leakage inductance. In [144, 145], it has been shown that wide constant-power operation ranges can be achieved for surface-mounted PM machines by employing FSCWs.
- Enhanced voltage quality in PMSM designs. The generated back-EMF is inherently sinusoidal in FSCW machines, despite the existence of higher harmonics in the air-gap MMF distribution created by the PM [129]. This leads to a smoother torque, an improved output voltage quality when operated as generator and an enhanced DC link voltage [146].

Despite the several advantages, the use of FSCWs has a number of drawbacks as well, such as:

- Higher rotor losses due to an increase in the harmonic content of the armature air-gap MMF waveform [130]. The increase on the MMF subharmonic content of certain FSCWs designs is particularly detrimental regarding rotor losses [147]. These may be partially compensated by laminating the rotor core and by segmenting the magnets either axially [148] or circumferentially [149].
- Unbalanced radial forces and alternating magnetic fields if the greatest common divisor between the number of poles and slots is low [129]. This, in turn, leads to higher stresses on the bearings and increased vibration and acoustic noise [150].
- Slightly lower main harmonic winding factors than integral-slot windings [126]. This penalty can be negligible if certain pole and slot combinations are chosen [131].

2.6.4 Characteristics according to the number of layers

As mentioned earlier, most commonly FSCW machines employ either a single-layer or a double-layer winding, with 1-layer windings having coils wound only around alternate teeth and 2-layer windings having coils placed around each tooth. Double-layer windings admit a greater number of pole and slot combinations to be chosen. The constrains and rules that allow a 2-layer winding to be transformed into a 1-layer winding are explained, for instance, in [133].
CHAPTER 2. STATE OF THE ART IN FAULT-TOLERANT PMSM DRIVES

The choice on the number of layers for a certain phase, pole and slot combination has an enormous impact on several characteristics of the machine. The main distinctions between single-layer and double-layer FSCW machines with the same phase, pole and slot combination are as follows [131]:

- Single-layer windings lead to higher winding factors for the main harmonic.
- The end-windings are shorter in double-layer FSCW machines.
- The slot fill factor in double-layer winding machines is lower due to the additional insulation required between distinct phase coils.
- Single-layer FSCWs lead to higher phase self-inductance and lower mutual inductance values. Additionally, the physical and thermal isolation between phase windings is higher in this type of machines, resulting in an improved fault-tolerant capability.
- The induced back-EMF waveform tends to be more trapezoidal in single-layer machines and more sinusoidal in double-layer winding machines.
- Single-layer FSCWs give rise to a higher air-gap MMF harmonic and subharmonic content. In particular, 1-layer winding combinations with more poles than slots lead to the highest MMF subharmonic content [137]. Armature MMF harmonics and subharmonic can be detrimental for the operation of the machine as they cause rotor losses, local rotor saturation, unbalanced radial forces, increased torque ripple, vibrations and noise, etc. [126].
- Manufacturing 1-layer windings is simpler and the process can be easier automated. In modular machine designs having identical coil groups, the coils are more isolated in the case a single-layer FSCW is used. Hence, the assembly and disassembly of each machine segment is easier and a better insulation can be achieved [132].

In addition to 1-layer and 2-layer windings, it is possible to design FSCWs with a higher number of layers. Specifically, 3-layer and 4-layer designs have been proposed in [151–153]. Adopting a higher number of layers helps reduce the harmonic content of the armature MMF and, in turn, the torque ripple and the rotor losses of the machine [132, 154]. It is particularly easy to obtain a 4-layer winding from a 2-layer one. The procedure in this case consists on dividing the stator coils in two sets and shifting one set with respect to the other. As a result, the winding factor for the main harmonic is reduced, but the armature MMF distribution has a lower harmonic content; specially regarding MMF subharmonics [132]. Although the number of layers can be increased [153], the improvements to be obtained in MMF harmonic reduction are quite unlikely to compensate for the increase in winding costs. The properties of 4-layer FSCWs are discussed in greater detail in the next chapter.

2.6.5 Non-conventional FSCW machines structures

The single-layer and double-layer machine configurations exemplified in Figure 2.14, with regular teeth and uniformly distributed coils spanning a single slot, are by far the most commonly used FSCW machine designs. Nevertheless, other FSCW machine structures have been proposed in the related literature in order to achieve certain objectives. Some of these less conventional configurations are discussed below, although their study is beyond the scope of the present thesis.

Mixed single/double-layer windings

In order to achieve a high degree of modularity, single-layer windings may be preferred to double-layer FSCWs, as they lead to machine segments separated by half-tooth boundaries. However, as earlier mentioned, 1-layer FSCWs lead to higher rotor losses and a compromise solution between both winding configurations may be required. Starting from a double-layer winding, by removing a coil per coil group, a higher degree of isolation between phase windings and an improved modularity can be achieved [132]. An example of a FSCW machine with a mixed winding is show in Figure 2.15a.
2.6. FRACTIONAL-SLOT CONCENTRATED-WINDING MACHINES

Machines with an irregular teeth distribution

Stator structures with unequal tooth widths have been used together with single-layer FSCWs in [136, 155–157], for instance. In the cited machine designs, the non-wound teeth are made thinner than the teeth carrying coils, that have tooth tips spanning almost a pole-pitch. By employing these structures with similar pole and slot pitches, the winding flux linkage and the fundamental winding factor are increased, the induced back-EMF becomes more trapezoidal and the torque density can be improved [156]. The phase self-inductance slightly increases over a machine with equal tooth widths and the mutual inductance between phases is marginally reduced [157]. An improvement in terms of an increased copper area is suggested in [136] as well. Owing to the trapezoidal induced voltage waveform, these structures are better suited for BLDC operation or for PMSM designs in which a significant back-EMF third harmonic may be advantageous [158]. In BLDC operation, machines with an irregular teeth distribution exhibit lower torque ripple than machines with equal tooth widths, whereas the situation reverses for synchronous operation [155].

The benefits gained from having unequal tooth structures gradually decrease as the number of poles increases [156]. The higher the number of poles, the higher the inter-pole leakage flux and, as a consequence, the air-gap flux density distribution and the back-EMF waveform become more sinusoidal. Ultimately, for high pole number machines, there is almost no difference in having equal or unequal tooth widths [159].

A different context, in which irregular teeth arrangements are adopted, is when FSCWs with open slots are used [160,161]. In this case, by having parallel stator slots and rectangular coil carrying teeth, pre-wound coils can be used, leading to an easy stator assembly and reduced manufacturing costs. Additionally, unequal tooth and slot widths should be used in the case of mixed single/double-layer FSCWs designs in order to optimize the amount of copper in each slot and increase the winding factor.

Concentrated windings spanning two slot pitches

As earlier commented, expanding a 2-layer FSCW into a 4-layer winding leads to a reduction in the harmonic content of the armature MMF waveform; specially regarding MMF subharmonics [132]. In order to reduced both the subharmonics and the higher order harmonics simultaneously, the use of multilayer windings with coils spanning two slot pitches ($y_q = 2$) is proposed in [162], for instance. This way, lower rotor losses and a reduced torque ripple compared to a conventional double-layer FSCW machine can be obtained. However, these structures lead to the slots not being fully utilized and to an increase in the end-winding length. An example of a FSCW machine making use of a 4-layer winding and coils spanning two slot pitches is show in Figure 2.15b.

![Figure 2.15: Non-conventional FSCW machines structure examples](image-url)
2.6.6 FSCW application examples

In addition to being extensively investigated, FSCW PMSM drives are already being considered in a wide variety of commercial applications. Below some examples are listed [5,128,130]:

- Aerospace: electrical actuators.
- EV and HEVs: traction motor (Honda, Toyota), starter/alternator (Toyota, ZF Sachs), in-wheel motors.
- Ship propulsion.
- Power generation: wind power, ocean wave power.
- Household appliances: washing machines (Whirlpool).
- HVAC: pump in air conditioning system (Panasonic).

2.7 Multiphase machines

In addition to the requirements of a maximum isolation between phases and a unity $d$-axis inductance in order to limit the short-circuit currents, an additional common demand for fault-tolerant drives is that the drive retains a sufficient level of performance even after a fault. This can be accomplished by adopting multiphase systems, in which the number of phases is higher than three. The higher number of phases implies a higher number of degrees of freedom for machine supply. These additional degrees of freedom over a standard 3-phase drive can be used to achieve a better performance, both at healthy and fault conditions [121]. Additionally, multiphase systems provide several advantages, as extensively covered in [27,121,127].

In the following subsections, the advantages and disadvantages arising from adopting multiphase machine drives are briefly discussed.

2.7.1 Advantages of using multiphase machines

Improved fault-tolerance

Multiphase machine drives inherently possess an improved fault-tolerant capability over a standard 3-phase machine due to the high level of redundancy. In a standard half-bridge supplied 3-phase machine, the loss of one phase renders the machine to a state similar to that of a single phase drive, which cannot start on its own and in which high torque pulsations result [27]. On the other hand, a multiphase drive can deliver a significant amount of torque at fault conditions, even if the supply from independent H-bridge inverters is not considered, as demonstrated by [123]. When the machine phases are independently supplied, a rotating armature main MMF harmonic and, hence, a net torque capability can be maintained with the loss of $m-2$ phases. If a single phase is lost and the machine phases are independently supplied, the drive can ideally deliver $\frac{m-1}{m}$ of the rated torque with no drive overrating. In practice, the torque that can be delivered depends on a number of factors, such as the type of fault, the adopted control strategy, the available DC bus voltage, etc.

Reduced phase current

In a multiphase AC drive, since the power is evenly divided among a larger number of phases, the phase current can be reduced without increasing the voltage per phase. This implies a reduced size and power rating of the individual power semiconductor switches [121,127]. In high power applications, such as in marine propulsion, this can be an advantage due to size limitations on the individual converters [163]. Moreover, in high power applications, it is common to find multilevel converters due to the power handling limits of individual inverters. An alternative to this is to employ multiphase systems, which lead to a simpler and more-reliable power conditioning equipment since the need for parallel and/or series
connection of semiconductor switches is reduced or even eliminated [27, 164]. An additional degree of reliability is gained by the drop in DC current harmonics by the reduction in phase current level [121].

**Improved armature MMF waveform**

Generally speaking, higher phase number windings produce armature magnetic fields with a lower harmonic content due to the higher choice of phase current phasors [27, 163]. In the case of integral-slot distributed windings, the orders of the lowest spatial MMF harmonics are $2m \pm 1$ for $m$-phase machines [127]. In the case of FSCWs, as the rich MMF harmonic and subharmonic content is a function of the particular winding combination, the effect of increasing the number of phases is more difficult to assess. Anyhow, the previous property holds in general terms.

The reduction in the armature air-gap MMF distribution harmonic content leads to a number of benefits, such as a [27, 127, 163, 165]:

- Reduced amplitude and higher frequency of the torque pulsations. The torque ripple can be low even for machines with a non-sinusoidal back-EMF waveform [166]. In $m$-phase machine drives, the torque ripple frequency is at least $2m$ times the fundamental frequency of the drive [127].
- Reduced acoustic noise.
- Lower rotor losses and improved efficiency.

**Increased main winding factor**

As the windings of a multiphase machine produce a more sinusoidal MMF waveform, the winding factor for the main harmonic is higher. This, in turn, leads to an increase of the torque per ampere ratio for the machine.

**Stator current harmonic injection**

The additional degrees of freedom for machine supply arising from the increased number of phases can be used to inject higher order current harmonics. This, in turn, allows to design trapezoidal or quasi-rectangular back-EMF waveform machines that combine the power density of a BLDC machine with the low torque ripple and the controllability of a PMSM [158]. According to [127], torque enhancement by injection of higher order current harmonics is only possible for non-sinusoidal winding distributions; namely integral-slot concentrated windings and FSCWs.

**Multi-motor supply from a single inverter**

The additional degrees of freedom derived from the higher number of phases allow various motors to be supplied by a single voltage source inverter (VSI) [167]. In particular, a $m$-phase current controlled VSI can supply $(m - 1)/2$ series connected stator windings, achieving an independent control of the speed and the torque of the various machines. This can, of course, lead to several advantages, such as a reduction in the number of power semiconductor switches and, hence, a lower inverter cost, better reliability of the system, lower switching losses, etc.; the possibility of controlling the multi-motor drive system from a single DSP unit and the possibility of directly utilizing the braking energy of some of the machines of the group by the machines that are operating in the motoring mode [167]. The major drawbacks of multi-motor drive systems is that there is an increase in the overall stator copper losses and that the machines must have a sinusoidal winding distribution; namely, a distributed winding [127].

**Higher number of voltage space vectors**

A multiphase system offers a higher number of voltage space vectors, which can be an interesting feature in Direct Torque Control (DTC) applications, where the direct control of the stator flux and torque is
sought. The higher choice of voltage space vectors offers an enhanced flexibility in selecting the optimal switching states and a finer adjustment of the flux and the torque can be obtained [121].

**Improved torque density**

The increase in the torque per ampere ratio stemming from the higher main winding factor, the reduction in rotor losses derived by the lower MMF harmonic content and the possibility of enhancing the torque by injecting higher order current harmonics can lead to an increase in the torque density of a multiphase machine over a same size 3-phase machine [121].

### 2.7.2 Disadvantages of using multiphase machines

**Higher number of power semiconductor switches**

Having a higher number of phases involves an increased number of power semiconductor switches in the drive. This, in turn, can imply a higher volume, weight and cost of the power converter. The higher device count leads to a higher complexity and possibly to a lower reliability of the overall system as well [69].

**Increased difficulty to achieve a sinusoidal winding distribution**

As the number of phases increases, the number of slots for a certain pole number integral-slot distributed winding machine must increase as well. Hence, machine designs with small diameters may face difficulties to realize near-sinusoidal MMF distributions [127]. For instance, a 3-phase 8-pole integral-slot distributed winding machine requires at least 24 slots, while a 6-phase machine with the same characteristics demands the number of slots to be doubled.

**Losses in the stator end-parts: clamping rings, frame and end-shields**

According to [168], high phase number FSCWs may lead to higher losses in the support structure of the machine than FSCWs with a lower number of phases. Although technically sound, the analysis in [168] just compares 6 candidate machines with different phase, pole and slot number combinations. No extensive simulation of FSCWs with any possible winding combination is conducted. Moreover, for the selected combinations, the loss computations are conducted for different fundamental frequencies. Since the armature MMF harmonic content of a PMSM largely depends on the particular phase, pole and slot combination, it is difficult to evaluate the influence of a single winding parameter, be it either the number of phases or the pole number, on another variable with just a few computations. Therefore, the claim by [168] should be more carefully checked before given an overall validity.

### 2.7.3 Examples of fault-tolerant multiphase PMSM drives

The benefits derived from increasing the number of phases largely overcome the drawbacks associated to multiphase machine drives in high power and safety critical applications. In the related literature, numerous examples of fault-tolerant multiphase PMSM drives can be found; of which 5-phase machine drives are predominant. In Table 2.3, some examples of fault-tolerant multiphase PM machine drives designed to have the maximum possible isolation between phases are listed.

According to [24], a system with more than 5 phases will be generally undesirable due to the increased complexity, cost and fault rate of the individual components.
2.8 Conclusions

This chapter has covered the main aspects of the state of the art in fault-tolerant PMSM drives. A general overview on PMSM drive applications and on the main failure mechanisms of these systems has been given. Different strategies to grant fault-tolerance to a PMSM drive have been discussed and compared in terms of cost, reliability and post-fault performance. The topology consisting of multiple independent phases with the maximum isolation between the distinct phases has emerged as one of the most promising candidates in terms of overall system cost and fault-tolerant capability. Therefore, it has been decided to orient the present research towards these fault-tolerant drive designs. The requirements and the penalties arising from this method have been reviewed and the main characteristics of FSCW machines and multiphase systems have been covered, as well.
Chapter 3

Winding Selection Criteria for Fault-Tolerant PMSMs

As stated in the state of the art given in the previous chapter, fractional-slot concentrated-winding (FSCW) machines have been proposed as the most advantageous solution for meeting the isolation requirements demanded by fault-tolerant applications. There are many combinations of phases, poles and slots that lead to a FSCW, either with 1, 2 or 4 layers. In the present chapter, criteria for choosing the most appropriate number of phases, poles, slots and layers for a fault-tolerant design are investigated. First, the winding choice criteria commonly found in the related literature are revised. Then, own selection criteria based on an analytical calculation of the machine inductances are presented.

3.1 Traditional criteria

The choice of the number of phases, poles and slots heavily influences the performance of a FSCW machine. The usual criteria when assessing the different possible combinations focus on the following aspects [131,132,138,177]:

1. Winding feasibility.
2. Non-overlapped coil windings ($y_q = 1$).
4. Winding factor.
5. Radial forces.
6. Cogging torque and torque ripple.
7. Rotor losses.
8. MMF subharmonics.
9. Peak armature reaction.
10. Self-inductance.
11. Mutual inductance between phases.
12. Number of phases.

These twelve points are thoroughly discussed in the following subsections. In addition to the criteria stated above, there will be an upper limit for the allowable number of pole pairs due to geometrical constraints, excessive losses in the machine and maximum switching frequency of the inverter. Likewise, the number of slots will be limited by the diameter of the machine, since too small a slot-pitch leads to problems when winding the machine and very weak teeth.
3.1.1 Winding feasibility

For a \( m \)-phase number machine with \( 2p \) poles and \( Q \) slots, the main constraint for a symmetrical FSCW to be feasible is that [132]:

\[
\frac{Q}{mt} \text{ is an integer}
\]  

(3.1)

where \( t = \gcd(p, Q) \) is the electrical periodicity of the machine. In this case, the coil throw or coil span (expressed in number of slots) is \( y_q = \text{round}(Q/2p) \). In general terms, the previous restriction allows to design odd phase number 2-layer FSCWs \( (n_l = 2) \). If the winding satisfies the following properties:

1. The number of slots is an even number.
2. The coil throw \( y_q \) is an odd number.

the 2-layer winding can be reduced to a 1-layer winding \( (n_l = 1) \) [126].

Equation (3.1) applies to odd phase number windings. When the number of phases is even, there are combinations that do not satisfy (3.1), but still give rise to a FSCW; such as the example presented in [62, 66]. Figure 3.1 represents a 6-phase, 8-pole, 12-slot machine. For this combination, \( t = 4 \) and \( Q/(mt) = 1/2 \).

Figure 3.1: 6-phase FSCW machine with fractional \( Q/(mt) \)

Expanding a 2-layer winding into a 4-layer winding is always feasible, although not always desirable. This matter will be analyzed further in later subsections.

3.1.2 Non-overlapped coil windings

When the coil throw equals one slot, \( y_q = 1 \), the coil windings do not overlap and the end-winding length is kept at its minimum. In most cases, this reduction of the copper volume with respect to the same number of poles integral-slot machine implies a reduction of the copper losses for a given torque [138]. Another advantage of having non-overlapped coil windings is that the process of automatically winding the stator coils is greatly simplified [132].

Therefore, it is recommended that \( 2p \approx Q \).

3.1.3 Modularity

A machine is said to be modular when the stator can be split in identical parts where all the coils belong to the same phase. As an example, Figure 3.2 shows a machine with separate teeth structures [139]. In its most simple definition, any machine with non-overlapped coils can be considered to be modular.

Modular machines allow the groups of coils to be separately assembled and removed, making it easier to disassemble and replace faulted components and simplifying the connections in the machine [132].
3.1. TRADITIONAL CRITERIA

An additional advantage of modular designs is that the slot fill factor can be increased, leading to a
decrease of the copper losses and thus, an increase of the torque density [139]. These designs are specially
interesting for large machines due to their easy maintenance and due to the fact that different parts can
be simultaneously assembled [132].

In general terms, the coils belonging to the same phase in a modular design are arranged in $N_{grp}$ groups.
When the number of phases is odd, $N_{grp} = |Q - 2p|$ [132]. For some phase, pole and slot combinations,
all the coil groups have the same number of coils, making the stator assembly even easier. This happens
when:

$$N_{cpg} = \frac{mQ}{2mN_{grp}} \text{ is an integer}$$  \hspace{1cm} (3.2)

being $N_{cpg}$ the number of coils per group. Equation (3.2) allows to chose combinations that lead to coil
groups with the same number of coils. For combinations that do not fulfill (3.2), it is possible to modify
the disposition of the coils in order to have identical coil groups at the expense of reducing the winding
factor for the main harmonic. For an example of change of connections, refer to [132].

### 3.1.4 Winding factor

The winding factor for a certain harmonic, $k_{w,\nu}$, is a measure of the electrical displacement between the
phasors of the same phase [138]. It can be graphically computed from the star of slots and it is always
$k_{w,\nu} \leq 1$. The effect of the slot/pole combinations in the winding factor for the main harmonic have been
extensively reported. For instance, [141] investigates the winding factors of 3-phase 1-layer and 2-layer
machines and [178] studies the main winding factor for high pole number machines.

It is desirable that the winding factor for the main harmonic is as high as possible in order to maximize
the machine’s torque [137]. High main winding factors are obtained when the number of poles is close to
the number of slots, $2p \approx Q$, since in those cases the pole-pitch almost equals the slot-pitch and the flux
linkage is maximized [131]. For instance, Figure 3.3 shows the main winding factor for 3-phase machines.

The evolution of the main winding factor is periodical with the number of poles per slot, $2p/Q$; being
the period equal to 2. Therefore, high winding factors can be obtained with high $2p/Q$ combinations (or
low number of slots per pole and phase, $q = Q/(2pm)$). However, the large number of poles of these
combinations limits their possible uses to, perhaps, magnetic gear applications [131].

Figure 3.3 shows how 1-layer windings can lead to higher main winding factors than their corresponding 2-
layer windings. This happens when $Q/(2t)$ is even [133]. Four-layer windings exhibit lower main winding
factors than 2-layer FSCWs, but higher order EMF harmonics are reduced as well. Since the main
reason to adopt a 4-layer FSCW is to reduce the harmonic content of the armature MMF distribution,
only combinations that do not lead to the same amplitude decrease for all the space harmonics are
convenient [154]. Figure 3.4 represents the winding factor for different harmonics for a 3-phase, 10-pole, 12-slot machine with 2 and 4-layers.

Finally, the main winding factor increases with the number of phases. For instance, when \(2p = 26\) and \(Q = 30\), it is possible to design both 3-phase and 5-phase 1-layer machines. The winding factor for the main harmonic is \(k_{w,p} = 0.9358\) for the 3-phase machine and \(k_{w,p} = 0.9639\) for the 5-phase one.

### 3.1.5 Radial forces

For any electrical machine, radial forces between the stator and the rotor exist. These radial forces are mainly attractive in nature and come from the magnet, coil current and steel interactions [137]. If the forces are not regularly distributed along the air-gap, an unbalance magnetic pulling force is created [131]. This unidirectional pulling force rotates with time and causes vibrations within the machine, generates noise and stresses the bearings [129, 178].

As long as the unbalance magnetic pull is considered, machines with a force asymmetry can be designed [179]. However it is much simpler to select pole and slot combinations that have periodicities in the winding layout so that opposing forces in the air-gap are compensated and no net radial force is created [138]. The higher the number of periodicities, the lower the radial forces at each side of the air-gap [137]. If the number of periodicities is an even number, diametrically opposite forces compensate and therefore, the machine becomes more tolerant to eccentricities.

At no load, in order for the radial forces to be regularly distributed, it suffices that the greatest common divisor of the number of poles and slots is greater than 1, \(\text{gcd}(2p, Q) > 1\). However, at load conditions, there is an additional contribution to the radial forces resulting from the interaction of the harmonics of the permanent magnet flux and the armature field [132]. In this case, it is necessary to examine the periodicities in the winding layout of the machine.
3.1. TRADITIONAL CRITERIA

Figure 3.5 represents the radial magnetic force distribution in the air-gap for two single-layer 3-phase FSCW PMSM machines. The first machine has $2p = 8$, $Q = 9$, $gcd(2p, Q) = 1$ and no periodicities in the winding sequence. Therefore, a net radial force appears between the rotor and the stator. On the contrary, the second machine has $2p = 10$, $Q = 12$, $gcd(2p, Q) = 2$ and an antisymmetric distribution of the coils in the slots. In this case, the radial forces compensate and no unbalanced magnetic pull is generated.

![Figure 3.5: Radial forces and net radial force for two 3-phase FSCW machines](image)

(a) No winding periodicities  
(b) Winding periodicity = 2

In [180] 3-phase FSCW combinations with $2p = Q \pm 2$, $2p = Q \pm 1$ and $q = 0.5$ are studied. According to the investigation, the dominant order of the radial forces is at least 2 when $2p = Q \pm 2$ and a double-layer winding is considered or, if $2p = Q \pm 2$ and a single-layer is used, when $Q/t$ is even. As reported by [181], undesired noise can also appear in machines that do not have an unbalance magnetic pull. For more on magnetic forces and noise in FSCW machines, see [132,150,179,181,182].

To sum up, an adequate rule to design FSCW machines with respect to the magnetic forces is that the number of winding periodicities is an even number and sufficiently high.

3.1.6 Cogging torque and torque ripple

Many applications require a smooth, ripple free torque in order to avoid vibrations in the machine and reduce the acoustic noise [143]. Having a low torque ripple is also important to improve speed control accuracy.

In a permanent magnet machine, one of the main contributors to the torque ripple is the cogging torque due to the interaction between the magnets and the stator slotting; the other being the air-gap MMF distribution harmonics. Many researches have addressed the influence of the number of poles and slots in the cogging torque waveform in a PM machine [143,183–185]. The number of cogging torque periods per slot-pitch is:

$$N_p = \frac{2p}{gcd(2p, Q)} = \frac{lcm(2p, Q)}{Q}$$

Having a high value of $N_p$ leads to a high frequency and low amplitude cogging torque waveform [63]. High frequency torque harmonics are more easily filtered by the load inertia and thus, having a reduced number of pole and slot periodicities or high $lcm(2p, Q)$ values helps achieve a smoother torque [137]. Pole and slot combinations with $2p = Q+k$, $k \in \mathbb{N}$ lead to a higher $N_p$ than combinations with $2p = Q-k$; despite both combinations having the same main winding factor [131].

The requirement of having a low number of periodicities to reduce the cogging torque is detrimental to reduce the radial forces in the air-gap, so a compromise between both criteria must be met.

It must be noted that having a low cogging torque does not always guarantee a low torque ripple [186]. For instance, the high stator magnetic potential and the higher MMF harmonic content of some 1-layer
CHAPTER 3. WINDING SELECTION CRITERIA FOR FAULT-TOLERANT PMSMS

FSCW designs with respect to their 2-layer counterparts, can lead to high PM stresses and iron saturation under high loads, causing an increased torque ripple [126, 138].

3.1.7 Rotor losses

Rotor losses are a major issue in FSCW machines [67]. They greatly affect the efficiency of the machine [187] and may be high enough so as to compromise machine operation [152].

Rotor losses in PM machines are caused by the existence of flux pulsations due to the stator slots and by the presence of air-gap MMF harmonics. When the machine is designed with open slots and 1-layer FSCWs, the flux pulsations may be very high; specially when \(2p \approx Q\) and the pole-pitch is close to the slot-pitch, leading to high rotor losses [132]. The other cause of rotor losses are the MMF harmonics due to the non-sinusoidal winding distribution. Among the air-gap MMF harmonics, all but the main harmonic induce currents in the conductive parts of the rotor (\(\nu \neq p\)). Therefore, the higher the harmonic content of the air-gap MMF, the higher the rotor losses. Of special consideration are the MMF subharmonics (\(\nu < p\)), since they lead to high and fast flux pulsations [188]. The effect of MMF subharmonics in the performance of the machine will be addressed in a later subsection.

The effect of MMF harmonics in FSCW machines with isotropic and anisotropic (e.g. reluctance) rotors is discussed in [126]. In [189] a rotor loss index is introduced in order to compare the effect of the winding configuration in the rotor losses of the machine. Since the proposed index is independent of the geometry of the machine, it allows for a comparison among the different winding topologies to be made. With reference to FSCW machines, such an index can be used to choose among different phase, pole and slot combinations [147]. In the same fashion, [137] derives a figure of merit for rotor losses based on an analytical model. The model is able to take into account losses in the back iron, permanent magnets, non-magnetic retaining sleeves and copper cladding (if any). The effect of the number of phases and pole and slot combinations in rotor losses is addressed. The effect of MMF harmonics in the losses of rotor clamping rings is investigated in [168, 190].

In general terms, FSCWs lead to a higher MMF harmonic content and, thus, higher rotor losses than integral-slot windings. As an example, figures 3.6 and 3.7 show the ideal air-gap MMF waveforms and the MMF harmonics of an integral-slot winding machine and a FSCW machine. Both machines have \(m = 3\) and \(p = 5\).

![MMF waveform](image)

(a) Integral-slot winding (no short-pitching) \((m = 3, p = 5, Q = 30, q = 1)\)

(b) Double-layer FSCW \((m = 3, p = 5, Q = 12, q = 2/5)\)

Figure 3.6: Ideal air-gap MMF waveform for integral-slot and FSCW machines

It is readily noticeable from Figure 3.7 the higher MMF harmonic content of the 2-layer FSCW machine with respect to the same pole number integral-slot winding machine. Single-layer FSCWs lead to even higher MMF harmonics and rotor losses than their corresponding 2-layer windings. Specifically, single-layer windings lead to higher amplitude subharmonics. On the contrary, using a 4-layer winding reduces the harmonic content of the air-gap MMF distribution. Figures 3.8 and 3.9 show the air-gap MMF waveforms and the MMF harmonics for the 1-layer and 4-layer FSCW solutions corresponding to \(m = 3\) and \(p = 5\). Of particular interest is the decrease of subharmonics when using a 4-layer winding, as
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Figure 3.7: MMF harmonic content for integral-slot and FSCW machines compared to its equivalent 2-layer winding. The amplitude reduction for the main harmonic and the slot harmonics ($\nu = kQ \pm p, k \in \mathbb{N}$) is low [191].

Figure 3.8: Ideal air-gap MMF waveform for two FSCW machines ($m = 3, p = 5, Q = 12$)

Figure 3.9: MMF harmonic content for two FSCW machines ($m = 3, p = 5, Q = 12$)

The orders of the MMF harmonics generated by odd phase number FSCWs are collected in Tables 3.1 and 3.2 [133]. According to [137], 2-layer FSCWs lead to lower rotor losses than the corresponding 1-layer windings, but the relative benefit lessens as the number of phases increases. Also, the difference becomes less significant as the number of poles and slots increases. In general terms, rotor losses tend to decrease when the number of poles and slots increase, while keeping the same $2p/Q$ ratio.
CHAPTER 3. WINDING SELECTION CRITERIA FOR FAULT-TOLERANT PMSMS

In addition to selecting an adequate phase, pole and slot combination, a number of design techniques are available to reduce rotor losses in FSCW PM machines, such as [131]:

- Use of a laminated rotor, instead of a solid one.
- Axial and/or circumferential segmentation of the permanent-magnets [148,149,192].
- Use of a higher resistivity PM material (e.g. bonded vs sintered) [193].
- Shielding by copper cladding [137,194].

3.1.8 Armature MMF subharmonics

MMF subharmonics are those with harmonic order $\nu < p$. As can be appreciated from Figures 3.7 and 3.9, FSCWs create subharmonics in the air-gap MMF distribution; being the order and the amplitude of the MMF subharmonics a function of the phase, pole and slot number combination. In general, the lower the number of layers for a particular combination, the higher the MMF subharmonic content. Almost all FSCWs, except those with $q = 1/2$, lead to armature MMF subharmonics [132].

The presence of MMF subharmonics is detrimental for the operation of the machine due to a number of side-effects. In the first place, subharmonic flux components penetrate deeply in the rotor and may cause significant eddy-current losses in the rotor conductive parts [126]. In [147] it was demonstrated how the air-gap of the machine acts as a kind of “low-pass filter” regarding rotor losses; meaning that low order harmonics might have a larger influence on rotor losses than high order ones. As the power loss density is material dependent, it was shown in the same paper that permanent-magnets are specially sensitive to low order MMF harmonics. In [188] it is argued that armature MMF subharmonics are the main cause of rotor losses in FSCW machines, due to their high amplitude.

Another consequence of MMF subharmonics is an unbalanced saturation of the iron parts among the rotor poles. This can lead to an increased torque ripple; specially in small air-gap machines [126]. Additionally, MMF subharmonics lead to low order radial forces between the rotor and the stator of the machine. This means that depending on the application, the stator yoke may have to be increased in order to avoid excessive low order vibrations [75]. As well as radial vibration of the stator, additional torque ripple may arise due to the interaction of low order MMF harmonics with low order variations in the permanent-magnet flux; due, for instance, to rotor eccentricity [75].

Figure 3.10 represents the armature flux distribution in two FSCW permanent magnet machines. The first machine corresponds to a single-layer example, while the second machine corresponds to a double-layer one. It can be easily spotted how the highest amplitude harmonic is the main harmonic for the 2-layer machine ($\nu = p = 5$), while it is the first order subharmonic for the 1-layer machine ($\nu = 1$). The higher penetration of the lower order harmonics in the rotor can be appreciated as well.

When the number of poles is higher than the number of slots ($2p > Q$), additional subharmonics are generated [137]. Double-layer windings with $t = p$ do not lead to MMF subharmonics and thus, can be
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3.1.9 Peak armature reaction

The maximum value of the armature MMF reaction is related to the high amplitude of MMF subharmonics in FSCW machines [126]. Figures 3.6 and 3.8 show how the peak value of the armature MMF distribution can be very different depending on the number of layers. In general terms, 1-layer FSCWs lead to the highest peak value of armature reaction.

High armature reaction can lead to undesirable local saturation, permanent magnet demagnetization, unbalanced magnetic pull, decrease of the average torque and increased torque ripple [126, 138]. This effect is specially relevant when the machine is subject to high loads or when the machine air-gap is small.

3.1.10 Phase self-inductance

Having a sufficiently high phase self-inductance is a requirement for machines designed for fault-tolerant or wide constant power speed range applications [66, 144]. In the first case, a high phase inductance helps limit the fault currents in case of a short-circuit, while in the second case, having a per-unit $d$-axis inductance and thus, a characteristic current equal to the rated current, grants an ideally infinite flux weakening range to a surface-mounted PM machine.

FSCWs inherently lead to higher phase inductance values than integral-slot windings with the same number of poles [132]. Among the different layer possibilities, 1-layer windings give rise to the highest self-inductance values. The higher harmonic content of the armature MMF distribution causes higher air-gap harmonic leakage components and therefore, higher phase self-inductance [131]. The slot leakage inductance increases as well due to the increase of the number of conductors per slot belonging to the same phase [126]. According to the same reference, the air-gap inductance component is twice as big in single-layer windings as compared to double-layer ones.

All the combinations with $2p > Q$, exhibit an inherently high inductance due to the rich content of MMF subharmonics [138].

3.1.11 Mutual inductance between phases

Designing the machine in order to have a low mutual inductance between phases is another requirement for fault-tolerant applications [66]. With a reduced magnetic coupling between phases, even under short-circuit conditions, the flux produced by the healthy phases is not linked by the faulted phase and the short-circuit current becomes limited. In a similar manner, a low mutual inductance between phases
prevents the healthy phases to be affected by the faulted phase and allows them to continue to operate normally [123].

FSCW machines are characterized by having lower mutual inductance values than the same number of poles integral-slot machines [126]. Additionally, 1-layer FSCWs do not give rise to a slot leakage component of the mutual inductances and therefore, yield a reduction of the mutual inductance with respect to the corresponding 2-layer windings. Some phase, pole and slot combinations are particularly advantageous to achieve a low mutual inductance between phases. In [75] a rule to choose favorable combinations with any phase number is introduced. However, the study is limited to a special winding configuration with single-layer windings and an even number of anti-phase coils per phase. Moreover, the study is related to reduced winding systems (electrical angle between phase windings of $\alpha_m = \pi/m$).

The authors in [126, 133] introduce some criteria to select combinations that lead to a negligible mutual inductance between phases with both 1-layer and 2-layer windings. The rules, stated for odd phase number machines, are as follows:

- The coil throw must equal one slot, $y_q = 1$.
- The number of spokes in the star of slots must be an even number, so:

$$\begin{align*}
\frac{Q}{2t} & \text{ must be even for 1-layer windings} \\
\frac{Q}{t} & \text{ must be even for 2-layer windings}
\end{align*}$$

When the rule regarding the number of spokes applies, the MMF contribution of each coil belonging to a phase is compensated by the negative contribution of another coil of that same phase. In consequence, the section of the air-gap affected by the MMF created by a phase is limited and the magnetic coupling between phases is kept at minimum [126]. For instance, Figures 3.11 and 3.12 show the winding layout, the corresponding star of slots and the air-gap MMF distribution when only one phase is supplied for two 3-phase double-layer machines.

As it can be appreciated in the figures, for the “$\frac{Q}{t}$ even” machine, there are couples of opposite phasors in the star of slots and hence, the contribution to the air-gap MMF distribution of each coil is compensated by a coil with opposite sign. Consequently, the air-gap MMF waveform equals zero everywhere except over the teeth where the coils belonging to the fed phase are placed. On the contrary, for the “$\frac{Q}{t}$ odd” machine, there are no couples of opposing phasors and a non-null MMF distribution that leads to a mutual coupling between phases is created. Although the considered examples have a double-layer winding, a negligible mutual inductance is achieved with single-layer windings when the number of spokes is even, as well [126].

A special case arises for 2-layer windings for which $2p = Q \pm 2t$. In these cases, adjacent teeth are wound by coils belonging to the same phase but opposing sign [126]. The MMF contribution of one coil is directly compensated by the contribution of the adjacent coil, limiting the armature flux path and reducing the magnetic loading on the stator yoke [133]. The double-layer machine example with “$\frac{Q}{t}$ even” shown in Figures 3.11b and 3.12b is an example of such case. Figure 3.13 compares the armature flux distribution when only one phase is fed for that same combination with 1-layer and 2-layer windings.

While in the 2-layer case, only part of the stator yoke carries the flux of one phase, when a single-layer winding is used, the whole stator yoke carries the magnetic flux produced by all the phases. Although in linear conditions there is no difference between both configurations with reference to magnetic coupling, when a non-linear, saturable stator material is used, such as a commonplace silicon steel lamination, an undesirable mutual coupling between phases may arise due to magnetic saturation [126].

### 3.1.12 Number of phases

Having more than 3 phases grants additional degrees of freedom for machine control and provides for an enhanced fault-tolerance over 3-phase machines. In fact, in case of a multiphase machine, an adequate fault-tolerant operation can be achieved just by modifying the control law, without any additional hardware [23].
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The choice of the number of phases in a fault-tolerant PMSM is mainly the result of a compromise between the increased fault-tolerance and the higher weight, cost and complexity of the associated inverter [69]. Although the size and power rating of the individual converters is reduced and simpler connections for the semiconductor switches may be used, there is a concern that the overall inverter size and complexity might be increased, putting at stake the inverter reliability [137].

The additional advantages of having an increased phase number besides an improved fault-tolerance, such as an increased torque density, improved MMF waveform, lower losses, etc. have already been covered in the previous chapter.
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Figure 3.13: Armature flux distribution in FSCW machines with \(2p = Q \pm 2t\) when only one phase is fed \((m = 3, p = 5, Q = 12)\)

3.2 Selection based on inductance calculation

Previous work regarding phase, pole and slot selection for fault-tolerant applications primarily focuses on obtaining 1-layer FSCWs [75,125]. Although 1-layer FSCWs have been mostly adopted in fault-tolerant designs, it is also possible to design 2-layer winding machines with a high phase self-inductance and a reduced mutual inductance between phases [63]. Double-layer and 4-layer FSCWs offer some advantages over their 1-layer counterparts, such as shorter end-windings, lower MMF harmonic content and thus, lower rotor losses [131,154].

On the other hand and, as already mentioned, the authors in [126, 133] introduce some criteria to select combinations that lead to a negligible mutual inductance between phases with both 1-layer and 2-layer windings. However, the derived rules are restricted to odd phase number machines. Single-layer combinations with any phase number are investigated in [75], but the study is limited to a special winding configuration with an even number of anti-phase coils per phase. Moreover, no numerical comparison regarding mutual inductances is generally established among the most promising combination candidates.

This section aims to fill this gap by studying the relationship between self-inductance and mutual inductance values in radial-flux PMSM with either 1-layer, 2-layer or 4-layer FSCWs. In order to do so, analytical expressions for the calculation of the inductances are derived.

Traditionally, the inductances of an electrical machine have been divided into a number of components according to the following classification of fluxes [134,135]:

- Magnetizing flux across the air-gap.

- Leakage fluxes:
  - Crossing the air-gap: differential leakage flux.
  - Not crossing the air-gap: slot, pole, tooth-tip and end-winding leakage fluxes.

In permanent magnet machines, the slot and air-gap components of the inductance tend to dominate over the rest [135,195]. Therefore, only these components are considered in the study.

This section is organized in the following way: first, analytical expressions for the slot-component of the self and mutual inductances in an electrical machine are obtained. Studying the derived formulas, a number of conclusions are extracted. The same process is conducted for the air-gap region of the machine. Next, an application example is presented, whose results are validated by a Finite Element Analysis (FEA). Finally, the conclusions of the study are summarized.
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

3.2.1 Self and mutual inductances - Slot component

Analytical expressions for the slot inductances

Traditionally, the slot leakage inductances have been related to the slot geometry by means of permeance coefficients, \( \lambda_{slot} \): \[ L_{slot,ii} = 2 \mu_0 l_{ef} \frac{N_i^2}{p} \lambda_{slot,ii} \] \[ L_{slot,ij} = -2 \mu_0 l_{ef} \frac{N_i^2}{p} \lambda_{slot,ij} \]

For a semi-closed rectangular slot machine as the one depicted in Figure 3.14, and under some simplifications, the magnetic field distribution in the slot can be analytically calculated, leading to the following expressions for the permeance coefficients of the self-inductance and the mutual inductance between phases (see Appendix C):

![Figure 3.14: Simplified slot geometry and nomenclature](image)

Expressions (3.7) and (3.8) include the terms corresponding to the leakage across the slot-opening region. Coefficients \( \beta_{ii}, \gamma_{ii}, \delta_{ii}, \epsilon_{ii} \) (3.7), \( \beta_{ij}, \gamma_{ij}, \delta_{ij} \) and \( \epsilon_{ij} \) (3.8) are dimensionless parameters that depend solely on the winding arrangement (see Appendix C). For a proper fault-tolerant machine design, \( \lambda_{slot,ii} \) should be high enough to have a unity phase self-inductance, while \( \lambda_{slot,ij} \) should tend to zero. The assumption of a simplified slot geometry is justified by the fact that the present section aims to study the suitability of different winding combinations regardless of the geometry of the machine.

Study of the slot inductance coefficients

The following section covers the study of the slot inductance coefficients for FSCWs with a different number of layers. Specifically, 1-layer, 2-layer and 4-layer windings with a number of phases ranging from...
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3 to 7 are investigated. Non-reduced winding systems are considered for 4 and 6-phase machines and normal systems for 3, 5 and 7-phase machines (see Appendix A). In all cases, an electrical angle between phase windings of $\alpha_m = \frac{2\pi}{m}$ is contemplated. The windings have been designed by means of the star of slots [133].

**Slot inductance coefficients: 1-layer FSCWs** Single-layer FSCWs are characterized by having an alternate teeth wound configuration. Figure 3.15 shows an example of a single-layer FSCW permanent magnet machine.

![Figure 3.15: 1-layer FSCW machine ($m = 3, p = 5, Q = 12$)](image)

For 1-layer FSCWs, the *sign* type vectors that define the winding arrangement and that are defined in equation (C.22) satisfy the following property:

$$[k_{UL,j}] = [k_{UR,j}] = [k_{DL,j}] = [k_{DR,j}] = [k_j], \forall j \in [1,Q]$$

so that:

$$
\begin{cases}
[\beta] = \frac{m}{qQ} \sum_{j=1}^{Q} ([k_j]^T[k_j]) = \frac{m}{qQ} \frac{Q}{m} [I_m] = \frac{1}{q} [I_m] \\
[\gamma] = [0_m] \\
[\delta] = [0_m] \\
[\epsilon] = \frac{m}{qQ} \sum_{j=1}^{Q} ([k_j]^T[k_j]) = \frac{1}{q} [I_m]
\end{cases} \tag{3.10}
$$

where $[I_m]$ and $[0_m]$ stand for the identity and null matrices of size $m \times m$, respectively. In summary:

$$
\begin{cases}
q\beta_{ii} = 1 & q\gamma_{ii} = 0 & q\delta_{ii} = 0 & q\epsilon_{ii} = 1 \\
q\beta_{ij} = 0 & q\gamma_{ij} = 0 & q\delta_{ij} = 0 & q\epsilon_{ij} = 0
\end{cases} \tag{3.11}
$$

Equation (3.11) demonstrates that the slot component of the mutual inductance between phases in a single-layer FSCW machine is ideally null; fact that is exploited in fault-tolerant machines [66].

**Slot inductance coefficients: 2-layer FSCWs** Double-layer FSCWs are characterized by having all teeth wound by coils spanning just one slot-pitch. Different options are possible with this type of winding, depending on the arrangement of the coils within the slots and on the connection between them. In the present thesis the options have been labeled as *vertical*, *horizontal* and *diagonal*. Figure 3.16 shows the different possible coil arrangements for a PMSM where $m = 3, p = 5, Q = 12$.

For $m$-phase, balanced windings, both the *vertical* and *diagonal* arrangements give rise to the same number of coils placed in the UL, UR, DL and DR zones of the slots for the different phases. This,
### 3.2. SELECTION BASED ON INDUCTANCE CALCULATION

(a) Vertical  
(b) Horizontal  
(c) Diagonal

Figure 3.16: 2-layer FSCW machines \((m = 3, p = 5, Q = 12)\)

however, does not always happen for horizontal type windings, which can lead to a magnetic unbalance, depending on the phase, pole and slot number combination. When the number of phases is 3, 5 or 7, the horizontal and the diagonal arrangement of the coils lead to the same distribution of the coils inside the slots, as can be seen in Figures 3.16b and 3.16c. This is not the case for an even number of phases, where examples of unbalanced windings exist. Figure 3.17 shows an example of the winding distribution of a 4-phase, 2-layer, unbalanced FSCW with a horizontal coil arrangement.

![Figure 3.17: 2-layer horizontal unbalanced FSCW \((m = 4, p = 1, Q = 4)\)](image)

For the winding arrangement show in Figure 3.17, it is:

\[
\begin{align*}
\beta_{AA} &= \beta_{CC} = 2 \\
\beta_{BB} &= \beta_{DD} = 1/2 \\
\beta_{AA} &\neq \beta_{BB}
\end{align*}
\]

(3.12)

Since unbalanced windings can lead to parasitic phenomena such as local saturation, increased torque ripple or unbalanced radial forces in the rotor, the present study considers the cases where no magnetic unbalance occurs, namely the vertical and diagonal cases.

**Vertical 2-layer FSCWs**  In this case:

\[
\begin{align*}
[k_{UL,j}] &= [k_{DL,j}] = [k_{L,j}] \\
[k_{UR,j}] &= [k_{DR,j}] = [k_{R,j}]
\end{align*}
\]

(3.13)
so that:

\[
\begin{align*}
[\beta] &= \frac{m}{4qQ} \sum_{j=1}^{Q} \left( ([k_{R,j}] + [k_{L,j}])^T ([k_{R,j}] + [k_{L,j}]) \right) \\
[\gamma] &= \frac{m}{4qQ} \sum_{j=1}^{Q} \left( ([k_{R,j}] - [k_{L,j}])^T ([k_{R,j}] - [k_{L,j}]) \right) \\
[\delta] &= [0_m] \\
[\epsilon] &= [\beta]
\end{align*}
\] (3.14)

By studying the matrices above, a number of properties are derived for the slot inductance coefficients. For double-layer vertical FSCWs where \( m = 3 - 7 \), it holds that:

1. \( q\beta_{ii} \geq 0 \).
2. \( q\gamma_{ii} \geq 0 \).
3. \( q\beta_{ii} + q\gamma_{ii} = 1 \).
4. \( q\beta_{ij} + q\gamma_{ij} = 0 \), \( i \neq j \).
5. \( q\delta_{ii} = q\delta_{ij} = 0 \).
6. \( q\beta_{ii} = q\epsilon_{ii}, q\beta_{ij} = q\epsilon_{ij} \).
7. Coefficients \( q\beta_{ii} \) and \( q\beta_{ij} \) are solely a function of the number of phases and the number of slots per pole and phase, \( q \). The remaining coefficients are derived by the relationships stated above.
8. All coefficients are periodical in the variable \( 2p/Q \) (that is, the number of poles per slot). The period of the coefficients in \( 2p/Q \) is 2, or fractions of 2.
9. There are regions in the \( 2p/Q \in (\frac{1}{m}, \infty) \) space for which \( q\beta_{ii} \neq q\gamma_{ii} \). Outside these regions and at the end-points, it holds that \( q\beta_{ii} = q\gamma_{ii} = 1/2 \).
10. For \( m = 3, 5 \) and 7:
   - The aforementioned regions lie between \( 2p/Q \in [k - 1/m, k + 1/m], k \in \mathbb{N} \).
   - Outside the regions, the slot components of the mutual inductance are zero except for the electrically most separated phases, \( q\beta_{ij} = q\gamma_{ij} = 0 \).
   - Inside the regions, the factors \( q\beta_{ii} \) and \( q\beta_{ij} \) have relative maxima and minima at \( 2p/Q = k \).
   - For the electrically most separated phases, factors \( q\beta_{ij} \) also have relative maxima and minima at the end-points of the regions.
   - Factors \( q\beta_{ii} \) vary in the range from 0 to 1, while factors \( q\beta_{ij} \) vary from \(-1/4\) to \(1/4\).
11. For \( m = 4 \) and 6:
   - There are double regions, with two relative maxima in the variables \( q\beta_{ii} \) and simple regions, with just one relative minimum in the same variables.
   - The double regions correspond to \( 2p/Q \in [k - 2/m, k + 2/m], k \in \mathbb{N}, k \) odd, while the simple regions correspond to \( 2p/Q \in [k - 1/m, k + 1/m], k \in \mathbb{N}, k \) even.
   - In the double regions \( q\beta_{ii} > 1/2 \). In these regions \( q\beta_{ii} \) have two relative maxima at \( 2p/Q = k - 1/m \) and \( 2p/Q = k + 1/m \) and a relative minimum at \( 2p/Q = k \). In the relative maxima \( q\beta_{ii} \) reach the value of 3/4, while in the relative minimum they descend to 1/2.
   - In the simple regions \( q\beta_{ii} < 1/2 \). In these regions \( q\beta_{ii} \) has one 0-valued relative minimum at \( 2p/Q = k \).
12. The aforementioned properties can be observed in Figure 3.18, in which some of the factors for vertical 3-phase and 4-phase 2-layer FSCWs are shown.
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

Usually pole and slot combinations with a ratio $2p/Q$ close to unity are chosen in order to maximize the winding factor for the main harmonic and reduce the length of the end-windings [132]. In the region $2p/Q \in [1 - 1/m, 1 + 1/m]$, defining parameter $z$ as $z = \left| \frac{2p}{Q} - 1 \right|$, the following relations hold:

- For $m = 3, 5$ and $7$, $q_{\beta ii} = 1 - \frac{m}{2} z$ and $q_{\beta ij} = \frac{m}{4} z$ for the electrically most separated phases. For the remaining combinations $q_{\beta ij} = 0$.

- For $m = 4$ and $6$, $q_{\beta ii} = \frac{1}{2} + \frac{m}{4} z$, $q_{\beta ij} = \frac{1}{2} - \frac{m}{4} z$ for the electrically most separated phases and $q_{\beta ij} = -\frac{m}{8} z$ for the electrically closer phases. For the remaining combinations, $q_{\beta ij} = 0$.

Diagonal 2-layer FSCWs In this case:

$$\begin{align*}
[k_{UL,j}] &= [k_{UR,j}] = [k_{U,j}] \\
[k_{DL,j}] &= [k_{DR,j}] = [k_{D,j}]
\end{align*}$$

so that:

$$\begin{align*}
[\beta] &= \frac{m}{8qQ} \sum_{j=1}^{Q} [[k_{D,j}]^T[k_{D,j}] + 3[k_{D,j}]^T[k_{U,j}] + 4[k_{U,j}]^T
[k_{U,j}]] \\
[\gamma] &= [0_m] \\
[\delta] &= [0_m] \\
[\epsilon] &= \frac{m}{4qQ} \sum_{j=1}^{Q} \left( ([k_{U,j}] + [k_{D,j}])^T([k_{U,j}] + [k_{D,j}]) \right)
\end{align*}$$

Matrix $[\epsilon]$ in (3.16) is the same as for vertical 2-layer FSCWs and its properties have been stated in the previous subsection. In the case of coefficients $q_{\beta ii}$ and $q_{\beta ij}$, their evolution is similar to the vertical arrangement case, but with the shape of the curves modified by an offset and a scale factor. The evolution of some of the factors for diagonal 2-layer FSCWs is shown in Figure 3.19. It holds that:

1. $q_{\beta ij} = \frac{3}{4} q_{\epsilon ij}$, $i \neq j$.
2. For the regions where $q_{\epsilon ii} = 1/2$, $q_{\beta ii} = 5/8$.
3. For $m = 3, 5$ and $7$, $q_{\beta ii}$ ranges from $1/4$ to $1$, while for $m = 4$ and $6$, said factor has two $13/16$-valued maxima per period at $2p/Q = k - 1/m$ and $2p/Q = k + 1/m$, $k \in \mathbb{N}$, $k$ odd, and $1/4$-valued minimum at $2p/Q = k$, $k$ even.
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For windings with a $2p/Q$ ratio in the interval $[1 - 1/m, 1 + 1/m]$:

- For $m = 3, 5$ and $7$, $q_{\beta_{ii}} = 1 - \frac{3m}{4}z$ and $q_{\beta_{ij}} = \frac{3m}{16}z$ for the electrically most separated phases. For the remaining combinations, $q_{\beta_{ij}} = 0$.
- For $m = 4$ and $6$, $q_{\beta_{ii}} = \frac{5}{8} + \frac{3m}{16}z$, $q_{\beta_{ij}} = \frac{3}{8} - \frac{3m}{32}z$ for the electrically most separated phases and $q_{\beta_{ij}} = -\frac{3m}{32}z$ for the electrically closer phases. For the remaining combinations, $q_{\beta_{ij}} = 0$.

**Slot inductance coefficients: 4-layer FSCWs**  
4-layer FSCWs have been recently introduced as an extension of 2-layer windings in order to reduce the harmonic content of the armature MMF [154]. Figure 3.20 shows an example of a 4-layer FSCW for a PMSM where $m = 3$, $p = 5$, $Q = 12$.

![4-layer FSCW](image)

**Figure 3.20: 4-layer FSCW machine ($m = 3$, $p = 5$, $Q = 12$)**

When extending a 2-layer winding into a 4-layer one, there are different "optimization" strategies that seek one of the following goals [154]:

- to maximize the winding factor for the main harmonic.
- to minimize a specific MMF harmonic.
- to reduce the number of coils (e.g. 3 coil sides per slot).

In order to maximize the winding factor for the main harmonic while reducing the harmonic content of the MMF distribution, the angular shift between layers must be kept as low as possible. Two distinct cases have to be considered depending on if $Q/t$ is an even or odd number (where $t = \gcd(p, Q)$ is the electric periodicity of the machine) [154]. The machine shown in Figure 3.20 is an example of a machine with $Q/t$ even. In case of $Q/t$ odd, two possibilities arise depending on how the layers are shifted (solutions (i) and (ii)). Figure 3.21 shows an example of a machine with $Q/t$ odd and the different solutions implemented:

Although expanding a 2-layer winding into a 4-layer winding is always feasible, not all 4-layer combinations are convenient in terms of reducing the harmonics of the MMF. Some of the combinations give
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

Figure 3.21: 4-layer FSCW machines ($m = 3, p = 4, Q = 9$)

rise to the same amplitude decrease for all the space-harmonics and thus, adopting a 4-layer FSCW in those cases makes no sense [154]. In the present study, only balanced and symmetrical $m$-phase, 4-layer FSCWs that lead to a convenient decrease of the MMF harmonics have been considered.

Unlike the slot inductance coefficients for 1-layer and 2-layer FSCWs, no general rules seem to exist in order to determine coefficient matrices $[\beta]$, $[\gamma]$, $[\delta]$ and $[\epsilon]$ for 4-layer windings. Therefore, they have to be calculated on an individual case basis for each phase, pole and slot combination. Nevertheless, some general properties of the coefficients can be stated:

1. Coefficients $q\beta_{ii}$ are upper-bounded by the equivalent coefficients for vertical 2-layer FSCWs. Likewise, coefficients $q\beta_{ij}$ are lower-bounded by the equivalent coefficients for vertical 2-layer windings.
2. Coefficients $q\gamma_{ii}$ and $q\gamma_{ij}$ perfectly match the equivalent coefficients for vertical 2-layer FSCWs.
3. Coefficients $q\delta_{ii}$ are upper-bounded by a triangular function periodic in $2p/Q$. The period of the function is $\frac{1}{m}$ and its maximum value is $1/8$. Coefficients $q\delta_{ij}$ are lower-bounded by an equivalent negative triangular function, whose minimum value is $-\frac{1}{16}$.
4. Coefficients $q\epsilon_{ii}$ and $q\epsilon_{ij}$ follow the same rules as coefficients $q\beta_{ii}$ and $q\beta_{ij}$, respectively, but they are generally unequal.

The aforementioned properties can be observed in Figure 3.22, in which the slot inductance factors for 5-phase, 4-layer FSCWs are shown.

**Figure of merit**

In order to identify which combinations of phases, poles and slots are more advantageous to grant fault-tolerance to a PMSM, the following figure of merit is introduced:

$$\psi = \sum_{i \neq j}^{m} \frac{|L_{ij}|}{L_{ii}}$$

(3.17)

Since a value of $\psi$ close to zero implies a reduced mutual inductance between phases, said factor allows for a comparison in terms of the machine’s fault tolerance among the different possible combinations to be made. If only the slot leakage inductances are considered:

$$\psi_{\text{slot}} = \sum_{i \neq j}^{m} \frac{|L_{\text{slot},ij}|}{L_{\text{slot},ii}} = \sum_{i \neq j}^{m} \frac{|\lambda_{\text{slot},ij}|}{\lambda_{\text{slot},ii}}$$

(3.18)
CHAPTER 3. WINDING SELECTION CRITERIA FOR FAULT-TOLERANT PMSMs

For 1-layer FSCWs, $\psi_{\text{slot}}$ equals 0.

For 2-layer vertical FSCWs:

\[
q_{\text{slot,ii}} = \frac{b}{12h} + q_{\epsilon,\text{ii}} \left( \frac{h}{3b} - \frac{b}{12h} + \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,}^{\infty} \frac{\sin^2 \left( \frac{h_0 \pi n}{b} \right)}{(\pi n)^3 \tanh \left( \frac{h_0 \pi n}{b} \right)} \right) \\
\geq q_{\epsilon,\text{ii}} \left( \frac{h}{3b} - \frac{b}{12h} + \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,}^{\infty} \frac{\sin^2 \left( \frac{h_0 \pi n}{b} \right)}{(\pi n)^3 \tanh \left( \frac{h_0 \pi n}{b} \right)} \right) \\
q_{\text{slot,ij}} = q_{\epsilon,ij} \left( \frac{h}{3b} - \frac{b}{12h} + \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,}^{\infty} \frac{\sin^2 \left( \frac{h_0 \pi n}{b} \right)}{(\pi n)^3 \tanh \left( \frac{h_0 \pi n}{b} \right)} \right)
\]

Therefore:

\[
\psi_{\text{slot}} = \sum_{i \neq j}^{m} \frac{\lambda_{\text{slot,ij}}}{\lambda_{\text{slot,ii}}} \leq \sum_{i \neq j}^{m} \left| \frac{\epsilon_{ij}}{\epsilon_{ii}} \right| = \varsigma
\]
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

- For 2-layer diagonal FSCWs:

\[
q_{\lambda_{\text{slot,ii}}} = q_{\beta_{ii}} \frac{h}{3b} + q_{\epsilon_{ii}} \left( \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,...}^{\infty} \frac{\sin^2 \left( \frac{b_0 \pi n}{2b} \right)}{(\pi n)^3 \tanh \left( \frac{b_0 \pi n}{b} \right)} \right)
\]
\begin{align*}
\geq q_{\epsilon_{ii}} \left( \frac{h}{3b} + \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,...}^{\infty} \frac{\sin^2 \left( \frac{b_0 \pi n}{2b} \right)}{(\pi n)^3 \tanh \left( \frac{b_0 \pi n}{b} \right)} \right) \\
q_{\lambda_{\text{slot,ij}}} = q_{\beta_{ij}} \frac{h}{3b} + q_{\epsilon_{ij}} \left( \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,...}^{\infty} \frac{\sin^2 \left( \frac{b_0 \pi n}{2b} \right)}{(\pi n)^3 \tanh \left( \frac{b_0 \pi n}{b} \right)} \right)
\end{align*}
\]
\[
\leq q_{\epsilon_{ij}} \left( \frac{h}{3b} + \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,...}^{\infty} \frac{\sin^2 \left( \frac{b_0 \pi n}{2b} \right)}{(\pi n)^3 \tanh \left( \frac{b_0 \pi n}{b} \right)} \right)
\]

and thus:

\[
\psi_{\text{slot}} = \sum_{i\neq j}^m \left| \frac{\lambda_{\text{slot,ij}}}{\lambda_{\text{slot,ii}}} \right| \leq \sum_{i\neq j}^m \left| \frac{\epsilon_{ij}}{\epsilon_{ii}} \right| = \varsigma
\]

Unlike \( \psi_{\text{slot}} \), which depends on the geometry of the machine, parameter \( \varsigma \) depends solely on the chosen combination of phases, poles and slots. Figure 3.23 shows the value of \( \varsigma \) in terms of the number of poles per slot, \( 2p/Q \), for 2-layer FSCW machines and different number of phases.

![Figure 3.23: Factor \( \varsigma \) for 2-layer FSCWs](image)

Since the main winding factor for FSCW machines increases as the number of poles per slot tends to unity, the most favorable combinations in terms of fault-tolerance for 2-layer FSCWs with \( m = 3, 5 \) and 7 are those in which \( 2p \approx Q \), as they lead to high winding factors and reduced slot components of the mutual inductance between phases. However, for FSCWs where \( m = 4 \) and 6, the combinations that give rise to a lower mutual coupling between phases do not correspond to those which lead to a high winding factor. For \( m = 4 \), the zeros of \( \varsigma \) are in \( 2p/Q = 2/3, 4/3 \), combinations that lead to a poor winding factor, \( k_w,p = 0.7797 \). In the case of 6-phase machines, high winding factors for the main harmonic can be achieved at the expense of having a higher mutual coupling between phases than for other phase numbers.

Figure 3.23 indicates that the \( 2p \approx Q \) rule is more stringent for 2-layer machines with a high number of phases, seeing as the relationship between the mutual and the self-inductance increases faster as the \( 2p/Q \) ratio shifts away from unity.

- For 4-layer FSCWs, unfortunately, no general rule can be stated based on the slot leakage coefficients. However, in general terms, they lead to lower self-inductances and higher mutual inductances between phases than their corresponding 2-layer equivalents.
By way of example, the slot inductance coefficients are compared for the phase, pole and slot combinations shown in Figures 3.20 and 3.21 in terms of the number of layers. The results are shown in Tables 3.3 and 3.4.

A further comparison has been carried out for a single machine geometry and the different layer possibilities. The analysis has been conducted for a trapezoidal slot machine, the standard in low power applications (e.g. Figures 3.15, 3.16 and 3.20). The machine characteristics are shown in Table 3.5. The machine has been designed in order to have a per-unit phase inductance when a single-layer winding is employed. This grants an optimal field-weakening region and fault-tolerant capabilities [144].

Table 3.3: Slot inductance coefficients for 3-phase, 10-pole, 12-slot FSCWs

<table>
<thead>
<tr>
<th>$q^2_{AA}$</th>
<th>$q^2_{AB}$</th>
<th>$q^3_{AA}$</th>
<th>$q^3_{AB}$</th>
<th>$q^4_{AA}$</th>
<th>$q^4_{AB}$</th>
<th>$q^5_{AA}$</th>
<th>$q^5_{AB}$</th>
<th>$q^6_{AA}$</th>
<th>$q^6_{AB}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.8333</td>
<td>0.0833</td>
<td>0.1667</td>
<td>-0.0833</td>
<td>0.0000</td>
<td>0.0000</td>
<td>0.0000</td>
<td>0.0000</td>
<td>0.0000</td>
<td>0.0000</td>
</tr>
</tbody>
</table>

Table 3.4: Slot inductance coefficients for 3-phase, 8-pole, 9-slot FSCWs

<table>
<thead>
<tr>
<th>$q^2_{AA}$</th>
<th>$q^2_{AB}$</th>
<th>$q^3_{AA}$</th>
<th>$q^3_{AB}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.8333</td>
<td>0.0833</td>
<td>0.1667</td>
<td>-0.0833</td>
</tr>
</tbody>
</table>

Table 3.5: Specifications for 3-phase, 10-pole, 12-slot FSCW machines

<table>
<thead>
<tr>
<th>Rated torque, $T_{nom}$</th>
<th>50 [N-m]</th>
<th>RMS rated current, $I_{nom}$</th>
<th>51.26 [A]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Corner speed, $N_{nom}$</td>
<td>600 [rpm]</td>
<td>DC bus voltage, $U_{DC}$</td>
<td>72 [V]</td>
</tr>
<tr>
<td>RMS PM flux linkage*, $A_{m1}$</td>
<td>65.03 [mWb]</td>
<td>Base inductance*, $L_b$</td>
<td>1.269 [mH]</td>
</tr>
<tr>
<td>Outer diameter, $D_{ext,s}$</td>
<td>200 [mm]</td>
<td>Effective length, $l_e$</td>
<td>100 [mm]</td>
</tr>
<tr>
<td>Slot-opening height, $h_0$</td>
<td>5.6 [mm]</td>
<td>Wedge height, $h_1$</td>
<td>2.0 [mm]</td>
</tr>
<tr>
<td>Slot height, $h$</td>
<td>28.3 [mm]</td>
<td>Stator yoke thickness, $h_y,s$</td>
<td>9.95 [mm]</td>
</tr>
<tr>
<td>Slot-opening width, $b_0$</td>
<td>2.4 [mm]</td>
<td>Slot bottom width, $b_1$</td>
<td>19.0 [mm]</td>
</tr>
<tr>
<td>Slot top width, $b_2$</td>
<td>33.7 [mm]</td>
<td>Tooth width, $w_t$</td>
<td>13.2 [mm]</td>
</tr>
<tr>
<td>Magnet thickness, $h_m$</td>
<td>5 [mm]</td>
<td>Air-gap thickness, $g$</td>
<td>1 [mm]</td>
</tr>
<tr>
<td>Magnet remanence, $B_r$</td>
<td>1.15 [T]</td>
<td>Magnet relative permeability, $\mu_{r,m}$</td>
<td>1.1</td>
</tr>
<tr>
<td>Number of series turns, $N_s$</td>
<td>40</td>
<td>Parallel paths</td>
<td>No</td>
</tr>
</tbody>
</table>

* In the case of 1-layer FSCW

Machine inductances have been calculated using the formulas derived in the thesis (see Appendix C), a classical approach and FE simulations. Consequently, the comparison further serves to validate the analytical expressions deduced in the present work. End-leakage inductance values have been neglected in the calculation of the total inductances as they tend to be insignificant in FSCW machines. In order to account for the trapezoidal geometry, the mean slot width has been considered as parameter $b$ in equations (3.7) and (3.8). For the wedge region permeance, the following expression has been added to the proposed formulas [134]:
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

\[ \lambda_{\text{wedge}} = \epsilon \frac{2h_1}{b_1 + b_0} \]  
\[ (3.23) \]

In the classical formulation, the slot leakage flux is assumed to be perpendicular to the slot depth and the magnetic field intensity is considered to be dependent only on the depth coordinate \[196,197\]. For a trapezoidal geometry slot, this yields \[134,198\]:

\[ \lambda_{\text{slot}} = \beta k_t \frac{h}{3b_1} + \epsilon \frac{h_1}{b_1 - b_0} \log_e \left( \frac{b_1}{b_0} \right) + \epsilon \frac{h_0}{b_1} \]  
\[ (3.24) \]

where:

\[ k_t = \frac{4t^2 - t^4 (3 - 4 \log_t t) - 1}{4 (t^2 - 1)^2 (t-1)} \]
\[ t = \frac{b_2}{b_1} \]  
\[ (3.25) \]

Table 3.6 presents the self and mutual phase inductance calculation results using the different methods.

<table>
<thead>
<tr>
<th>Q/t</th>
<th>1-layer</th>
<th>2-layer vert.</th>
<th>2-layer diag.</th>
<th>4-layer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proposed</td>
<td>Slot self-inductance</td>
<td>0.5600</td>
<td>0.4231</td>
<td>0.4236</td>
</tr>
<tr>
<td></td>
<td>Slot mutual inductance</td>
<td>0.0000</td>
<td>-0.0685</td>
<td>-0.0682</td>
</tr>
<tr>
<td></td>
<td>( \psi_{\text{slot}} )</td>
<td>0.0000</td>
<td>0.3236</td>
<td>0.3222</td>
</tr>
<tr>
<td></td>
<td>( \psi_{\text{slot}} )</td>
<td>0.0000</td>
<td>0.3333</td>
<td>0.3333</td>
</tr>
<tr>
<td>Classical</td>
<td>Slot self-inductance</td>
<td>0.4850</td>
<td>0.3637</td>
<td>0.3685</td>
</tr>
<tr>
<td></td>
<td>Slot mutual inductance</td>
<td>0.0000</td>
<td>-0.0606</td>
<td>-0.0583</td>
</tr>
<tr>
<td></td>
<td>( \psi_{\text{slot}} )</td>
<td>0.0000</td>
<td>0.3333</td>
<td>0.3162</td>
</tr>
<tr>
<td>FEM</td>
<td>Slot self-inductance</td>
<td>0.5526</td>
<td>0.4172</td>
<td>0.4175</td>
</tr>
<tr>
<td></td>
<td>Slot mutual inductance</td>
<td>0.0000</td>
<td>-0.0674</td>
<td>-0.0672</td>
</tr>
<tr>
<td></td>
<td>( \psi_{\text{slot}} )</td>
<td>0.0000</td>
<td>0.3231</td>
<td>0.3219</td>
</tr>
<tr>
<td></td>
<td>Total self-inductance</td>
<td>0.9945</td>
<td>0.6534</td>
<td>0.6536</td>
</tr>
<tr>
<td></td>
<td>Total mutual inductance</td>
<td>-0.0014</td>
<td>-0.0757</td>
<td>-0.0755</td>
</tr>
</tbody>
</table>

Note: All inductance values given in per-unit system

Table 3.6: Self and mutual inductances for 3-phase, 10-pole, 12-slot FSCW example

Results show a significant improvement of the proposed formulas over the classical formulation. In the example, inductance values for vertical and diagonal 2-layer FSCWs are almost the same due to the fact that the slot leakage inductance is dominated by the slot-opening component. Said component is the same for both coil arrangements.

The example illustrates the importance of calculating precisely the slot leakage inductances in a surface-mounted permanent-magnet FSCW machine, as the slot component accounts for a large amount of the total inductance. Additionally, it shows how 4-layer FSCWs lead to lower self-inductances and higher mutual inductances between phases than the corresponding 2-layer combinations.

It must be noticed that the derivation of the analytical formulas and the FE simulations have been conducted under the assumption of infinite iron permeability. In order to account for the effect of iron saturation in the slot leakage, the slot and slot-opening widths can be artificially increased in the formulas as it is usually done with the rotor slots of induction machines \[198\]. A more comprehensive approach regarding iron saturation involves modelling the magnetic circuit of the machine in terms of a permeance network \[199–202\].
Summary

In short:

- For 1-layer FSCW machines, the slot component of the mutual inductance between phases is intrinsically null, there being no criterion, a priori, to select the most appropriate combinations of phases, poles and slots to provide fault-tolerance to the machine.

- For 2-layer FSCW machines where \( m = 3, 5 \) and 7 there are combinations that lead to a reduced mutual coupling between phases, while keeping high winding factors. These combinations correspond to \( 2p \approx Q \). Factor \( \varsigma \) can be used to confront different winding choices, allowing a comparison between the optimal number of phases in terms of fault-tolerance.

- Double-layer FSCWs where \( m = 4 \) and 6 are not suitable for fault-tolerant PMSMs since they do not simultaneously give rise to low mutual inductances between phases and high winding factors for the main harmonic.

- Additionally, from the previous section, it can be concluded that 4-layer FSCWs are not appropriate for fault-tolerant applications, as they generally lead to lower self-inductances and higher mutual inductances between phases than 2-layer machines with the same phase, pole and slot combination.

3.2.2 Self and mutual inductances - Air-gap component

Analytical expressions for the air-gap inductances

Again, the air-gap inductances can be expressed in terms of permeance coefficients:

\[
L_{\text{agap},ii} = 2\mu_0 l_{ej} \frac{N_s^2}{p} \lambda_{\text{agap},ii} 
\]

\[
L_{\text{agap},ij} = -2\mu_0 l_{ej} \frac{N_s^2}{p} \lambda_{\text{agap},ij} 
\]

Considering an air-gap region geometry as the one depicted in Figure 3.24, the magnetic field distribution in the air-gap can be analytically calculated as well, leading to the following expressions for the air-gap permeance coefficients (see Appendix D):

\[
\lambda_{\text{agap},ii} = \sum_{n=1,2,3...}^{\infty} \frac{1}{\pi n} \text{sinc}^2 \left( \frac{n\alpha_0}{2} \right) \left( \epsilon_{ii} + \zeta_{n,ii} \right) \left( \frac{R_s}{R_r} \right)^n + \left( \frac{R_r}{R_s} \right)^n
\]

Figure 3.24: Air-gap geometry and nomenclature
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

\[ \lambda_{agap,ij} = \sum_{n=1,2,3,...}^{\infty} \frac{1}{\pi n} \text{sinc}^2 \left( \frac{n\alpha_0}{2} \right) (\epsilon_{ij} + \zeta_{n,ij}) \left( \frac{(R_s/R_r)^n + (R_r/R_s)^n}{R_s/R_r} \right)^n \]  

(3.29)

Coefficients \( \epsilon_{ii} \) and \( \epsilon_{ij} \), follow the rules mentioned in the previous section, while coefficients \( \zeta_{n,ii} \) and \( \zeta_{n,ij} \) depend both on the winding topology and the subscript index, \( n \). These coefficients are periodic in the variable \( n \), where the period is equal to the number of slots, \( Q \), or fractions of said number.

Study of the air-gap inductance coefficients

Under the assumptions that the air-gap thickness is very small compared to the air-gap radius, \( g \ll R = \frac{R_s + R_r}{2} \), and \( \alpha_0 \approx 0 \), the air-gap permeance coefficients reduce to:

\[ \lambda_{agap} \approx \frac{R}{\pi g} \sum_{n=1,2,3,...}^{\infty} \frac{1}{n^2} (\epsilon + \zeta_n) \]  

(3.30)

Thus, the relationship between the air-gap inductances becomes independent of the machine’s geometry and a new factor that is solely a function of the winding distribution, \( \xi \), can be introduced:

\[ \psi_{agap} = \sum_{i \neq j}^{m} \left| \frac{\lambda_{agap,ij}}{\lambda_{agap,ii}} \right| \approx \sum_{i \neq j}^{m} \left| \sum_{n=1}^{\infty} \frac{1}{n^2} (\epsilon_{ij} + \zeta_{n,ij}) \right| = \xi \]  

(3.31)

FSCWs with an odd number of phases exhibit almost null mutual air-gap inductance when one of the following relationships applies [133,177]:

- \( Q/(2t) \) even for 1-layer windings.
- \( Q/t \) even for 2-layer windings.
- \( 2p = Q \pm 2t \).

where \( t \) is the electric periodicity of the machine, \( t = \text{gcd}(p,Q) \). These properties can be appreciated in Figure 3.25, in which parameter \( \xi \) is displayed for 5-phase FSCWs. For the sake of clarity, all figures represent the decimal logarithm of \( \xi \).

However, the aforementioned rules do not hold for FSCWs with an even number of phases, as it can be concluded from observing Figures 3.26a – 3.26c, where parameter \( \xi \) is represented for 4-phase windings. In the case of 4-phase and 6-phase FSCWs, combinations for which \( 2p = Q \pm 2t \) do not always lead to a negligible inductance between phases and in order to have couples of opposite MMF contributions in the air-gap [133], the preceding relationships should be modified into the new pair of rules \( (r_1) \) and \( (r_2) \) that are more restrictive than its previously stated counterparts:

- \( (r_1): Q/(2mt) \) even for 1-layer windings.
- \( (r_2): Q/(mt) \) even for 2-layer windings.

For 4-layer windings, there are no combinations that lead to a negligible coupling between phases.

The machine examples whose characteristics are reported in Table 3.5 illustrate the case of combinations that follow the \( (r_1) \) and \( (r_2) \) rules. As a further comparison, machine inductances have been calculated using the formulas derived in the thesis (see Appendix D) and FE simulations. Again, the comparison serves to validate the analytical expressions deduced in the present work. In the calculation of the air-gap inductances, surface magnets have been taken into account by calculating an effective air-gap thickness:

\[ g_{ef} = g + \frac{h_m}{\mu_r,m} \]  

(3.32)
CHAPTER 3. WINDING SELECTION CRITERIA FOR FAULT-TOLERANT PMSMs

Figure 3.25: $\log_{10} \xi$ for 5-phase FSCWs

Figure 3.26: $\log_{10} \xi$ for 4-phase FSCWs
3.2. SELECTION BASED ON INDUCTANCE CALCULATION

Table 3.7 presents the self and mutual phase inductance calculation results using both calculation methods.

<table>
<thead>
<tr>
<th></th>
<th>Proposed</th>
<th>FEM</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1-layer</td>
<td>2-layer</td>
</tr>
<tr>
<td>Air-gap self-inductance</td>
<td>0.4462</td>
<td>0.2387</td>
</tr>
<tr>
<td>Air-gap mutual inductance</td>
<td>0.0000</td>
<td>−0.0078</td>
</tr>
<tr>
<td>$\psi_{agap}$</td>
<td>0.0000</td>
<td>0.0010</td>
</tr>
</tbody>
</table>

Note: All inductance values given in per-unit system

Table 3.7: Self and mutual inductances for 3-phase, 10-pole, 12-slot FSCW example

The obtained results show a remarkable agreement between the proposed formulas and the FE method. The notable contrast between parameters $\xi$ and $\psi_{agap}$ for 2-layer windings lies on the fact that the approximations upon which the relation "$\psi_{agap} \approx \xi$" has been established are not totally valid ($g_{ef}/R = 0.1079, \alpha_0 = 0.0443$ [rad]). Anyhow, tendencies according to the number of layers still hold. Moreover, the example restates how 4-layer FSCWs lead to low self-inductances and high mutual inductances between phases.

In addition to allowing a comparison between pole and slot combinations for a given phase number to be made, factor $\xi$ provides for a means of selecting the optimal number of phases in order to achieve the lowest possible mutual inductance between phases. Figures 3.27a and 3.27b show the value of $\log_{10} \xi$ for a number of combinations that follow the $(r_1)$ and $(r_2)$ rules, respectively, with the phase number ranging from 3 to 7. The figures indicate that, for combinations following the $(r_1)$ and $(r_2)$ rules, the increase in factor $\xi$ when parameter $2p/Q$ shifts away from unity is higher for machines with a high number of phases. For 2-layer windings, the minimum attainable relationship between the air-gap permeance coefficients is higher than for its 1-layer counterparts. When the number of phases is odd, factor $\xi$ has relative minima at $2p/Q = 1/2, 2/2, 3/2...$. For 2-layer combinations with $m = 4$ and $6$, $\xi$ reaches its minimum values for points different from $2p/Q = 1$.

![Figure 3.27: $\log_{10} \xi$ for FSCWs following the $(r_1)$ and $(r_2)$ rules](image)
CHAPTER 3. WINDING SELECTION CRITERIA FOR FAULT-TOLERANT PMSMS

Summary

As a conclusion of the study:

- Lowest mutual inductance values are obtained for 1-layer windings with $Q/(2mt)$ even. Combinations for which $2p/Q \approx 1$ are particularly advantageous, as they lead to high winding factors and minor mutual to phase inductance ratios.

- Even phase number 2-layer FSCWs are not suitable for fault-tolerant designs, as they do not simultaneously give rise to high main winding factors and mutual inductances as reduced as when the number of phases is odd.

- Lowest self-inductance and highest mutual inductance values are obtained for 4-layer FSCWs, making them inappropriate for fault-tolerant applications.

In order to be valid for even phase number FSCW machines as well, the rule to obtain a negligible mutual inductance between phases stated in [126,133], should be modified into:

\[
\begin{align*}
Q/(2mt) & \text{ must be even for 1-layer windings} \\
Q/(mt) & \text{ must be even for 2-layer windings}
\end{align*}
\]

(3.33)

3.2.3 Application example

Considering solely the slot and air-gap components of the inductances, the expressions for the calculation of the phase self-inductance and the mutual inductance between phases are:

\[
L_{ii} = L_{\text{slot},ii} + L_{\text{agap},ii} = 2\mu_0 l_{ef} \frac{N_s^2}{p} (\lambda_{\text{slot},ii} + \lambda_{\text{agap},ii})
\]

(3.34)

\[
L_{ij} = L_{\text{slot},ij} + L_{\text{agap},ij} = -2\mu_0 l_{ef} \frac{N_s^2}{p} (\lambda_{\text{slot},ij} + \lambda_{\text{agap},ij})
\]

(3.35)

Unfortunately, these expressions depend on the geometry of the machine, so when trying to evaluate the merits of a particular combination of phases, poles and slots, the effects of these variables are masked by the effect of the geometry. In order to decouple the effects of the geometry from the benefits of a certain combination, a number of different PMSMs have been analytically designed under the same design constraints.

The machines have been designed for an Integrated Starter Alternator (ISA) application, whose specifications are shown in Table 3.8.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated power, $P_{\text{nom}}$</td>
<td>2 [kW]</td>
</tr>
<tr>
<td>Corner speed, $N_{\text{nom}}$</td>
<td>600 [rpm]</td>
</tr>
<tr>
<td>DC bus voltage, $U_{\text{DC}}$</td>
<td>42 [V]</td>
</tr>
<tr>
<td>Outer diameter, $D_{\text{ext,s}}$</td>
<td>170 [mm]</td>
</tr>
<tr>
<td>Stacking factor, $k_F$</td>
<td>0.97</td>
</tr>
<tr>
<td>Air-gap thickness*, $g$</td>
<td>4 [mm]</td>
</tr>
<tr>
<td>Stack length, $l$</td>
<td>70 [mm]</td>
</tr>
<tr>
<td>Slot fill factor (Copper area/Slot area), $k_r$</td>
<td>0.40</td>
</tr>
<tr>
<td>Air-gap flux density, $B_{\text{agap}}$</td>
<td>0.8 [T]</td>
</tr>
<tr>
<td>Stator yoke flux density, $B_{\text{y,s}}$</td>
<td>1.4 [T]</td>
</tr>
<tr>
<td>Tooth flux density, $B_t$</td>
<td>1.4 [T]</td>
</tr>
<tr>
<td>Rated current density, $j_{\text{nom}}$</td>
<td>5 [A/mm$^2$]</td>
</tr>
<tr>
<td>Parallel paths</td>
<td>No</td>
</tr>
<tr>
<td>Skewing</td>
<td>No</td>
</tr>
</tbody>
</table>

* Accounting for surface mounted magnets
† No-load, peak value

Table 3.8: Specifications for ISA application machines
Other imposed constraints and/or assumptions are:

- Inner rotor, radial flux machine.
- Star-connection with no neutral line.
- Unity phase self-inductance, \( L_{ii} = 1 \text{ pu} \), calculated according to (3.34).
- Only combinations for which the winding factor is higher or equal to 0.90 have been considered. In the case of 4-phase machines, given the limited number of combinations that fulfilled such restriction, the criterion has been relaxed to \( k_w \geq 0.85 \).
- In order to reduce the presence of unbalanced radial forces, only combinations for which \( \text{gcd}(2p, Q) \geq 2 \) have been investigated [129].
- For constructional reasons, only those combinations for which \( \frac{1}{2} \leq 2p/Q \leq \frac{3}{2} \) have been studied.
- Surface mounted magnets. The rotor geometry is not considered and it is assumed that the radial air-gap flux density waveform created by the magnets is sinusoidal with a peak value equal to \( B_{agap} \).
- No reluctance torque is considered. The machine torque corresponds solely to the torque produced by the magnets.
- Constant width teeth. When calculating the slot component of the inductances, parameter \( b \) is equated to the mean slot width.
- The tooth-tip height equals half the tooth width. Thus, in case of a single phase short-circuit, half the magnet flux is bridged through the tooth head, avoiding excessive saturation [66].
- Infinite permeability of the ferromagnetic material.
- Skin effect is ignored in the calculations.

FSCW machines with all possible phase, pole and slot number combinations have been analytically designed in the range \( p \in [1, 30], m \in [3, 7] \) for 1 and 2 layers. Four-layer FSCWs have been disregarded as a consequence of the previous study. All designed machines possess a phase self-inductance of 1 pu. The comparison between the different combinations of phases, poles and slots is performed based on parameter \( \psi \) (3.17).

Figure 3.28 shows the results obtained for the 1-layer and 2-layer machines that met all the imposed restrictions. For the sake of clarity, the decimal logarithm of \( \psi \) is shown in the figure.

![Figure 3.28: \( \log_{10} \psi \) for ISA machine examples](image)

(a) 1-layer FSCWs. (\( \circ \)) Empty symbols: \( Q/(2mt) \) even. (\( \bullet \)) Full symbols: \( Q/(2mt) \) odd
(b) 2-layer FSCWs. (\( \circ \)) Empty symbols: \( Q/(mt) \) even. (\( \bullet \)) Full symbols: \( Q/(mt) \) odd

In observing Figure 3.28 it can be concluded that, for the given application:

- Single-layer windings give rise to lower inductances between phases than 2-layer FSCWS.
- On average, 1-layer windings for which \( Q/(2mt) \) is even have lower mutual inductance values than those for which \( Q/(2mt) \) is odd. Since combinations with a higher number of poles than slots have an inherently high inductance due to the rich content of MMF subharmonics, combinations for which
2p > Q have, on average, lower $\psi$ values than those with 2p < Q. The rule “$2p \approx Q$” is not as significant as the analysis conducted in the previous section indicated and, except for the “Q/(2mt) even” condition, there is no clear criterion for 1-layer windings to determine which phase, pole and slot combinations are particularly advantageous in order to achieve a reduced mutual inductance between phases.

- On the contrary, for 2-layer FSCWs, since the slot component plays a significant role in the total mutual inductance between phases, the “Q/(mt) even” rule is less significant than for 1-layer windings. Likewise, the effect of having 2p > Q in $\psi$ is reduced. In this case, the “$2p \approx Q$” rule becomes more important for odd phase number machines. The higher the phase number, the more stringent the aforementioned criterion is, as the relationship between the mutual and the self-inductance increases faster as the $2p/Q$ ratio shifts away from unity.

- As previously stated, 2-layer even phase FSCW machines are not suitable for fault-tolerant applications, since they do not simultaneously give rise to low mutual inductances between phases and high winding factors.

- For the specific application studied, the most promising combinations for each phase number in terms of the machine’s fault-tolerance are listed in Tables 3.9 and 3.10.

<table>
<thead>
<tr>
<th>Phases</th>
<th>Poles</th>
<th>Slots</th>
<th>$Q/(2mt)$</th>
<th>$2p &gt; Q$</th>
<th>$k_w$</th>
<th>$\log_{10} \psi$</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>14</td>
<td>12</td>
<td>even</td>
<td>yes</td>
<td>0.9659</td>
<td>-6.3624</td>
</tr>
<tr>
<td>4</td>
<td>36</td>
<td>32</td>
<td>even</td>
<td>yes</td>
<td>0.9061</td>
<td>-4.4274</td>
</tr>
<tr>
<td>5</td>
<td>22</td>
<td>20</td>
<td>even</td>
<td>yes</td>
<td>0.9877</td>
<td>-6.5956</td>
</tr>
<tr>
<td>6</td>
<td>26</td>
<td>24</td>
<td>even</td>
<td>yes</td>
<td>0.9077</td>
<td>-5.4164</td>
</tr>
<tr>
<td>7</td>
<td>34</td>
<td>28</td>
<td>even</td>
<td>yes</td>
<td>0.9439</td>
<td>-5.2914</td>
</tr>
</tbody>
</table>

Table 3.9: ISA 1-layer combinations with lowest $\psi$

<table>
<thead>
<tr>
<th>Phases</th>
<th>Poles</th>
<th>Slots</th>
<th>$Q/(mt)$</th>
<th>$2p &gt; Q$</th>
<th>$k_w$</th>
<th>$\log_{10} \psi$</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>58</td>
<td>60</td>
<td>even</td>
<td>no</td>
<td>0.9541</td>
<td>-1.3345</td>
</tr>
<tr>
<td>4</td>
<td>18</td>
<td>16</td>
<td>even</td>
<td>yes</td>
<td>0.8887</td>
<td>-0.5741</td>
</tr>
<tr>
<td>5</td>
<td>38</td>
<td>40</td>
<td>even</td>
<td>no</td>
<td>0.9827</td>
<td>-1.0920</td>
</tr>
<tr>
<td>6</td>
<td>46</td>
<td>54</td>
<td>odd</td>
<td>no</td>
<td>0.9297</td>
<td>-0.5007</td>
</tr>
<tr>
<td>7</td>
<td>58</td>
<td>56</td>
<td>even</td>
<td>yes</td>
<td>0.9802</td>
<td>-0.8987</td>
</tr>
</tbody>
</table>

Table 3.10: ISA 2-layer combinations with lowest $\psi$

### 3.2.4 FE validation

In order to validate the criteria obtained, the inductances of a number of candidate machines have been calculated by a FE analysis. The simulated machines have the characteristics listed in Table 3.11. The following additional restrictions have been applied to the most promising candidates in order to reduce the number of FEM simulations:

- Phase number between 4 and 6. This grants an enhanced post-fault control over a 3-phase machine, without excessively increasing the weight, cost and complexity of the associated inverter [69].

- Number of pole pairs between 10 and 20 to have an operating frequency in the range 100 – 200 Hz at base speed. At higher speeds, the increase in iron losses is assumed to be relieved by the field weakening operation.

- Maximum number of slots about 40 due to mechanical and economic constraints [135].

- Only 1-layer windings are considered for even phase number machines.

Interestingly, for all selected combinations, the relationship $2p = Q \pm 2t$ holds.

The results of the simulations are collected in Table 3.12, together with the results of the derived analytical expressions. The inductance values are split into their slot-leakage and air-gap components. The first
and second data column categories correspond to analytical and FEM results, both considering infinite
permeability of the stator material. Last data column category shows the phase self and mutual inductances
at rated current when a standard M600-50A lamination steel is considered as stator material [203].
Inductances by the FE method have been calculated using a magnetic energy approach (see Appendixes C
and D). For the calculation of mutual inductances, same current magnitudes with opposite sign are
introduced in phases $i$ and $j$, $i_j = -i_i$.

![Table 3.11: Most promising candidates for ISA application](image)

<table>
<thead>
<tr>
<th>Machine</th>
<th>Analytic</th>
<th>FEM (infinite permeability)</th>
<th>FEM at rated current (M600-50 lamination)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>slot</td>
<td>air-gap</td>
<td>sum</td>
</tr>
<tr>
<td>I</td>
<td>121.9</td>
<td>117.8</td>
<td>239.6</td>
</tr>
<tr>
<td>II</td>
<td>190.9</td>
<td>202.4</td>
<td>393.3</td>
</tr>
<tr>
<td>III</td>
<td>214.8</td>
<td>232.3</td>
<td>467.1</td>
</tr>
<tr>
<td>IV</td>
<td>266.5</td>
<td>222.6</td>
<td>489.1</td>
</tr>
</tbody>
</table>

Table 3.12: FE validation of designed machine inductances

Analytical and FE results considering infinite permeability show a fairly good agreement between the
two calculation methods and justify the simplifications undertaken in the derivation of the analytical
formulas. Relative errors in the calculation of the self-inductances are kept around 10%. Moreover, for
rated current and considering the M600-50A steel, FE results show that the designed machines have
a phase self-inductance close enough to the base inductance; meaning that the fault currents will be
effectively limited in case of a short-circuit fault. The main discrepancies arise regarding the mutual
inductances when the non-ideal iron material and 1-layer combinations are considered. The reason has
been explained in the previous section: while for the 2-layer machine each stator portion carries the flux
of a single phase, for the 1-layer machines the whole stator yoke carries the magnetic flux produced by
all the phases. Therefore, saturation of some parts of the stator yoke can increase the mutual coupling
between phases [126].

The influence of magnetic saturation on the inductances is shown in Figures 3.29a and 3.29b, where the
current and inductance magnitudes are expressed in the per-unit system. Current is expressed in terms
of rated current and the inductances are referred to the base inductance values (Table 3.11).

The figures indicate that for the sized 1-layer machines, saturation is slightly present at rated current
conditions. Referring to this point, it must be noted that the machines have been sized assuming infinite
permeability material, as the purpose of the design has been to allow a fast comparison between winding
topologies to be made. A proper machine design for any application should take magnetic saturation into
account, as heavy load conditions may result in significant differences between the analytically calculated
and the actual inductance values. In any case, results in Table 3.11 and Figures 3.29a and 3.29b reflect
the impact the winding topology has on the mutual inductances, regardless of magnetic saturation.

### 3.3 Conclusions

In this chapter, the main design principles for choosing the optimum number of phases, poles and slots for fault-tolerant PMSMs have been exposed. Namely, traditionally employed criteria in designing FSCWs have been reviewed and new rules to select the most promising combinations for fault-tolerant applications have been proposed. Moreover, two numerical factors that depend solely on the winding layout have been introduced (ς, ξ). These factors allow to evaluate the merits of a particular phase, pole and slot combination in terms of magnetic isolation regardless of the geometry of the machine.

In particular, 1-layer, 2-layer and 4-layer FSCWs with a phase number ranging from 3 to 7 have been studied. Normal winding systems and non-reduced winding systems have been considered for odd and even phase number machines, respectively (α_m = 2π/m).

The main conclusions of the study are as follows:

- Single-layer FSCWs following the rules “Q/(2mt) even” and “2p ≈ Q” lead to the lowest mutual to self inductance ratios. They additionally have the advantages of giving rise to non-overlapped coil windings, having high main winding factors, low cogging torque, etc. For surface-mounted permanent magnet machines, the reduction in the values of the mutual inductances that is obtained by choosing combinations with “2p ≈ Q” can be masked by the large effective air-gap.

- It is possible to design 2-layer winding machines with a low mutual coupling between phases when the number of phases is odd. In these cases, specially with surface-mounted permanent magnets, it is particularly important to select combinations for which “2p ≈ Q”, due to the high influence the slot leakage component wields over the total inductance.

- Even phase number 2-layer windings and 4-layer FSCWs are not suitable for fault-tolerant applications as they lead to significantly larger mutual coupling between phases than other winding configurations.

The above criteria have been validated by means of an application example and a finite element analysis. Additionally, the most promising machine candidates in terms of fault-tolerance for a given ISA application have been obtained.

It must be noted, that among the revised criteria for choosing FSCWs, some rules collide one with another and the final choice for the number of phases, poles and slots will be application dependent. A proper design should consider the following factors: radial forces, cogging torque and torque ripple, rotor losses, self-inductance and mutual inductance values, local saturation, etc. For fault-tolerant FSCW machines, the most promising combinations are those for which:

- The number of spokes per phase in the star of slots are even.
3.3. CONCLUSIONS

- The number of poles is close to the number of slots, so that non-overlapped coils can be used, the winding factor is high, the phase self-inductance is high and low mutual inductances between phases are obtained.

- Single-layer windings offer better isolation characteristics than 2-layer FSCWs, but the influence of the stator material saturation on the magnetic coupling between phases must be addressed.
Chapter 4

Design of a Fault-Tolerant PMSM Prototype

In this chapter the design of a fault-tolerant PMSM prototype is described. First, a design methodology for fault-tolerant PMSMs with a maximum isolation between phases is presented. Then, the required specifications and design constraints for the prototype are established. After the prototype has been designed, a Finite Element analysis and a thermal and hydraulic analysis are conducted in order to check the validity of the design. Next, some mechanical considerations regarding the manufacture of the machine are contemplated. Finally, the data for the prototype, together with its operating curves are presented.

4.1 Design methodology

A design methodology for fault-tolerant PMSMs has been developed. The general procedure is schematically represented in Figure 4.1. The methodology is based on [204–209].

First, the machine is sized according to the application specifications and required constraints. In this process, all the geometrical and winding parameters of the machine are obtained. Then, the performance of the machine in terms of electromagnetic behavior is evaluated. This serves as initial check for the sized machine. If the machine performance is deemed satisfactory, the design process continues. Otherwise, either the machine variables or the initial data for the sizing procedure are modified and the process is repeated until an adequate design is achieved.

It is important to notice that both in the sizing and evaluation procedures, an analytical formulation is used. The employed analytical formulas are well established in the related literature [131, 210–214]. One of the advantages of adopting analytical formulas for machine calculation is that a number of geometries can be sized and calculated in a short amount of time as compared to other methods based on machine discretization; such as the FEM or methods based on permeance networks [215]. From a reduced number of parameters, the complete machine geometry and performance characteristics are obtained almost instantaneously. A further advantage of the analytical methods is that parameter sensibility analysis can be very easily conducted, obtaining knowledge about the influence of the variation of a variable in a moment.

A drawback of the analytical methods is that they inherently follow a number of approximations and simplifications and do not give results as precise as the ones obtained with other methods, such as the FEM. Owing to this reason, after the machine has been sized and evaluated, a FE analysis is conducted in order to validate the soundness of the design. The FEA not only serves as a validation tool for the design and the employed analytical formulation, but it also helps optimize the electromagnetic behavior of the machine since more complex aspects can be taken into account.

After the FEA, a thermal and hydraulic analysis is conducted in order to check whether the temperatures in different parts of the machine are kept within valid boundaries. The most sensitive machine elements
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

Figure 4.1: Design methodology for fault-tolerant PMSMs
4.1. DESIGN METHODOLOGY

to the effect of high temperatures are the insulation covering the copper conductors and the permanent magnets. On the one hand, the motor insulation lifetime heavily depends on the temperature, with a rule of thumb stating that for every 10 °C increase in temperature, the insulation lifetime is reduced by 50% [15]. On the other hand, the PM magnetization curve varies widely with temperature, being the higher the temperature, the higher the risk of demagnetizing the magnets [216]. If the temperatures resulting from the thermal analysis are adequate and consistent with previous estimations, the design process continues. Otherwise, the machine performance figures are reevaluated and/or some design actions are taken in order to have suitable temperatures.

Regarding thermal and hydraulic evaluation, it is possible to couple both analysis with the analytical performance evaluation step [208, 209, 217]. This way, a reduction of the overall design process time is achieved. However, since the machine characteristics may change as a result of the FEA and losses are usually estimated more precisely, it is always convenient to repeat the thermal and hydraulic analysis after the electromagnetic design is concluded.

The next step in the design process involves conducting a mechanical analysis of the machine. In this stage, vibrations, noise level, machine constructability, etc. are checked. If the results from the mechanical analysis are satisfactory, the design process ends. Otherwise, either some of the machine variables or the initial data for the sizing procedure must be modified and the design process restarted.

It must be noted that the design process is eminently iterative due to the interactions that exist among different machine variables. For example, magnet and stator winding temperatures affect the flux linkage and stator resistance, respectively, and therefore, they must be taken into account when assessing the electromagnetic behavior of the machine. Likewise, if the radial forces in the air-gap are too high, a different winding topology or a larger air-gap may have to be considered in order to ensure the durability of the bearings. Figure 4.2 represents the basic interactions between the different aspects of machine design.

Owing to the many existing interactions, the modification of a number of variables is envisaged after each design step in order to reach a solution fulfilling all the imposed constraints. After the design procedure is completed, the final step should always be to build a machine prototype and test both the actual design and the correctness of the proposed design procedure.

### 4.1.1 Design inputs

In the sizing process, the whole machine geometry and winding data are obtained from a reduced number of basic parameters. The data inputs from which the design process is started are divided among:

- **Application specifications.** They are the variables set by the considered application. This can be the torque-speed characteristics, the voltage supply (mains or inverter), the type of service, the machine topology, maximum dimensions, etc.

- **Design constraints.** They are imposed by the available materials and construction methods, possible hardware and software limitations, etc.
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

- **Initial data.** Among the machine parameters, there are some that may be freely set by the designer within some reasonable boundaries. These parameters are usually chosen or estimated based on past designs or experience with similar characteristic machines. These parameters are necessary in a preliminary sizing of the machine, but can be later modified in subsequent steps. Examples include the air-gap length, flux density values, the electric loading, etc.

### Application specifications

The application specifications considered in the present thesis are:

- **Machine type** - Motor, generator or general purpose machine. It establishes whether the machine will be operating mostly in motoring mode, generating mode or in either conditions.
- **Voltage supply** - Mains or inverter supplied.
- **Machine topology** - Radial flux, axial flux or transversal flux machine.
- **Type of PM machine** - Stator PM machine or rotor PM machine.
- **Rotor type** - Surface-mounted magnets, surface-inset, interior permanent-magnets, etc.
- **Induced voltage waveform** - Sinusoidal (PMSM) or trapezoidal (BLDC).
- **Winding topology** - Selected phase, pole and slot combination and corresponding winding layout. In this case, the winding topology is considered to be a design specification since it is determined by the fault-tolerant requirement of the application.
- **Winding connections** - Star connection (wye), delta connection, etc. For multiphase machines, more connection options than for 3-phase machines exist. For instance, a 5-phase machine can be star, pentagon or pentacle connected.
- **Housing type and cooling method** - Totally enclosed, semi-open or open machine. Natural convection, forced ventilation, liquid cooling, etc.
- **N\textsubscript{nom}** - Corner or base speed [rpm]; that is, the speed beyond which the power is kept constant. Usually, this corresponds to the division between normal and flux-weakening operation, although it is not necessary.
- **N\textsubscript{max}** - Maximum speed [rpm].
- **T\textsubscript{nom}** - Rated torque [N·m]; that is, the torque the machine is able to deliver under continuous operation at base speed.
- **T\textsubscript{max}** - Maximum torque the machine is able to deliver during temporary service [N·m]. It is required that a maximum time period for such a temporary service is specified.
- **P\textsubscript{nom}** - Rated output power [kW]. The power the machine is able to deliver under continuous operation at base speed.
- **P\textsubscript{max}** - Maximum power the machine is able to deliver during temporary service [kW].
- **U\textsubscript{nom}** - Rated phase voltage (RMS) [V\_RMS]. If the machine is inverter supplied, the rated phase voltage should be related to the DC bus voltage, U\textsubscript{DC} [V].
- **I\textsubscript{nom}** - Rated phase current (RMS) [A\_RMS].
- **I\textsubscript{max}** - Maximum phase current during temporary service (RMS) [A\_RMS]. If the machine is inverter supplied, it will depend on the current-carrying capacity of the power semiconductor devices.
- **Geometrical constraints** - A number of geometrical constraints can be imposed depending on the application, such as a maximum outer diameter, a maximum machine length, a maximum rotor inertia, etc.
4.1. Design Methodology

Design Constraints

The design constraints imposed by the available materials and construction methods are, in the present case:

- $C_{Fe}$ - Employed iron material. It has been assumed to be a constraint for this thesis. It defines the magnetization curve, the electrical conductivity, the thermal conductivity, the density, etc.

- $C_{PM}$ - Employed permanent magnet material. It defines the magnet remanence, coercitivity, the relative permeability and other physical properties. The magnetic properties for the magnets are usually given at a certain temperature, $T_0$ (e.g. $T_0 = 20 \degree C$). It is commonly assumed that both the remanence and the intrinsic coercitivity values vary linearly with magnet temperature:

\[
B_{r,m} = B_{r,m0} \left(1 + \alpha_{B} (T - T_0)\right)
\]

\[
iH_c = iH_{c0} \left(1 + \alpha_{iH} (T - T_0)\right)
\]

(4.1)

where $\alpha_{B}$ and $\alpha_{iH}$ represent the magnet remanence and intrinsic coercitivity temperature coefficients, respectively.

- $k_{Fe}$ - Stacking factor or stack lamination factor. Takes into consideration the voids and insulation between laminations so that the “magnetic stack length” is always equal or less than the actual stack length.

- $k_f$ - Slot fill factor. Defined as the ratio between the copper area in a slot and the total slot area, disregarding the slot-opening and wedge sections. Depends on the design (copper bars or wounded stator), insulation requirements, construction method, etc.

- Geometrical constraints - A number of geometrical constraints can be imposed based on the available materials and manufacturing methods, such as a minimum slot-opening width, a minimum wedge-height, a maximum tooth height to yoke thickness ratio, etc.

Initial Data

The variables pre-established by the designer and that are needed for a first approximation to the design process are:

- $A$ - Electric loading or linear current density [A/m].

- $B_{agap}$ - Air-gap flux density due to the permanent magnets (peak value of the main harmonic) [T].

- $B_t$ - Stator tooth flux density (peak value) [T].

- $B_{y,r}$ - Rotor yoke flux density (peak value) [T].

- $B_{y,s}$ - Stator yoke flux density (peak value) [T].

- $g$ - Air-gap thickness [m].

- $j$ - Slot current density [A/m²].

- $\alpha_{PM}$ - Magnet arc width referred to the pole-pitch, $0 < \alpha_{PM} < 1$.

- $T_w$ - Winding temperature [°C].

- $T_{PM}$ - Permanent magnet temperature [°C].

The previously stated variables are referred to the nominal operation point of the machine; that is, for the machine operating continuously under rated torque and speed conditions. Initial values for these variables are usually taken from experience with similar characteristic machines [135,198]. As previously stated, the values for these variables can change during the design process in order to have a more sound design.
4.1.2 Analytical sizing process

The sizing process is the very first step of the machine design process. In this process, a first version of the machine geometry is obtained from some basic machine parameters. The machine sized this way must comply with the restrictions imposed and with the undertaken assumptions regarding the initial data; such as the specified winding and permanent magnet temperatures. In subsequent analysis, a number of geometrical, material, etc. variables can be modified in order to optimize the design and/or analyze the effect of such variables in the performance of the machine.

The flowchart for the analytical sizing process is shown in Figure 4.3. The sizing process has been adapted from [204,207–209] and thus, it will not be covered in detail. The interested author is invited to refer to the cited references.
4.1. DESIGN METHODOLOGY

First, from the main application specifications, a number of derived electrical specifications are obtained. Explicitly, these are the phase current \((I)\), the phase inductance \((L_{ii})\) and the flux linkage due to the permanent magnets \((\Lambda_m)\). These are calculated from the following equations:

\[
T = \frac{mp}{2} (\lambda_{di}q - \lambda_{q}i_d) \quad (4.2)
\]

\[
\begin{align*}
u_d &= R_{ph}i_d - \lambda_{q}\omega_e \\
u_q &= R_{ph}i_q + \lambda_{d}\omega_e
\end{align*} \quad (4.3)
\]

\[
I_{ch} = \frac{\Lambda_m}{L_d} = \frac{\lambda_m}{\sqrt{2}L_d} = I \quad (4.4)
\]

The first two equations impose the steady state electromagnetic torque and voltage values at rated conditions, while the third equation is related to the fault-tolerant behavior of the machine. Equation (4.3) ignores loss components other than copper losses. Under linear conditions:

\[
\begin{align*}
\lambda_d &= \lambda_m + L_d i_d \\
\lambda_q &= L_q i_q
\end{align*} \quad (4.5)
\]

For a surface-mounted permanent magnet (SPM) machine, the inductance values in both the \(d\)-axis and the \(q\)-axis are fairly similar and the following assumption can be made: \(L_{ii} = L_d = L_q\). For an interior permanent magnet (IPM) machine, a prior estimation of the saliency-ratio, \(L_d/L_q\) is needed. In any case, if the machine resistance is neglected in a first approximation, the previous equations set all the necessary conditions for the calculation of the phase current, the phase inductance and the flux linkage due to the permanent magnets. In subsequent calculations, more refined estimations of these parameters can be obtained.

Secondly, the needed rotor volume is calculated by relating the electromagnetic torque \((T)\) to the electric loading \((A)\) and the magnetic loading \((B_{agap})\) values [135, 198]. Since the machine length or the diameter to length ratio is usually specified, this directly yields the stack length \((l)\) and the air-gap diameter of the machine \((D_{agap})\). Once the air-gap diameter has been obtained, the air-gap flux due to the permanent magnets is easily obtained.

Having set the flux density values for the stator teeth \((B_t)\) and the stator yoke \((B_{y,s})\), the tooth width \((w_t)\) and the stator yoke thickness \((h_{y,s})\) can then be estimated. At this stage, the number of coil turns \((Z)\) in order to achieve the specified flux-linkage and voltage levels is calculated. The number of parallel conductors within the slot \((Z_{par})\) depends on the chosen slot current density \((j)\). Then, from the slot area needed to accommodate the total number of conductors in the slot, the slot height \((h)\) is obtained.

In fault-tolerant designs, the slot-opening or tooth-tip height \((h_0)\) is commonly chosen to be a significant fraction of the tooth width in order to avoid an excessive saturation of the tooth-tip region under a single phase short-circuit fault [66]. The slot-opening width \((b_0)\) is then selected in order to have a unity phase self-inductance.

At this point, the main dimensions for the machine stator have been set and the magnet thickness \((h_m)\) needed to obtain the desired air-gap flux density is calculated. This calculation can be carried out analytically [218, 219] or by constructing a magnetic circuit model. Among the magnetic circuit models many different possibilities exist with a varying level of complexity [215, 220]. Anyhow, once the magnet thickness is obtained, the computation of the rest of geometrical parameters and flux density values in the different machine regions is straightforward.

The final step is to check all the machine dimensions, electrical parameters and other imposed constraints. Usually, the first resulting machine of a sizing process does not meet all the imposed restrictions, since a number of calculations are based on initial estimations and guesses. The sizing process can then be refined by changing some of the machine dimensions and/or initial data. Once all the imposed constraints are met the sizing process ends and the design is moved onto a next step.
4.1.3 Analytical performance evaluation

Once the motor has been sized and its geometrical and winding parameters obtained, the performance characteristics of the machine are analytically calculated. The evaluation or calculation process can also be called after any of the subsequent machine design steps, since many of the physical and geometrical variables may have changed.

The flowchart for the analytical evaluation performance process is shown in Figure 4.4. Again, the evaluation process has been adapted from [204, 207–209].

![Figure 4.4: Performance evaluation flowchart](Image)

The analytical performance evaluation process starts with the geometric and winding data of the machine. Among the many geometrical and winding parameters, there are some that are defined as conducting parameters and are used to calculate the rest of variables. This are the air-gap diameter ($D_{agap}$), the air-gap thickness ($g$), the slot height ($h$), etc. Since at any time the performance evaluation procedure is called a variable may have changed, the very first step of the evaluation process is to recalculate the non-conducting geometrical and winding parameters.

After that, the flux density values in the different regions of the machine are obtained. As discussed earlier, this can be achieved via analytical formulas or by computing a machine equivalent magnetic...
4.1. DESIGN METHODOLOGY

circuit. This calculation relies on a first estimation of the machine current in order to address the effect of armature reaction. Once the magnetic field distribution in the machine is known, electrical parameters such as the flux-linkage, the inductances and the electrical resistance can be calculated. The electrical parameters together with the imposed load and speed conditions determine the electrical magnitudes of the machine.

The next step in the performance evaluation process is to estimate the power losses in different parts of the machine. A number of different analytical models are available in order to estimate power losses. The main part of the research effort has been centered in iron [221–223] and magnet losses [149,224,225]. Analytical expressions that estimate losses in the bearings can be found in [226].

Knowing the loss distribution in the machine allows to perform a thermal and hydraulic analysis. In the thermal analysis, the temperatures of different elements of the machine are calculated. The hydraulic analysis is used to estimate the convection coefficients and windage losses in the machine. As it has been previously mentioned, performing a thermal and hydraulic evaluation of the machine at this stage is optional. If the temperature evaluation is coupled with the electromagnetic performance evaluation, an overall reduction of the time required by the machine design process is expected [208, 209, 217]. In case analytical tools for thermal and hydraulic evaluation are not available, this step can be left for latter on.

Anyhow, after the analytical evaluation process has been completed, convergence for the machine current value (and temperature values, if available) should be checked. As long as there is a variation on the phase current magnitude, the magnetic field distribution in the machine should be recalculated in order to account for the new estimate of the armature reaction field. Once the algorithm reaches convergence, the performance evaluation process ends.

4.1.4 Electromagnetic Finite Element analysis

The Finite Element method (FEM) has become a standard technique for the electromagnetic analysis of electrical machines and many other physical systems [134]. It is a numerical method that relies on domain discretization in order to solve the partial differential equations that govern a certain process. In the case of electrical machines, this are Maxwell equations and the constitutive equations for the materials. Like any other numerical method, the FEM is a method that finds an approximate solution to the problem at hand, as the problem is reduced via discretization. Hence, having a proper meshing of the problem geometry is critical in order to obtain meaningful results.

Nowadays, a number of commercial software solutions for electromagnetic 2D and 3D FEA are available [227–230]. Numerical calculation programs do not usually have a specific purpose and allow to evaluate any system and geometry, with some imposed boundary conditions. Normally, they allow the coupling of the electromagnetic solution with other analysis tools, such as: electrical circuit simulation, thermal analysis, vibro-acoustic analysis and other multiphysics simulations.

The main advantage of the FEM over the analytical methods for electrical machine analysis is that more accurate results can be obtained since a number of complex phenomena can be taken into account, e.g: non-linear material behavior, magnetic field distribution change by eddy currents in the conductive materials, faulted elements, complicated geometries, etc. On the other hand, the time needed to build a machine model and the time needed for problem resolution are much greater in the FEM. Typically, FEA is used as a design validation and optimization tool for when an advanced design of the machine is available.

4.1.5 Thermal analysis

The goal of a thermal analysis is to determine the temperature distribution in the machine. As previously mentioned, the temperature of certain machine elements, such as the permanent magnets, conditions the behavior of the machine to a large extent. A magnet temperature increase means a lower remanence value and therefore, a lower phase flux linkage. The machine will need a higher supply current to deliver the required torque, with the consequent power loss increase and efficiency reduction. Moreover, too high
winding and magnet temperatures may risk the lifetime of such elements, so that machine temperature
evaluation becomes essential for a proper machine design.

The thermal evaluation methods for electrical machines are broadly divided in two categories [231]: alge-
braic methods and numerical analysis methods. Algebraic methods consist mainly in lumped-parameter
thermal networks [232]. The time needed to define all the heat flows inside the machine and build a model
can be large, but once constructed, solving the thermal model is almost instantaneous. On the other
hand, numerical methods, such as the Finite Element method or Computational Fluid Dynamics (CFD),
are general purpose methods, allow to model any geometry and obtain extremely precise results, but the
time needed to model and simulate a system can be huge. In the following, the main characteristics of
each method are briefly discussed [233].

**Lumped-parameter thermal networks**

Lumped-parameter thermal networks are one of the classical methods for thermal analysis in engineering.
The machine geometry is divided in a number of elements and the thermal interaction between such
elements is represented by a thermal resistance; there being an equivalence between thermal networks
and electrical circuits. Thermal resistances can be used to model any of the three possible thermal
exchange mechanisms: conduction, convection and radiation. In order to simplify machine analysis, a
number of assumptions are usually made:

- Heat flow in the axial, radial and circumferential directions are independent one from another.
- Heat flow in the circumferential direction is commonly neglected.
- A single temperature is used to describe the thermal behavior of each element.
- Power loss and heat storage are uniformly distributed along each element.

Generally speaking, this has been the preferred method for thermal analysis in electrical machines in
the past, although until recent times, quite simple thermal networks have been employed. The advance
in computing has favored the development of more complex machine thermal models. Moreover, the
advent of reference software Motor-CAD [234], which contemplates the study of many kinds of electrical
machines, has facilitated research and analysis in a variety of topologies. The complexity of modern day
thermal networks for electrical machine analysis can be appreciated in Figure 4.5.

![Figure 4.5: Lumped-parameter thermal network for a PMSM [234]](image-url)
4.1. DESIGN METHODOLOGY

One of the most important applications of the lumped-parameter thermal networks for electrical machines lies in the thermal analysis during the design stage. The high degree of parametrization of these models and their fast simulation makes them ideal candidates for a first analysis of the distribution of temperatures inside the machine.

Thermal Finite Element method

The Finite Element method can be applied to the thermal analysis of electrical machines. In this case, the Fourier law for heat conduction is applied to each mesh element:

\[ \alpha \nabla^2 T + \frac{\dot{q}_G}{\rho C_p} = \frac{\partial T}{\partial t} \quad (4.6) \]

where \( \alpha \) is the material’s thermal diffusivity, \( T \) the temperature, \( \dot{q}_G \) the specific heat generation and \( \rho \) and \( C_p \) the material’s density and specific heat capacity, respectively.

Although the FEM leads to very precise results regarding heat conduction in the solid materials of the machine, it needs additional information in order to account for heat convection and contact between different materials. Specifically, FEM based thermal analysis software often requires the introduction of the convection coefficients for the various surfaces as additional parameters. A number of commercially available software programs for this type of analysis exist [228,230,235,236].

Computational Fluid Dynamics

The set of numerical tools that allow to study the behavior of a fluid in a system is termed Computational Fluid Dynamics (CFD). In an electrical machine, they can be used to analyze the effects of ventilation, calculate the convection coefficients in different parts of the machine and analyze the thermal behavior of the machine. Like in the case of the thermal FEM, a number of programs to perform CFD calculations exist [230,237–239].

As it happens with the FEM programs, the degree of precision obtained with the CFD methods is very high. However, the time required to build a machine model and simulate a system is high, so they are mostly used in an advance stage of the machine design process or to study very specific phenomena, such as the heat convection in the end-windings of a machine [240].

4.1.6 Hydraulic analysis

A hydraulic analysis allows to determine the behavior of a fluid in an electrical machine. This can be the speed of air inside the machine or the hydraulic losses in case an external cooling circuit is employed. Since the temperature distribution in a machine heavily depends on the coolant fluid flows, the hydraulic and thermal analysis of a machine are often coupled together. Having a precise knowledge of the hydraulic behavior of the machine will allow to have a good estimate of the heat convection, the windage losses, the performance of possible fans, etc.

The preferred method for hydraulic analysis in electrical machines is to employ CFD programs. This has been used, for example, to obtain the speed-dependent convection curves in a machine [241], to study the heat convection in the end-windings [240], analyze the air flow in a self-ventilated traction machine [242], improve the thermal management of a synchronous generator [243] and design the air-cooling system for a PMSM [244].

An alternative to the CFD programs to study the air flow inside a machine is to use algebraic methods [245,246]. This is the method employed, for instance, by Motor-CAD [234] and other authors [247–250]. The method is analogous to the lumped-parameter thermal networks used in the thermal analysis of a system. A number of hydraulic resistances that model the behavior of a fluid inside the machine are defined. The models defined this way are easily parametrizable and their solving time is very low.
4.1.7 Mechanical analysis

The next step in the design process involves conducting a mechanical analysis of the machine. In this stage, forces, vibrations, noise level, machine constructability, etc. are checked. Usually this is accomplished with the help of Computer-Aided Design (CAD) software and additional functionalities for the analysis of structural mechanics, vibrations and acoustics. Again, the FEM is the preferred tool for carrying out this type of calculations [230,251,252].

If the results from the mechanical analysis are satisfactory, the design process ends. Otherwise, either some of the machine variables or the initial data for the sizing procedure must be modified and the design process restarted.

4.2 Required specifications and design constraints

In this section, the application specifications and design constraints chosen for the fault-tolerant PMSM prototype are described. The values of the initial pre-established variables, needed for a first approximation to the sizing process are presented as well.

The main constraint for the designed prototype has been that it had to be designed to fit in the housing of an old commercial asynchronous machine. Therefore, the prototype machine specifications and design restrictions have been chosen accordingly. Figure 4.6 shows the different elements of the disassembled asynchronous motor, while its main characteristics are collected in Table 4.1.

![Figure 4.6: Different elements of the original machine](image)

<table>
<thead>
<tr>
<th>Characteristics</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>4.0 [kW]</td>
</tr>
<tr>
<td>Torque</td>
<td>13.2 [N·m]</td>
</tr>
<tr>
<td>Voltage</td>
<td>415*[V_{RMS}]</td>
</tr>
<tr>
<td>Frequency</td>
<td>50 [Hz]</td>
</tr>
<tr>
<td>Number of phases</td>
<td>3</td>
</tr>
<tr>
<td>Number of stator slots</td>
<td>24</td>
</tr>
<tr>
<td>Number of pole pairs</td>
<td>1</td>
</tr>
<tr>
<td>Number of rotor slots</td>
<td>20</td>
</tr>
<tr>
<td>Housing length</td>
<td>200 [mm]</td>
</tr>
<tr>
<td>Stack length</td>
<td>120 [mm]</td>
</tr>
<tr>
<td>Total length (stack + end-wdg.)</td>
<td>192 [mm]</td>
</tr>
<tr>
<td>Stator bore diameter</td>
<td>88 [mm]</td>
</tr>
<tr>
<td>Slot height</td>
<td>12 [mm]</td>
</tr>
<tr>
<td>Stator yoke thickness</td>
<td>24 [mm]</td>
</tr>
<tr>
<td>Slot-opening width</td>
<td>2 [mm]</td>
</tr>
<tr>
<td>Slot bottom width</td>
<td>7 [mm]</td>
</tr>
<tr>
<td>Tooth width</td>
<td>5 [mm]</td>
</tr>
<tr>
<td>Lamination thickness</td>
<td>0.6 [mm]</td>
</tr>
<tr>
<td>Skewing</td>
<td></td>
</tr>
<tr>
<td>Rotor: 1 slot</td>
<td></td>
</tr>
<tr>
<td>Protection</td>
<td>IP54</td>
</tr>
<tr>
<td>Insulation</td>
<td>Class F (80° rise)</td>
</tr>
</tbody>
</table>

* △-connected

Table 4.1: Main characteristics of the original asynchronous machine
4.2. REQUIRED SPECIFICATIONS AND DESIGN CONSTRAINTS

4.2.1 Machine specifications

The machine specifications have been chosen so that they complied, as far as possible, with the ISA application requirements presented in Chapter 3, but taking into account the additional constraints imposed by the limited space of the housing to be used. The capacity of building a PMSM with similar power and speed ranges and comparable size has been demonstrated in a number of papers, such as the ones presented in Table 4.2.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Year</th>
<th>Power [kW]</th>
<th>Speed [rpm]</th>
<th>DC bus voltage [V]</th>
<th>Outer diameter [mm]</th>
<th>Stack length [mm]</th>
<th>Rotor type</th>
<th>Application</th>
</tr>
</thead>
<tbody>
<tr>
<td>[253]</td>
<td>2002</td>
<td>2.35</td>
<td>(750/-)</td>
<td>288</td>
<td>275</td>
<td>130</td>
<td>SPM</td>
<td>In-wheel</td>
</tr>
<tr>
<td>[88, 89]</td>
<td>2005</td>
<td>0.75</td>
<td>(900/3000)</td>
<td>42</td>
<td>120</td>
<td>40</td>
<td>IPM</td>
<td>Steering</td>
</tr>
<tr>
<td>[144, 145]</td>
<td>2005</td>
<td>4.0-6.0</td>
<td>(600/6000)</td>
<td>42</td>
<td>272</td>
<td>60</td>
<td>SPM</td>
<td>ISA</td>
</tr>
<tr>
<td>[254]</td>
<td>2008</td>
<td>0.55</td>
<td>(600/6000)</td>
<td>42</td>
<td>130</td>
<td>55</td>
<td>IPM</td>
<td>ISA</td>
</tr>
<tr>
<td>[255]</td>
<td>2010</td>
<td>1.0</td>
<td>(300/3000)</td>
<td>250</td>
<td>140</td>
<td>75</td>
<td>IPM</td>
<td>ISA</td>
</tr>
<tr>
<td>[256]</td>
<td>2012</td>
<td>5.0</td>
<td>(1350/5050)</td>
<td>120</td>
<td>150</td>
<td>135</td>
<td>SPM</td>
<td>Traction</td>
</tr>
<tr>
<td>[257]</td>
<td>2013</td>
<td>0.8</td>
<td>(570/4000)</td>
<td>311</td>
<td>130</td>
<td>80</td>
<td>IPM</td>
<td>Traction</td>
</tr>
<tr>
<td>[257]</td>
<td>2013</td>
<td>5.0</td>
<td>(570/4000)</td>
<td>311</td>
<td>260</td>
<td>80</td>
<td>IPM</td>
<td>Traction</td>
</tr>
</tbody>
</table>

Table 4.2: Literature examples of EV related machines

Machine topology

A radial-flux, inner rotor, surface mounted permanent magnet (SPM) machine has been chosen as machine type. The main reason for such a choice has been the easier manufacture of a SPM machine over a IPM machine [258].

It is true that an IPM design offers some advantages, such as a more robust rotor design, an increased air-gap flux density capability through flux focusing, improved flux-weakening characteristics and an additional torque component due to saliency [258, 259]. Additionally, the magnets are better protected against demagnetization when an IPM design is used [260]. However, IPM rotor designs suffer from an increased leakage flux and are more difficultly magnetized and manufactured [261]. It is also argued in [159] that the reluctance torque in an IPM machine is relatively low or even negligible when a FSCW is employed. Furthermore, rotor losses as well as stator iron losses may be increased when the magnets are placed inside the rotor [159].

In the last decades, a new family of PMSMs with the magnets placed on the stator has emerged, such as the doubly salient PM machine [262–264], the flux reversal PM machine [265] and the flux switching PM machine [266]. Although these machines offer some advantages over the rotor PM machines, such as a robust rotor structure, reduced cost and improved fault-tolerance [64], their torque density is somewhat lower, the stator manufacture is more complex and their use is not as widespread [258].

Taking into account the previous considerations, it has been decided to design a traditional radial-flux, inner rotor, SPM machine.

Induced voltage waveform

AC permanent magnet machines are broadly classified among PMSM and BLDC machines. Permanent magnet synchronous machines are designed so that the induced voltage waveform is as sinusoidal as possible and are supplied with sinusoidal currents. Brushless DC machines, on the contrary, produce a trapezoidal induced voltage waveform and are supplied with quasi-rectangular currents.

A BLDC machine offers the advantages of a simple control and an enhanced torque density over a PMSM, but the torque ripple, noise and vibrations are higher. Furthermore, the controllability of a BLDC machine is poorer, specially in the flux weakening region [260]. It has been demonstrated in [158, 267] that a multiphase machine with the induced voltage composed of a first and a third harmonic offers the advantages of both the PMSM and BLDC machines. Specifically, a 5-phase machine is considered in the above references.
However, since the purpose of this thesis has been focused since the beginning on PMSM machines, it has been decided to design a machine with a sinusoidal induced voltage waveform.

**Speed**

A speed of 600 rpm has been selected as base speed for the application. It can be noticed that this speed is the same as for the machines sized in Chapter 3. It is also equal to the base speed of some of the machines presented in Table 4.2.

It has been shown in [144, 268] that having a characteristic current equal to the rated current grants an ideally infinite flux weakening range to any PM machine. This is the case of the prototype machine, as it has been designed to have a per-unit self-inductance. Therefore, the maximum speed of the machine will be mainly limited by machine losses and by the ability of the inverter to generate an adequate output voltage waveform for a set switching frequency. Although flux weakening will alleviate up to a point the increase in iron losses as the speed increases, both the iron and magnet losses will likely increase with the speed. Also, too high a d-axis current may demagnetize the magnets, so it can be an additional limiting factor for the machine’s maximum speed.

Since the aim of the prototype machine has been to demonstrate the fault-tolerant capabilities of a permanent magnet machine, little attention has been paid to the maximum speed attainable by the machine. This aspect will be commented latter, when analyzing the results of the design process.

**Winding topology**

The designed prototype is a 5-phase, 22-pole, 20-slot, 1-layer machine ($m = 5, p = 11, Q = 20, n_l = 1$). This corresponds to the most promising combination in terms of mutual to self-inductance ratio among the machines sized in Chapter 3. This very same combination is studied for instance in [137, 168, 269–271].

Having eleven pole pairs and a base speed of 600 rpm implies having a fundamental electric frequency of 110 Hz at rated conditions. This frequency can be considered to be moderate for this type of machine, so special attention has to be paid to stator iron and rotor losses.

The machine’s diameter is large enough to accommodate the selected number of poles and slots, so no construction problems are expected regarding the slot and pole dimensions.

The winding layout for the considered combination and the corresponding star of slots is represented in Figure 4.7. The chosen winding has to the following properties:

- Electrical periodicity $t = 1$.
- Coil throw $y_q = 1$ (non-overlapped coils).
- Winding factor for the main harmonic $k_{w,p} = 0.9877$. 

Figure 4.7: Winding layout and corresponding star of slots for the selected combination ($m = 5, p = 11, Q = 20, n_l = 1$)
4.2. REQUIRED SPECIFICATIONS AND DESIGN CONSTRAINTS

- Greatest common divisor of the number of poles and slots \( \text{gcd} (2p, Q) = 2 \). The winding layout has an antisymmetric distribution of the coils in the slots, so the radial forces in the air-gap are compensated and no net radial force is generated. The lowest order of the radial forces in the stator is 2.

- The number of cogging torque periods per slot-pitch is \( N_p = \frac{2p}{\text{gcd}(2p, Q)} = 11 \), so a high frequency and very low amplitude cogging torque is expected.

- Figure 4.8 shows the ideal air-gap armature MMF waveform and its harmonics. As it can be appreciated in the figures, there is a strong first order subharmonic that has to be addressed in order to limit permanent magnet losses.

![Figure 4.8](image)

- The mutual inductance to phase self-inductance ratio coefficients for the slot and the air-gap are \( \varsigma = 0 \) and \( \xi = 2.594 \cdot 10^{-8} \), respectively, so a negligible mutual inductance between phases is expected at linear conditions. The magnetic loading in the stator parts will to be carefully checked in order to avoid a magnetic coupling between phases due to material saturation.

Electromagnetic torque

The machine has been designed for an electromagnetic torque at rated conditions of 27.06 N·m \((P = 1.7 \text{ kW})\). This is about twice the torque density of the original asynchronous machine. The reasons why such a large increase in the torque density is considered possible, are:

- Lower rotor losses in a PM machine than in a asynchronous machine. If appropriate measures are taken in order to limit the effect of the armature MMF harmonics, rotor losses can be much lower than in the original machine.

- Shorter end-windings due to the winding topology. Shorter end-windings means lower copper losses than for the same size and torque integral-slot winding machine.

- Having a large number of poles, the stator yoke thickness will be smaller than in the original machine and thus, heat transfer to the outside of the housing will be improved.

Ultimately, the machine torque will be limited by the cooling capacity of the machine [135], so a thermal analysis will have to be conducted in order to check the ability of the machine to deliver the considered torque.

Voltage

The available DC bus voltage for testing the prototype is 300 V. Considering a sinusoidal-PWM strategy supply and assuming a 5% voltage drop in the inverter semiconductor switches, a maximum voltage of around 100 \( V_{RMS} \) will be available for machine supply \((U_{\text{max}} = 0.95U_{DC}/\sqrt{8} \approx 100 \text{ V})\). Hence, a rated phase voltage of 100 \( V_{RMS} \) has been considered for the prototype.
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

Geometrical constraints

Since the prototype machine had to fit in the existing housing, it has been designed with the same stack dimensions that the original asynchronous machine. Specifically, a stator outer diameter of 166 mm and a stack length of 120 mm have been imposed as geometrical constraints. The outer diameter has been increased by 6 mm to allow the machining of the inner diameter of the housing and ensure a proper fitting of the stack inside the housing.

Actually, the stack length could have been increased due to the reduction in the end-winding volume and a higher power machine could have been built. However, in order to ease the exit of the phase cables and allow for a proper comparison, it has been decided to keep the machine stack length unchanged. Another reason for doing so has been to leave large enough clearances between the stack and the housing so that no additional losses in the frame and the end-shields were generated due to the special winding configuration. As demonstrated in [168], stator end-effects can be significant in FSCW machines if no proper measures are taken. In the present thesis, due to the high clearance values, eddy current losses in the housing have been disregarded.

4.2.2 Design constraints

Iron material

A standard non-grain-oriented electrical steel lamination has been selected as iron material for both the stator and the rotor. The electrical steel material, a M600-50A lamination [272], has been chosen among the available materials by the supplier. From the magnetic viewpoint, the different material grades are characterized by their permeability and power loss values. The higher the alloy content of the material, the higher the electrical resistivity and thus, the lower the losses. However, an increased alloy content also results in a higher cost, reduced permeability and lower saturation threshold. The M600-50A lamination has a good balance in terms of cost, permeability and power losses. According to the manufacturer it is suitable for economical applications and medium sized electrical machines.

Generally speaking, having a high permeability value is more important for small machines like the considered one in order to reduce machine volume. However, taking into account the moderately high frequency of the application, a medium permeability and hence, medium power loss material grade has been chosen. It has been decided to manufacture both the stator and the rotor with the same lamination material in order to reduce the rotor losses and make the best use of the steel laminations.

The magnetization curve and power loss density versus supply frequency and polarization are represented in Figure 4.9.

![Magnetization curve and power loss density](image)

Figure 4.9: Magnetization curve and power loss density for the M600-50A lamination [272]
4.2. REQUIRED SPECIFICATIONS AND DESIGN CONSTRAINTS

PM material

Regarding PM material, N44SH grade NdFeB magnets have been selected [273]. The magnet properties are presented in Table 4.3.

<table>
<thead>
<tr>
<th>Property</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnet remanence*, ( B_{r,m} )</td>
<td>1.37 [T]</td>
</tr>
<tr>
<td>Intrinsic coercitivity*, ( H_c )</td>
<td>1.600 [kA/m]</td>
</tr>
<tr>
<td>Max. energy product*, ( (BH)_{max} )</td>
<td>358 [kJ/m³]</td>
</tr>
<tr>
<td>Remanence temp. coefficient, ( \alpha_{B_r} )</td>
<td>-0.0012 [1/°C]</td>
</tr>
<tr>
<td>Intrinsic coercitivity temp. coeff., ( \alpha_{H_c} )</td>
<td>-0.0055 [1/°C]</td>
</tr>
<tr>
<td>Coercitivity*, ( H_c )</td>
<td>≥ 963 [kA/m]</td>
</tr>
<tr>
<td>Magnet relative permeability, ( \mu_{r,m} )</td>
<td>1.054</td>
</tr>
<tr>
<td>Max. working temp., ( T_{PM,max} )</td>
<td>150 [°C]</td>
</tr>
<tr>
<td>Density</td>
<td>7.400 [kg/m³]</td>
</tr>
<tr>
<td>Electrical resistivity</td>
<td>1.6 [µΩ·m]</td>
</tr>
</tbody>
</table>

* At \( T_{PM} = 22 \) °C

Table 4.3: Magnet characteristics - NdFeB N44SH [273]

NdFeB magnets have been selected since they give rise to the highest remanence and \((BH)\) energy product among the different magnet materials. SmCo magnets have better temperature characteristics but are more expensive and have a lower energy density. Ferrite magnets are cheaper, but their magnetic characteristics are poor. AlNiCo magnets have a high remanence but a low coercitive and are more suitable for other motor applications, such as memory motors [274,275].

High temperature grade magnets have been selected in order to guarantee magnet safety even under heavy load and fault conditions. Magnet material cost has not been an issue for the considered design.

The magnetization curves for the selected material at different temperatures are shown in Figure 4.10. Assuming that the magnet temperature will always be lower than 150 °C, it can be considered that the magnets will not demagnetize as long as the magnet flux density is higher than 0.6 T.

![Figure 4.10: Magnetization curve for NdFeB N44SH magnets [273]: magnetic flux density (solid line), magnetic polarization (dashed line)](image)

Geometric factors

A stack lamination factor of 0.96 and a slot fill factor of 0.45 have been contemplated. Both can be considered to have typical values for machines this size and power. Due to the special coil configuration, with just a single phase coil inside each slot and having the coils spanning a single slot-pitch, it is expected that the slot fill factor could be increased. Furthermore, the designed prototype is a low-voltage machine and therefore, the electrical insulation requirements are not excessive. Hence, it is thought that a rather conservative slot fill factor has been chosen. This issue will be checked after prototype manufacture.

Other geometrical constraints

In order to avoid an excessive saturation of the tooth-tip region under a single phase short-circuit fault, the tooth-tip height has been designed to be equal to half the tooth width. Hence, in case of a fault, half the magnet flux is bridged through the tooth head [66].
4.2.3 Initial data

Electric loading

One of the main aspects to determine the size of a machine is its electric loading. In [135] the following variation ranges are given for the linear current density, \( A \), and the slot current density, \( j \), for well-designed PMSMs:

<table>
<thead>
<tr>
<th>Linear current density, ( A )</th>
<th>( 35 - 65 ) [kA/m]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current density, ( j )</td>
<td>( 4 - 6.5 ) [A/mm(^2)]</td>
</tr>
<tr>
<td>Thermal product, ( A \cdot j )</td>
<td>( 14 - 42.25 ) ([10^4\text{A}^2\text{m}^3])</td>
</tr>
</tbody>
</table>

Table 4.4: Recommended electric loading values for PMSMs [135]

The linear current density \( A \) tends to be higher for large machines, while small machines tolerate higher slot current densities \( j \) than large machines [135]. The product of \( A \) and \( j \), sometimes termed thermal product \( A \cdot j \), is more or less independent of the machine size. The permitted electric loading values are defined on the basis of the thermal insulation class and the cooling of the machine. The higher the thermal class of the insulation materials and the better the cooling, the higher the electric loading of a machine can be. Using a low loss iron material also helps to increase the electric loading figures.

Since the designed prototype is a totally enclosed machine that relies on natural convection as a cooling method, the values corresponding to the lower limits of the varying ranges in Table 4.4 have been selected for an initial calculation: \( A = 35 \) kA/m, \( j = 4 \) A/mm\(^2\).

Magnetic loading

The magnetic loading values in a PMSM machine are mainly limited by the iron losses and the low remanence and coercitivity values of present day magnets. Table 4.5 collects recommended flux density values for PMSMs [135]:

<table>
<thead>
<tr>
<th>Air-gap flux density, ( B_{\text{gap}} )</th>
<th>( 0.80 - 1.05 ) [T]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator tooth flux density, ( B_t )</td>
<td>( 1.0 - 1.5 ) [T]</td>
</tr>
<tr>
<td>Stator yoke flux density, ( B_{y,s} )</td>
<td>( 1.5 - 2.0 ) [T]</td>
</tr>
<tr>
<td>Rotor yoke flux density, ( B_{y,r} )</td>
<td>( 1.0 - 1.6 ) [T]</td>
</tr>
</tbody>
</table>

Table 4.5: Recommended magnetic loading values for PMSMs [135]

Owing to the moderately high electric frequency of the application, low flux density values have been chosen for the air-gap and the stator teeth in an attempt to limit iron losses: \( B_{\text{gap}} = 0.85 \) T, \( B_t = 1.5 \) T. Flux density values in the stator and rotor yokes have not been determined by machine loadability criteria, but rather by mechanical considerations. Since the number of poles in the designed prototype is high, the required stator and yoke thickness in terms of magnetic loading will be low. However, such small yoke thickness values would imply constructional problems (insufficient structural stiffness, opening of the stack in a fan-like way, etc.). A minimum stator yoke thickness to slot height ratio of \( h_{y,s}/h = 0.25 \) has been imposed. In the case of the rotor yoke, its thickness is determined by the rotor outer lamination diameter and the shaft diameter.

An additional consideration to select low flux density values in the iron parts of the machine has been to avoid an undesired magnetic coupling between the phases and an unbalanced saturation due to the armature MMF subharmonics [126].

Air-gap thickness

The air-gap thickness of an electrical machine is one of its most relevant parameters. In a PMSM, having a small air-gap helps saving magnet material, but at the same time, increases iron losses due to slot permeance and armature MMF harmonics. Despite the great importance of the air-gap thickness, no
theoretical optimum value has been found for it. In a PMSM, the determination of the air-gap and
magnet thickness is a demanding optimization task [135].

Some empirical rules to choose the air-gap thickness of an electrical machine are proposed in [135]. A
first rule relates the air-gap thickness with the machine’s power:

\[
g = \begin{cases} 
10^{-3} \left(0.2 + 0.01P^{0.4}\right) & \text{when } p = 1 \\
10^{-3} \left(0.18 + 0.006P^{0.4}\right) & \text{when } p > 1 
\end{cases} \quad (4.7)
\]

If the machine is to be supplied with an inverter drive, it is intended for extremely heavy duty or high
pulsation losses are expected (as in large machines with open slots), the value given by equation (4.7)
should be increased by 60 – 100% to reduce rotor losses.

A second rule relates the air-gap thickness with the maximum permitted armature reaction. In order
to ensure that the flux density on one side of a magnetic pole is not reduced excessively, the following
relationship should hold true [135]:

\[
g \geq \gamma \tau_p \frac{A}{B_{agap}} \quad (4.8)
\]

where \(\tau_p\) is the pole pitch and \(\gamma\) a geometrical coefficient that is around \(\gamma = 7 \cdot 10^{-7}\) for constant air-gap
salient-pole synchronous machines. For the design at hand, equation (4.7) yields a value of \(g = 0.48\) mm
and equation (4.8), \(g \geq 0.36\) mm, assuming a stator bore diameter of 88 mm (the same value of the
original asynchronous machine).

However, additional considerations should be taken into account. First of all, since the designed machine
is just a prototype, no proper magnetizing equipment was available and pre-magnetized magnets have
been used for rotor manufacture. In order to ease the placement of the rotor inside the stator and cover
for possible eccentricities, the air-gap has been increased significantly. Another reason to increment the
air-gap thickness has been to limit rotor losses due to the high harmonic and subharmonic content of the
armature MMF waveform due to the special winding topology.

An air-gap thickness of 1.5 mm has been finally selected. This is a very conservative value substantiated
mainly in avoiding problems when placing the rotor inside the stator. If the designed machine was series
produced, the air-gap thickness should be definitely decreased in order to reduce magnet volume and
hence, machine costs.

**Magnet arc**

Assume that for a magnet arc width of \(\alpha_{PM}\) (\(\alpha_{PM} = \alpha_{PM}\tau_p\)), the air-gap flux density waveform due
to the permanent magnets is rectangular as displayed in Figure 4.11. This corresponds to an ideal case
where both the stator and rotor materials are infinitely permeable, stator slotting effect is ignored and
the air-gap thickness is negligible. The RMS value of such a waveform, \(Y\), and the RMS value for the
first harmonic, \(Y_1\), are, respectively:

\[
\begin{cases} 
Y = \sqrt{\alpha_{PM}} \\
Y_1 = \frac{2\sqrt{2}}{\pi} \sin \left(\frac{\alpha_{PM}\pi}{2}\right) 
\end{cases} \quad (4.9)
\]

and the Total Harmonic Distortion (THD):

\[
THD_y = \frac{\sqrt{Y^2 - Y_1^2}}{Y_1} \quad (4.10)
\]

The THD of the rectangular waveform attains its minimum value for \(\alpha_{PM} = 0.7420\). In reality, the large
effective air-gap of the machine will cause the air-gap flux density waveform to be much less rectangular
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

Figure 4.11: Rectangular air-gap flux density waveform

and the minimum THD of the actual waveform will be achieved for a lower $\alpha_{PM}$. Since the prototype machine is intended to work as a PMSM, the induced voltage waveform should be as sinusoidal as possible and therefore, a value of $\alpha_{PM}$ equal to 0.70 has been considered in initial calculations. This parameter has been objective to subsequent optimization with the aid of a FEA.

Temperatures

Having an initial guess of the steady state magnet and winding temperatures is necessary in order to proceed with the sizing process of the machine, as they will affect all the electrical variables of the machine due to their influence on the flux linkage and the phase resistance. The authors in [135] recommend not exceeding a temperature of 100 $^\circ$C for NdFeB magnets to prevent magnet demagnetization due to demagnetizing armature reaction fields.

Taking into account that low electric and magnetic loading values have been considered for the design, a magnet and winding temperature of 100 $^\circ$C has been estimated (that is, a temperature increase of 75 $^\circ$C for an ambient temperature of 25 $^\circ$C).

4.3 Finite Element analysis

With the application specifications and design constraints chosen in the previous section, a machine has been analytically sized and evaluated. In order to verify the adequateness of the design and optimize its behavior, a number of FE simulations have been conducted. The simulations have been carried out with the electromagnetic FEM software Flux by CEDRAT [228].

When analyzing an electrical machine by the FEM, some simplifications are usually made to keep the complexity of the problem on an adequate scale. These simplifications are mainly related to the lamination material and the resulting magnetic field distribution. The laminations used in the manufacture of electrical machines, usually have an anisotropic, non-linear and hysteretic behavior. Material properties depend on physical magnitudes, such as material temperature and supply frequency, and therefore, the magnetization curve for the material can be very complex [276].

The first usual simplification is to neglect the hysteresis phenomenon and properties dependency on temperature and frequency. Materials are represented by a unique $B(H)$ relationship, corresponding to the static magnetization curve. This is justified by the low coercitivity values of the commonly used electrical steel and the small variations in the magnetic field distribution with frequency and temperature [272]. The second simplification is to ignore material anisotropy due to the material’s crystalline structure and stack heterogeneity. A third simplification is to disregard the eddy currents induced in the lamination material when calculating the magnetic field distribution. This is justified by the high resistivity values of the commonly employed silicon steels and by the small thickness of the laminations. Owing to the small diameter of the selected copper conductors, skin effect in the conductors has been neglected as well. The last simplification contemplated in the present thesis is to ignore end-effects in the machine and consider the system as a 2D one. This is a common assumption in radial flux machines. End-winding leakage
4.3. FINITE ELEMENT ANALYSIS

Inductance has been analytically calculated and added to the electrical circuit equations coupled to the FE model [214].

Regarding hysteresis and eddy currents phenomena, these are addressed when calculating the iron losses in the machine. This is usually accomplished by applying a loss model, such as the Steinmetz equations or a mathematical hysteresis model [223]. It is important to point out that, in the present thesis, loss components other than copper losses, $P_{Cu}$, are subtracted from the electromagnetic power, $P$, in order to compute the shaft output power, $P_{shaft}$:

$$
\begin{cases}
    P_{in} = P + P_{Cu} \\
    P_{\text{loss}} = P_{\text{loss}} - P_{Cu} = P_{Fe,t} + P_{Fe,y,s} + P_{Fe,y,r} + P_{PM} + P_{wind} + P_{bear} + \ldots \\
    P_{\text{shaft}} = P - P_{\text{loss}}
\end{cases}
$$

This convention has been adopted in order to have consistent results with the electromagnetic FEM software employed. In Flux [228], iron losses are not considered during FE calculations, but they are calculated as a post-processing result.

Losses in the machine have been calculated in two different ways. For the rated speed and torque conditions, a transient FE simulation has been carried out. From this simulation, losses in the magnets have been calculated considering the magnets as solid conductors and losses in the iron parts have been obtained using the modified Bertotti formulas [223]. In this method, iron losses are separated in three terms: static hysteresis losses, dynamic eddy current losses and excess losses, and the following loss model is applied to each mesh element:

$$
P_{Fe} = V (p_h \hat{B}^2 f + p_{ec} \left< \left( \frac{dB}{dt} \right)^2 \right> + p_{exc} \left< \left( \frac{dB}{dt} \right)^{1.5} \right>)
$$

where $V$ is the element volume, $\hat{B}$ the peak flux density value in a period, and $k_h$, $k_{ec}$ and $k_{exc}$ are loss coefficients that are determined by curve fitting the previous model to the measured loss curve of the material.

The second method has been applied in the calculation of losses at different speeds. This second approach relies on the calculation of the values of:

$$
\hat{B}, \left< \left( \frac{dB}{dt} \right)^2 \right> \text{ and } \left< \left( \frac{dB}{dt} \right)^{1.5} \right>
$$

for each stator teeth and yoke section for a certain load condition ($i_d$, $i_q$). These values are obtained via magnetostatic FE simulations and post-processed in the MATLAB software environment [277]. Each stator teeth and yoke section is considered as a whole and rotor losses are ignored. The comparison between both methods is given in Table 4.6 for rated conditions. In order to have a broader comparative, iron losses calculated with the Loss Surface model [278] are presented as well.

<table>
<thead>
<tr>
<th>Loss model</th>
<th>Method 1</th>
<th>Method 2</th>
<th>Loss Surface</th>
</tr>
</thead>
<tbody>
<tr>
<td>FE simulation</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Stator teeth loss, $P_{Fe,t}$</td>
<td>30.69 [W]</td>
<td>34.70 [W]</td>
<td>32.31 [W]</td>
</tr>
<tr>
<td>Stator yoke loss, $P_{Fe,y,s}$</td>
<td>10.71 [W]</td>
<td>8.00 [W]</td>
<td>9.64 [W]</td>
</tr>
<tr>
<td>Rotor iron loss, $P_{Fe,y,r}$</td>
<td>0.36 [W]</td>
<td>– [W]</td>
<td>0.23 [W]</td>
</tr>
<tr>
<td>Magnet loss, $P_{PM}$</td>
<td>0.61 [W]</td>
<td>– [W]</td>
<td>0.61 [W]</td>
</tr>
<tr>
<td>Total, $P_{loss}$</td>
<td>42.37 [W]</td>
<td>42.70 [W]</td>
<td>42.79 [W]</td>
</tr>
</tbody>
</table>

Table 4.6: Comparison between loss calculation methods for rated conditions (assuming $T_w = 100 \, ^\circ C$, $T_{PM} = 100 \, ^\circ C$)

The comparative shows that both Bertotti and Loss Surface models give rise to very similar results, which leads to a improved confidence in the used calculation methods. Moreover, even though the second
method ignores rotor losses, these are fairly low and iron loss distributions in the stator are consistent. Since the data needed to apply the second method must be calculated just once regardless of machine speed conditions, iron loss calculation with the second method is much faster and thus, this has been the method employed when calculating machine losses at different speeds. In the future, analytical methods to include rotor iron and magnet losses as well should be implemented, e.g. [149, 224, 225].

In the following, the main conducted FE simulations are introduced. The rest of FEA results are presented on a latter section, where all the relevant data regarding the final prototype design is shown.

### 4.3.1 Magnet arc width

In the first place, the effect of the magnet arc width in the induced voltage waveform has been studied. The aim of the analysis has been to diminish the THD of the back-EMF waveform. Figure 4.12 shows the simulated induced voltage THD versus varying values of the relative magnet width, $\alpha_{PM}$. It can be appreciated that the minimum value of the THD is achieved for a lower $\alpha_{PM}$ than the initially considered one ($\alpha_{PM} = 0.7$). On the basis of the conducted analysis, a magnet arc width of $\alpha_{PM} = 0.66$ has been considered in consequent designs.

![Figure 4.12: Induced voltage THD vs magnet arc width](image)

### 4.3.2 Slot-opening width

A second FEA has been conducted in order to guarantee that a unity phase self-inductance is achieved. In a SPM equipped with FSCWs, the phase inductance can be varied most efficiently by modifying the geometry of the tooth tip. Having a slot-opening height equal to half the tooth width, the simplest way of varying the phase inductance is by modifying the slot-opening width, $b_0$. Figure 4.13 presents the results obtained analytically by using the formulas derived in this thesis (see Appendixes C and D) together with those obtained by the FEM.

![Figure 4.13: d-axis inductance vs slot-opening width](image)

The results confirm the high degree of precision of the proposed analytical formulas. The comparative is notable considering that the analytical formulas have been derived under a number of simplifications,
including the assumption of an infinite permeable material. On the basis of the obtained results, a slot-opening width of \( b_0 = 3.1 \) mm has been selected. Regarding this point, it must be noted that the calculations have been conducted assuming a magnet temperature of 100 °C. As it will be latter shown, the final temperature for the designed prototype is somewhat lower (around \( T_{PM} = 72.4 \) °C), meaning that the resulting flux linkage due to the permanent magnets, \( \Lambda_{m1} \), will be higher than expected and hence, the inductance will be lower than the required base inductance. This fact will be further discussed in a latter section.

### 4.3.3 Magnetic coupling

Finally, the effect of material saturation on magnetic coupling has been checked. Figure 4.14 shows the variation on the mutual inductance between phases with machine loading. As it has been done in the previous chapter, mutual inductances by the FEM have been calculated using a magnetic energy approach and introducing same current magnitudes with opposite sign in two phases.

![Figure 4.14: Influence of magnetic saturation on the mutual inductances for the designed prototype](image)

The results indicate that magnetic coupling between phases is extremely low up to phase currents twice their rated value. Since magnetic saturation is a very localized phenomena, it affects the mutual coupling between the various phases differently, leading to uneven values of the mutual inductances. Although the previous simulation gives a measure of the magnetic coupling between phases, the current excitation corresponds to quite an unrealistic case (current in just two phases). Advancing a result from the post-fault control chapter, in a 5-phase PMSM under a single phase open-circuit fault, it is common to shift the currents in the non-faulted phases and increase their magnitude by \( 1.382 \) in order to maintain the electromagnetic torque [123,279]. In case phase A is open-circuited, the currents in the rest of the phases are modified in the following way:

\[
\begin{align*}
    i_A &= \sqrt{2} I \cos (\theta_e + \psi - 0\alpha_m) \\
    i_B &= \sqrt{2} I \cos (\theta_e + \psi - 1\alpha_m) \\
    i_C &= \sqrt{2} I \cos (\theta_e + \psi - 2\alpha_m) \\
    i_D &= \sqrt{2} I \cos (\theta_e + \psi - 3\alpha_m) \\
    i_E &= \sqrt{2} I \cos (\theta_e + \psi - 4\alpha_m)
\end{align*}
\]

Under an open-circuit fault, the flux linkage in phase A corresponds to just the flux linkage due to the permanent magnets and the currents in the rest of the phases. Subtracting the flux linkage due to the PMs, the effect of the mutual coupling between phases is shown in Figure 4.15. As it can be appreciated, magnetic coupling between phases is low even at heavy loading conditions. It should be noted that due to the loss of a phase and magnetic saturation, in order to achieve twice the rated torque, the phase current in the healthy phases is tripled (for \( T = 2T_{nom}, I \approx 3I_{nom} \)).

Owing to these results, it can be concluded that the machine design is adequate in terms of magnetic isolation between phases.
Chapter 4. Design of a Fault-Tolerant PMSM Prototype

4.4 Thermal analysis

In the present thesis, Motor-CAD has been used as a thermal analysis tool [234]. The constant development of the program, together with the many literature examples that have used it over the years with satisfactory results [231, 280, 281], support the confidence in the conducted analysis. Anyhow, good engineering practice calls for latter experimental validation of the thermal analysis results.

Figure 4.16 represents the geometrical Motor-CAD model for the designed prototype. The lumped-parameter thermal model for the machine is shown in Figure 4.5.

Although Motor-CAD allows for transient simulations and even simulations under fault conditions [234], during the design process of the machine, only the steady state copper and magnet temperatures at rated conditions have been calculated. Ideally, a number of additional simulations should have been conducted in order to check the capability of the machine to thermally withstand the continuous operation in the whole speed range both under healthy and faulted conditions. However, due to time constraints during the design process, this has not been possible. It is expected to perform these simulations in a near future and check the results with the aid of experimental tests.

The results of the thermal analysis are presented in a latter section, together with the rest of the prototype data.

4.5 Mechanical considerations

A number of measures have been taken in order to ensure the constructability of the machine. The present section discusses some key design aspects from the mechanical point of view. Due to time constraints, it has not been possible to perform a force and vibration analysis on the machine structure. A proper machine design should address these aspects, together with the generated acoustic noise level.
4.5. MECHANICAL CONSIDERATIONS

4.5.1 Magnet segmentation

It has been shown in a number of papers [148, 149, 192], that radial and axial segmentation of the permanent magnets is a good measure to reduce rotor losses. Owing to the large MMF harmonic and subharmonic content of the chosen winding topology, it has been decided to segment the magnets in order to keep magnet losses and hence, temperatures at adequate levels. Due to the small pole-pitch of the designed machine, only axial segmentation has been applied to the permanent magnets. Each rotor pole has been divided in 12 axial segments.

4.5.2 Magnet fixation

In surface-mounted, inner rotor, permanent magnet machines, the magnets must be fixated to the rotor by some mechanism in order to ensure that the magnets stick to the rotor despite the centripetal acceleration forces acting on them. Different methods are usually applied to ensure magnet fixation in industrial SPM machines. Magnets can be fixated to the rotor by means of an adhesive, by covering the magnets by a bandage [282] or with the aid of a solid cover [283]. In the latter case, the employed cover material must be non-ferromagnetic in order not to increase magnet leakage flux dramatically. Commonly, carbon, non-magnetic steels [284] or copper [194] are used for the solid covers and carbon fiber and fiberglass for magnet banding [285]. Covering the magnets with a conductive, non-magnetic material can be an effective way of reducing eddy-current loss in the permanent magnets and rotor laminations [137,194].

In the related literature, analytical formulas for the design of the covers for high-speed SPM machines have been proposed [285]. Due to the low speed of the considered application (600 rpm), employing a bandage or cover as has been deemed unnecessary and the magnets have been fixated to the rotor just by means of an adhesive: LOCTITE AA326 [286]. The selected adhesive, is a high strength mono-component adhesive that is cured at ambient temperature. According to the manufacturer, it is ideal for magnet bonding and has a service temperature up to 120 °C.

In order to check the validity of the selected adhesive, a small calculation has been conducted. First, the attraction force between a magnet and the rotor stack has been calculated with the aid of a FEA. Then, the tensile stress due to the centripetal acceleration on each magnet has been calculated. The expression for the force and the stress can be derived upon simple calculus:

\[
\begin{align*}
F &= \int \left(r_\omega^2 \right) dm = \frac{1}{24} \rho \omega^2 \alpha_{PM} \alpha_{pole} \alpha_m \left(D_{ext,PM}^3 - D_{ext,r}^3 \right) \\
S &= \frac{1}{2} \rho \omega^2 \alpha_{pole} \alpha_m D_{ext,r} \\
P &= \frac{F}{S} = \frac{1}{12} \rho \omega^2 \frac{1}{D_{ext,r}} \left(D_{ext,PM}^3 - D_{ext,r}^3 \right) 
\end{align*}
\]  

(4.14)

where \( F \) and \( P \) stand for the acceleration force and stress, respectively. \( S \) is the magnet inner surface, \( \rho \) the magnet density and \( \omega \) the mechanical rotation speed (in rad/s). The results for the analysis are shown in Table 4.7. Negative values denote attraction stresses between the magnet and the rotor, while positive values denote repulsion stresses.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Attraction stress between rotor and magnet (no load)</td>
<td>-0.2160 [N/mm²]</td>
</tr>
<tr>
<td>Attraction stress between rotor and magnet (rated load)</td>
<td>-0.1375 [N/mm²]</td>
</tr>
<tr>
<td>Stress due to centripetal acceleration (at 600 rpm)</td>
<td>0.0064 [N/mm²]</td>
</tr>
<tr>
<td>Stress due to centripetal acceleration (at 6000 rpm)</td>
<td>0.6386 [N/mm²]</td>
</tr>
<tr>
<td>Adhesive tensile strength (at 22 °C)</td>
<td>24 [N/mm²]</td>
</tr>
<tr>
<td>Adhesive tensile strength (at 120 °C)</td>
<td>18 [N/mm²]</td>
</tr>
</tbody>
</table>

Table 4.7: Stresses over a magnet

Since the tensile strength of the adhesive is higher than the difference between the stress due to centripetal acceleration and the attraction stresses between magnet and rotor, it can be considered that the magnets will remain fixated to the rotor under any condition. In reality, the previous calculation relies on a perfect
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

contact between the magnets, the adhesive and the rotor laminations. It has been shown in [208] that the tensile stress needed to separate a magnet from a rotor made of laminations is much lower than the one that could be expected assuming a perfect contact between the materials. In fact, a reduction of 75% in the effort needed to separate the magnets was experimentally measured for the LOCTITE AA326 adhesive (tensile stress of 6 N/mm² at 22 °C). In any case, the large safety factor there still is, even at speeds much higher than the rated one, provides confidence on that no additional magnet fixation measures are required.

In addition to using an adhesive to fixate the magnets to the rotor, small tips have been included in the rotor lamination in order to ensure a proper fitting of the magnets. Figure 4.17 shows a sketch of the rotor lamination where the aforementioned tips can be appreciated. The thickness of the tips has been chosen as low as possible in order to keep magnet leakage flux at bay. Table 4.8 shows the flux linkage values due to the permanent magnets with and without rotor tips.

Figure 4.17: Sketch for rotor lamination

<table>
<thead>
<tr>
<th>Without rotor tips</th>
<th>With rotor tips</th>
</tr>
</thead>
<tbody>
<tr>
<td>PM flux linkage (1ˢᵗ harm.), Λₘ₁</td>
<td>102.76 [mWb₉₉₉₉₉₉]</td>
</tr>
</tbody>
</table>

Table 4.8: Flux linkage variation due to rotor tips (Tₚₘ = 75 °C)

4.5.3 Pressure plates

Two distinct steel parts have been designed to act as pressure plates (or clamping rings) and keep the rotor and stator stack laminations compacted. A non-magnetic stainless steel (AISI 304) has been used in order no to interfere with the magnetic field distribution in the machine. Sketches for both parts are shown in Figure 4.18.

Additional losses due to the conductive nature of the designed parts have not been considered. In reality, as demonstrated in [168,190], losses in the metallic pressure plates of an electrical machine can have a significant impact on machine performance, specially when using FSCWs. The rich harmonic content of the air-gap MMF distribution due to the stator windings can lead to high losses in the rotor clamping rings. In order to evaluate these losses, the authors in [168,190] propose to perform so-called 2.5D FEM simulations (assumption of geometry axi-symmetry with periodic angular variation of the electromagnetic fields).

The reasons to ignore these loss components in the present design have been, on the one hand, the large effective air-gap and the moderate application frequency. The conservative physical air-gap, together
with the fact that the outer diameter of the rotor pressure plate (60 mm) is much lower than the mean air-gap diameter (88.5 mm), will contribute to difficult the penetration of MMF harmonics into the rotor metallic material. Additionally, the rated frequency for the design (110 Hz) is moderate as compared to the high frequencies studied in [190] (rated frequency: 233.3 Hz, max. frequency: 1167 Hz).

In the future, this assumption should be checked via FEM simulations and/or experimental tests.

4.6 Prototype data

The following section presents the results of the design process for the fault-tolerant PMSM prototype. The prototype data is divided among different categories: geometric data, winding data, material data, etc. The operating characteristics for the machine have been obtained via 2D FE simulations considering a magnet temperature of 75°C ($B_{r,m} = 1.28$ T, $\mu_{r,m} = 1.0543$) and a winding temperature of 85°C ($R_{ph} = 0.9695\Omega$). The simulations have been conducted for a star-connected machine under purely sinusoidal currents.

In addition to the prototype data and operating curves, the assembly process of the machine is shown as well.

4.6.1 Machine characteristics

The main characteristics of the designed prototype are collected in Tables 4.9 – 4.13.
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

### Table 4.9: Geometric data for the PMSM prototype

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of stator slots, $Q$</td>
<td>20</td>
</tr>
<tr>
<td>Stator outer diameter, $D_{\text{ext, } s}$</td>
<td>166 [mm]</td>
</tr>
<tr>
<td>Air-gap mean diameter, $D_{\text{gap}}$</td>
<td>88.5 [mm]</td>
</tr>
<tr>
<td>Rotor iron outer diameter, $D_{\text{ext, } r}$</td>
<td>77 [mm]</td>
</tr>
<tr>
<td>Stack length, $l$</td>
<td>120 [mm]</td>
</tr>
<tr>
<td>End-winding overhang, $l_{\text{end} - \text{wdg}}$</td>
<td>23 [mm]</td>
</tr>
<tr>
<td>Slot-opening height, $h_0$</td>
<td>3 [mm]</td>
</tr>
<tr>
<td>Slot height, $h$</td>
<td>29.5 [mm]</td>
</tr>
<tr>
<td>Slot opening width, $b_0$</td>
<td>3.1 [mm]</td>
</tr>
<tr>
<td>Slot top width, $b_2$</td>
<td>17.4 [mm]</td>
</tr>
<tr>
<td>Air-gap thickness, $g$</td>
<td>1.5 [mm]</td>
</tr>
<tr>
<td>Magnet arc width, $\alpha_{PM}$</td>
<td>0.66</td>
</tr>
<tr>
<td>Rotor tip thickness, $h_{r, \text{tip}}$</td>
<td>1 [mm]</td>
</tr>
<tr>
<td>Rotor inertia (shaft incl.), $J_r$</td>
<td>4.811 $\times 10^{-3}$ kg m$^2$</td>
</tr>
</tbody>
</table>

### Table 4.10: Winding data for the PMSM prototype

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of phases, $m$</td>
<td>5</td>
</tr>
<tr>
<td>Number of stator slots, $Q$</td>
<td>20</td>
</tr>
<tr>
<td>Number of coil groups, $N_{\text{grp}}$</td>
<td>2</td>
</tr>
<tr>
<td>Number of coils per phase, $N_{\text{ph}}$</td>
<td>2</td>
</tr>
<tr>
<td>Number of turns per coil, $Z$</td>
<td>94</td>
</tr>
<tr>
<td>Conductor wire diameter, $D_{\text{cond, } w}$</td>
<td>0.566 [mm]</td>
</tr>
<tr>
<td>Number of parallel conductors, $Z_{\text{paral}}$</td>
<td>6</td>
</tr>
<tr>
<td>Slot fill factor, $k_f$</td>
<td>0.423</td>
</tr>
<tr>
<td>Conductor copper diameter, $D_{\text{cond, } Cu}$</td>
<td>0.5 [mm]</td>
</tr>
<tr>
<td>End winding mean length per turn</td>
<td>20.4 [mm]</td>
</tr>
</tbody>
</table>

### Table 4.11: Material data for the PMSM prototype

<table>
<thead>
<tr>
<th>Material</th>
<th>Property</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Iron material M500-50A</td>
<td>Lamination thickness</td>
<td>0.5 [mm]</td>
</tr>
<tr>
<td>Density</td>
<td>Young's modulus</td>
<td>210 – 220 [GPa]</td>
</tr>
<tr>
<td>Yield strength</td>
<td>Tensile strength</td>
<td>405 [MPa]</td>
</tr>
<tr>
<td>Electrical conductivity</td>
<td>Thermal conductivity</td>
<td>36 [W/(m·K)]</td>
</tr>
<tr>
<td>Specific heat</td>
<td></td>
<td>452 [J/(kg·K)]</td>
</tr>
<tr>
<td>Magnet material NdFeB</td>
<td>Magnet grade</td>
<td>N4SH</td>
</tr>
<tr>
<td>Density</td>
<td>Electrical conductivity</td>
<td>0.625 [MS/m]</td>
</tr>
<tr>
<td>Thermal conductivity</td>
<td>Specific heat</td>
<td>420 [J/(kg·K)]</td>
</tr>
<tr>
<td>Copper wire</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Copper density</td>
<td>Electrical conductivity at 20 °C</td>
<td>58 [MS/m]</td>
</tr>
<tr>
<td>Thermal conductivity</td>
<td>Specific heat</td>
<td>400 [J/(kg·K)]</td>
</tr>
<tr>
<td>Pressure plate material</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Material AISI 304</td>
<td>Density</td>
<td>8000 [kg/m$^3$]</td>
</tr>
<tr>
<td>Young's modulus</td>
<td>Yield strength</td>
<td>215 [MPa]</td>
</tr>
<tr>
<td>Tensile strength</td>
<td>Electrical conductivity</td>
<td>1.39 [MS/m]</td>
</tr>
<tr>
<td>Thermal conductivity</td>
<td>Specific heat</td>
<td>500 [J/(kg·K)]</td>
</tr>
</tbody>
</table>
## 4.6. PROTOTYPE DATA

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electromagnetic power, $P$</td>
<td>1700 [W]</td>
</tr>
<tr>
<td>Phase current, $I$</td>
<td>4.820 [A&lt;sub&gt;RMS&lt;/sub&gt;]</td>
</tr>
<tr>
<td>Speed, $N$</td>
<td>600 [rpm]</td>
</tr>
<tr>
<td>Input power, $P_n$</td>
<td>1812 [W]</td>
</tr>
<tr>
<td>Copper loss, $P_{Cu}$</td>
<td>112.6 [W]</td>
</tr>
<tr>
<td>Stator yoke loss, $P_{Fe,y,s}$</td>
<td>10.51 [W]</td>
</tr>
<tr>
<td>Magnet loss, $P_{PM}$</td>
<td>0.5334 [W]</td>
</tr>
<tr>
<td>Windage loss, $P_{wind}$</td>
<td>4 [W]</td>
</tr>
<tr>
<td>Power factor</td>
<td>0.7547</td>
</tr>
<tr>
<td>Current angle, $\phi$</td>
<td>0 [°]</td>
</tr>
<tr>
<td>PM flux linkage ($1^{st}$ harm.), $\Lambda_{m1}$</td>
<td>102.5 [mWb&lt;sub&gt;RMS&lt;/sub&gt;]</td>
</tr>
<tr>
<td>PM flux linkage ($5^{th}$ harm.), $\Lambda_{m5}$</td>
<td>0.4 [mWb&lt;sub&gt;RMS&lt;/sub&gt;]</td>
</tr>
<tr>
<td>$d_1$-axis inductance, $L_{d1}$ (no load)</td>
<td>18.87 [mH]</td>
</tr>
<tr>
<td>$d_1$-axis inductance, $L_{d1}$ (rated load)</td>
<td>18.34 [mH]</td>
</tr>
<tr>
<td>$d_1$-axis inductance, $L_{d1}$ (short-circuit)</td>
<td>18.89 [mH]</td>
</tr>
<tr>
<td>Short-circuit current, $I_w$</td>
<td>5.425 [A&lt;sub&gt;RMS&lt;/sub&gt;]</td>
</tr>
<tr>
<td>Air-gap PM flux density, $B_{agap}$</td>
<td>0.8464 [T]</td>
</tr>
<tr>
<td>Stator yoke flux density, $B_{y,s}$</td>
<td>0.7824 [T]</td>
</tr>
<tr>
<td>Linear current density, $A$</td>
<td>32.05 [kA/m]</td>
</tr>
<tr>
<td>Thermal product, $A \cdot j$</td>
<td>13.11 [10&lt;sup&gt;3&lt;/sup&gt;A&lt;sup&gt;2&lt;/sup&gt;/m&lt;sup&gt;2&lt;/sup&gt;]</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Housing temp. (outer surface), $T_{hous}$</td>
<td>69.1 [°C]</td>
</tr>
<tr>
<td>Stator teeth temp. (tooth center), $T_{t}$</td>
<td>74.6 [°C]</td>
</tr>
<tr>
<td>End-winding temp., $T_{end-wdg}$</td>
<td>75.4 [°C]</td>
</tr>
<tr>
<td>Rotor back-iron temp., $T_{b,r}$</td>
<td>72.3 [°C]</td>
</tr>
<tr>
<td>Bearing temperature, $T_{bear}$</td>
<td>69.0 [°C]</td>
</tr>
<tr>
<td>Stator yoke temp., $T_{y,s}$</td>
<td>71.9 [°C]</td>
</tr>
<tr>
<td>Winding temp. (center of the coil), $T_{w}$</td>
<td>81.4 [°C]</td>
</tr>
<tr>
<td>Magnet temp., $T_{PM}$</td>
<td>72.4 [°C]</td>
</tr>
<tr>
<td>Rotor shaft temp. (front), $T_{shaft}$</td>
<td>68.1 [°C]</td>
</tr>
<tr>
<td>Ambient temp., $T_{a}$</td>
<td>25.0 [°C]</td>
</tr>
</tbody>
</table>

### Table 4.12: Working characteristics of the prototype at rated conditions (assuming $T_w = 85$ °C, $T_{PM} = 75$ °C)

### Table 4.13: Estimated machine temperatures at rated conditions (losses calculated assuming $T_w = 85$ °C, $T_{PM} = 75$ °C)

When analyzing the design results, the first consideration that should be made is that the temperatures in the machine, as obtained from the thermal analysis with Motor-CAD, are substantially lower than the initially estimated values (100 °C for both the magnets and the windings). Even though the torque density for the prototype has been doubled compared to the original asynchronous machine, the predicted temperature increase in the machine is fairly low. This is the result of having lower rotor losses and an improved heat transfer to the outside of the housing due to a thinner stator yoke. A same size, higher power machine could have been designed by choosing higher electric and magnetic loading values.

A side effect of the lower magnet temperature is that the flux linkage due to the permanent magnets is higher than initially considered. Consequently, the current needed to deliver the rated torque is lower and thus, the base inductance of the machine is higher. It must be recalled that the slot-opening geometry has been designed assuming a magnet temperature of 100 °C and therefore, the resulting phase self-inductance is lower than the base inductance ($L_b = 21.27$ mH). A proper machine design should have taken this aspect into consideration as modify the slot geometry in order to have a unity phase self-inductance. Although this iterative process is contemplated in the proposed design methodology (see Figure 4.1), due to time constraints it has not been possible to redesign the machine and the prototype has been built with the characteristics shown in Tables 4.9 – 4.13.

Due to the aforementioned fact, the characteristic current of the machine is higher than the current at rated load (12.6% increase) and the conditions for a fault-tolerant design are not totally met. However, due to the large safety margin in machine temperatures, no real problems are expected under fault conditions. The increased copper losses under a short-circuit fault will increase magnet temperature and reduce the flux linkage due to the permanent magnets. Hence, it is expected that machine temperatures at fault conditions will stabilize at reasonable values. This aspect needs to be checked via experimental tests.

As expected, the power factor for the machine at rated conditions is low due to the large phase self-
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

inductance. It approaches the theoretical value of \( 1/\sqrt{2} = 0.7071 \) for a SPM machine. The fact that the self-inductance is a bit lower than 1 pu, together with the phase resistance, contribute to increase the power factor. The relatively low efficiency of the machine is justified by the fact that it is a small-size, low speed machine and by the moderate frequency of the application \[198\]. Machine efficiency could be easily improved just by using a lower loss lamination material. Also, the considered slot fill factor is rather conservative. If the stator was manufactured in a modular way and the coils were automatically wound, the slot fill factor could be vastly increased and hence, copper losses reduced. Owing to the very conservative air-gap, rotor losses are low.

The simulated cogging torque and torque ripple values are very low, as it can be appreciated in Figure 4.19. In the FE simulation of the electromagnetic torque at rated conditions, perfectly sinusoidal phase currents have been considered.

\[
T_{cogg} \quad [N \cdot m]
\]

(a) No load cogging torque

\[
T \quad [N \cdot m]
\]

(b) Electromagnetic torque at rated conditions

Figure 4.19: Cogging torque and electromagnetic torque waveforms for the designed prototype (from FE simulations)

4.6.2 Operating curves

Figure 4.20 shows the operating curves of the machine when the rated current and voltage values are set as a limit \( I_{\text{max}} = 4.82 \text{ A}_{\text{RMS}}, U_{\text{max}} = 100 \text{ V}_{\text{RMS}} \). The Maximum Torque per Ampere (MTPA) strategy is considered. The results have been calculated by post-processing the magnetostatic FE simulations from which the data for the loss calculations have been obtained. Losses at different speeds have been calculated using the second method described in the FE analysis section of this same chapter.

The aforementioned figure shows that, from the base speed onward, the machine is able to deliver more electromagnetic power than the rated one up to a speed of about 5300 rpm. This increase in power is due to an improvement in the power factor when current in the \( d \)-axis is introduced. The constant power speed range of the machine is almost 1:9. It is not infinite due to the fact that the \( d \)-axis inductance is somewhat lower than 1 pu. The term “field-weakening” or “flux-weakening” is easily understandable looking at the curves for stator teeth and yoke flux density values. Although not displayed, under no case is the mean flux density in the magnets under the demagnetization limit of 0.60 T.

The machine efficiency curves for different percentages of the rated electromagnetic power are shown in Figure 4.21. Again the rated current and voltage values are set as a limit and the MTPA strategy is considered. It must be noted that in the calculation of the efficiency, only copper losses and stator iron losses have been considered. Analytical models for the calculation of magnet and rotor iron losses, as well as bearing and windage losses, should be implemented in the future.

4.6.3 Prototype figures

The different parts of the machine, along with the assembly process, can be seen in Figures 4.22 – 4.25.
4.6. PROTOTYPE DATA

Figure 4.20: Operating curves vs speed for the designed prototype (assuming $T_w = 85^\circ C$, $T_{PM} = 75^\circ C$)

Figure 4.21: Machine efficiency curves (assuming $T_w = 85^\circ C$, $T_{PM} = 75^\circ C$)
CHAPTER 4. DESIGN OF A FAULT-TOLERANT PMSM PROTOTYPE

Figure 4.22: Stator of the manufactured prototype

Figure 4.23: Rotor of the manufactured prototype

Figure 4.24: Assembled prototype mounted on base-plate
4.7 Conclusions

In this chapter the whole design process of a fault-tolerant permanent magnet machine has been described. After introducing the design methodology, a brief comment on the chosen application specifications and design constraints has been given. The presented methodology, allows to obtain all the geometrical and constructional aspects of the machine from some very basic specifications. Once the machine has been analytically sized and calculated, a FE analysis and a thermal and hydraulic analysis have been conducted in order to check the validity of the design. Some key aspects of the mechanical design of the machine have been covered as well. Finally, the machine data and its operating characteristics have been presented.

Although the design of an electrical machine is an iterative process in nature due to the many interactions existing between variables, owing to time constraints, a non-optimal solution has been reached. Expected temperature rise in the prototype machine is fairly low; meaning that a higher power machine could have been built. The phase self-inductance is a bit lower than unity due to a magnet temperature lower than the initially estimated one. This results in a short-circuit current higher than the rated current and a non-infinite constant power speed range. If the machine was to be redesigned, these aspects should be addressed.

Anyhow, the validity of the design and the conducted analysis should be checked via experimental tests. These tests will serve as a validation tool for the fault-tolerant capabilities of the designed prototype and for the proposed design methodology.
Chapter 5

Fault Analysis of a PMSM Drive

The present chapter describes the development of a 5-phase permanent magnet synchronous machine drive model suitable for fault analysis. The aim of the proposed model is to serve as a tool to predict machine behavior, extract fault signatures and test post-fault remedial strategies. First, a literature review on PMSMs modeling is presented. Then, the implementation of the drive model is described and the results from fault simulations are presented. Finally, a Finite Element analysis is conducted in order to check the validity of the proposed model.

5.1 Modeling of a PMSM drive

When analyzing the behavior of an electrical machine subject to faults, developing methods for fault diagnosis or proposing new fault-tolerant control strategies, models to predict the performance of a PMSM drive are needed. These models should give accurate estimates of the most representative machine variables, while being computationally efficient in order to allow the inclusion of the control strategy in the simulations.

In general terms, the behavior of a PMSM depends on a large number of variables, such as: the input voltages, the supply frequency, the temperature distribution, etc. As it has been discussed in the previous chapter, in order to keep the complexity of the problem on an adequate scale, it is usual to make some assumptions regarding material properties [276]. Most commonly, the following simplifications are adopted:

- Variation of material properties with physical variables (e.g. temperatures, frequency, etc.) is neglected.
- The reaction field due to induced eddy currents in the iron laminations and in the permanent magnets is not considered.
- For low voltage machines making use of random-wound copper wires, skin effect is ignored.

With the previous assumptions, the voltage equations for a \( m \)-phase PMSM are:

\[
\begin{align*}
    u_{ph}(t) &= R_{ph}i_{ph}(t) + \frac{d\lambda_{ph}(t)}{dt} \\
    \lambda_{ph}(t) &= \lambda_{ph}(\theta_{e}(t), i_{A}(t), i_{B}(t), \ldots i_{m}(t)).
\end{align*}
\]

where \( u_{ph} \) stands for the phase voltage, \( R_{ph} \) is the phase resistance, \( i_{ph} \) the phase current and \( \lambda_{ph} \) the flux linked by each phase. Essentially, the phase flux linkages are a function of the electrical angle between the rotor and the stator and the instantaneous phase currents: \( \lambda_{ph}(t) = \lambda_{ph}(\theta_{e}(t), i_{A}(t), i_{B}(t), \ldots i_{m}(t)) \). If the rotor position and the phase currents are selected as state variables, modeling a PMSM involves just having a proper definition of these functions. However, building and accurate model that retains the main machine characteristics while being fast to simulate is not easy task.
Apart from the most simple linear models [34, 287], a number of methods have been presented in the related literature to model the previous relationships. Generally speaking, the different methods obtain the flux linked by each phase by approximating in one way or another the magnetic field distribution in the machine. In the following, a short literature review on PMSM modeling is conducted. Then, a simple 5-phase machine model is presented and finally, some comments regarding more advanced aspects of machine modeling are given.

5.1.1 Previous work on PMSM modeling

Overall, the methods to calculate the magnetic field distribution in a PMSM can be divided into four categories, going from the simplest and less precise, to the most complicated and accurate one:

- One-dimensional analytical methods.
- Two-dimensional analytical methods.
- Magnetic circuit modeling.
- FEM.

One-dimensional analytical methods

Methods equivalent to the permeance function approach commonly employed in the analysis of induction machines have been used for synchronous machines as well [134,135]. These methods generally compute the radial air-gap flux density distribution as the quotient between the magnetic potential and the air-gap length $g$ [288] (or the product between the MMF and the permeance):

$$B_{agap}(\theta) = \mu_0 \left( -U_s(\theta) + U_r(\theta) \right) / g(\theta)$$

where $U_s$ and $U_r$ are the stator and rotor magnetic potential, respectively, and $g$ is the air-gap length. Due to the non-uniform armature MMF distribution, stator slotting, rotor eccentricity etc. all three quantities can be modeled as angle dependent.

In these methods, the tangential component of the air-gap magnetic field is overlooked and most commonly, iron is considered infinitely permeable. In order to account for magnetic saturation, saturation factors or simple magnetic circuits are usually employed [289]. Such simple magnetic circuits can also be used to account for the effect of stator slotting or for salient pole machines [290].

As a quite recent example, in [291] an analytical model for synchronous reluctance and IPM machines with a single flux barrier is presented. The model is extended in [292] and [293] to two and four flux barrier configurations, respectively. The previous models are used in the minimization of both the torque ripple and the stator iron losses. In [294] a one-dimensional model for outer rotor parallel-magnetized SPM machines is described. The model is employed to minimize the cogging torque by optimizing the magnet pole arc. Many other examples of one-dimensional magnetic field models for PMSMs can be readily found in the related literature, specially with relation to the computation of the cogging torque, e.g. [295–298].

Two-dimensional analytical methods

The simple one-dimensional calculation methods described in the previous point can, quite accurately, predict the magnetic field distribution in small air-gap machines. For SPM or slot-less machines, in which the effective air-gap is large, the air-gap flux density variation with the radial component may be significant and more refined methods are usually preferred. In this sense, a lot of effort has been made to accurately describe the magnetic field distribution due to the permanent magnets and the armature excitation in radial flux SPM machines, e.g. [210–213,218,299].
The aforementioned works directly solve the Ampère-Maxwell equation in terms of the magnetic vector potential $A$. This way, obtaining the magnetic field distribution in different parts of the machine is equivalent to solving a number of associated Laplace or Poisson problems with certain boundary conditions. From the magnetic vector potentials, both the radial and the tangential components of the flux densities can be derived. The main simplification in these models is to assume an infinite or constant iron permeability. Saturation of the rotor and the stator iron can be accounted for by artificially increasing the air-gap length.

Starting from the work in [302], a first family of these methods solves the Laplace problem associated to the air-gap and magnet regions of a SPM machine in polar coordinates by considering a smooth uniform air-gap [210,211]. The effect of stator slotting is modeled by multiplying the radial component of the flux density with a permeance function depending on the slot geometry. In [212,303] a 2D relative permeance function is calculated based on a single infinitely deep rectilinear slot and by applying a conformal transformation method. A very similar approach is considered in [304,305] to predict the magnetic field distribution in IPM machines. In [306,307] the concept of complex relative permeance is introduced in order to have a better estimate of the tangential component of the air-gap magnetic field. In [182,308] the effect of stator slotting is accounted for by introducing the relative permeance function as a new boundary condition. A model for calculating the armature magnetic field distribution and the winding inductances in slotless SPM machines is described in [309].

These analytical models have been improved over the years to account for complex phenomena, such as: rotor eccentricity [308,310], tangential PM magnetization and variation of magnet magnetization with the radial component [311], radial or parallel magnetized magnets and overlapping and non-overlapping stator windings [218], induced rotor eddy current reaction field [312], etc. In addition to calculating the magnetic field distribution in a PMSM, these models have been used to predict the back-EMF waveform [213,306], the air-gap winding inductances [303], the electromagnetic and cogging torques [182,306,313,314], the rotor eddy current losses [149,192,312,315], the magnetic forces [179,182,316,317], etc.

A similar approach using rectangular coordinates has been employed to predict the magnetic field distribution inside the slot of AC machines [195,318–320]. From these, more accurate expressions of the slot leakage inductances than classically used formulas based on a one-dimensional analysis have been proposed.

Instead of assuming a smooth uniform air-gap and introducing a permeance function, a second family of two-dimensional calculation methods apply conformal transformation techniques to map the complex slot and air-gap geometry of an electrical machine into simpler domains for which the magnetic field distribution can be analytically solved [321–323]. These transformations may be defined by analytical complex functions [324] or by numerical computations in the case more complex slot geometries are considered [325].

Finally, the last family of two-dimensional calculation methods solves the magnetic field distribution in the whole machine by dividing it into different subdomains, i.e.: magnets, air-gap, slot-openings and slots, and by applying the corresponding boundary conditions to the interfaces between subdomains [299,326,327]. The first exact machine models that simplify the slot geometry [328–330] have been latter improved to include the tooth-tips [331–333]. The effect of rotor eccentricity has been included in [334]. Subdomain models have been used to calculate the back-EMF [299,332], the armature reaction [335], the electromagnetic and cogging torques [299,332,336], the eddy current losses in the magnets [337] and in the windings [338–340], etc.

Very comprehensive reviews on one-dimensional and two-dimensional analytical models for the magnetic field calculation in PM machines can be found in [299,329,341].

**Magnetic circuit modeling**

The analytical two-dimensional models described in the previous point are a powerful and precise tool for the magnetic field analysis in PMSM machines. Specifically, the results obtained with the subdomain models have been reported to be almost indistinguishable from those obtained by the FEM in case an infinite iron permeability is considered [333,336]. However, in order to address the effect of iron saturation, saturation factors are most commonly employed [219] which can lead to inaccurate results.
in the case of heavily saturated machines. In the case of IPM machines, in which the rotor leakage flux and magnetic saturation play a significant role, it is not easy to analytically predict the rotor magnetic field distribution [342]. In these cases, modeling the machine by means of a magnetic circuit is a good choice [215,343–345].

Instead of calculating the precise magnetic field distribution in each point of the machine, these methods calculate the flux densities in the different machine regions by employing a lumped parameter equivalent magnetic circuit. The machine is modeled in terms of permeances (or reluctances) and magnetic potential sources (or magnetic flux sources) derived from the geometry of the machine. Hence, the name of magnetic circuit modeling [346]. This approach has also been termed Magnetic Equivalent Circuit (MEC) method [347], reluctance network method [220], Permeance network method [348], Dynamic Reluctance Mesh Modeling [349], etc. The basic idea in all these methods is the same: to discretize the machine into a number of elements or flux-paths for which the direction of the magnetic field is assumed beforehand. This assumption regarding the magnetic field distribution in the machine requires the magnetic flux-paths to be carefully modeled. When the effective air-gap of the machine is large or under heavy saturation conditions, this is no trivial matter [349]. Therefore, it is common to divide each machine element (e.g. a tooth) into several lumped elements to account for the possibility of the magnetic flux flowing into more than one direction [350–352].

These lumped parameter magnetic circuits have been used in the modeling of all kinds of electrical machines, e.g.: SPM machines [345,353], IPM machines [215,344,346], induction machines [31,352,354], switched reluctance machines [347,355], synchronous machines [351,356], etc. In addition to the simple magnetic models used in the sizing and analysis of electric devices of all kinds [342,357,358], more complex models that account for the motion of the rotor or the effect of the skewing have been proposed. The motion of the rotor is usually modeled by redefining the air-gap magnetic circuit for each rotor position [346,359,360] or by defining the air-gap permeances as position dependent parameters [220,345,348]. In addition to two-dimensional geometries, in order to investigate the effects of skewing and end-leakage fluxes, a number of 3D magnetic circuit models have been proposed as well [345,361–363]. The magnetic circuit models have been applied in the dynamic simulation of PMSM drives [349,364], as well as in the fault analysis of these systems [365].

As they are based on a lumped parameter representation of the machine, it may be argued that the magnetic circuit models cannot capture the finer details of the flux density waveform like the analytical methods do; specially if the air-gap permeances are modeled based on the overlapping area between the rotor and the stator surfaces. In order to overcome this limitation, some authors propose to first calculate the permeances in the most critical elements (e.g. air-gap, slot-opening) in an analytical or numerical way (2D Laplace equation solving, FEM, Schwarz-Christoffel conformal mapping, etc.) and then couple the calculated permeances with the rest of elements of the magnetic circuit [366–368]. Another option to couple the magnetic circuit and 2D analytical methods is through the boundary conditions of the last one [369].

**Finite Element method**

The last common method applied in the modeling and analysis of electrical machines is the FEM. It is a very flexible method that can account for complex phenomena such as: iron non-linearities, rotor motion, eddy current reaction field, rotor eccentricity, etc. [370]. It produces very accurate results at the expense of an increased computational time if compared to other methods. The working principles of the FEM are widely known and will not be covered here [371,372].

The FEM has been extensively used in the simulation and fault analysis of PMSM drives [373,374]. However, due to its longer calculation time, it is still preferred for validating other simpler analytical or numerical methods [375].

**Conclusions for previous work on PMSM modeling**

The main conclusions on the conducted literature review on PMSM modeling are presented below.
5.1. MODELING OF A PMSM DRIVE

The analytical one-dimensional methods have proven to be quite adequate in the determination on basic machine parameters, such as the air-gap flux and the mean electromagnetic torque [134, 135]. However, since they neglect the tangential component of the air-gap flux density, more refined methods are needed in order to accurately predict phenomena like the cogging torque or the magnetic forces [336]. The analytical 2D methods can be very precise in terms of the calculation of time and space magnetic field harmonics. Nevertheless, saturation and cross-coupling effects cannot be easily accounted for and magnetic equivalent circuit methods may be preferable [342]. These address effectively the effects of iron saturation or 3D leakage fluxes, but may need to be coupled to other analytical or numerical methods to give an accurate description of the air-gap magnetic field distribution [368]. The FEM produces the more accurate results, but is too time consuming, specially if the control algorithm of the whole drive system needs to be incorporated in the calculations [21].

In addition to the methods discussed above, other analytical and numerical techniques have been proposed to predict the magnetic field distribution in electrical machines, such as the charge model method [376] or the boundary element method [377]. However, their use is not as widespread.

Since all the previous methods give an estimate of the magnetic field distribution in the machine, they can be used in the dynamic simulation and fault analysis of PMSMs. The models can be directly coupled with the electrical circuit equations [378] or be used to compute the machine parameters (flux linkages, inductances, etc.) previous to the simulation of the drive [379].

5.1.2 Model proposal

When analyzing a PMSM, it is common to represent the electrical variables in a reference frame rotating at synchronous speed. For a 3-phase machine, this involves transforming the time-dependent phase variables into the \(dq0\) reference frame. For a 5-phase PMSM, this can be done in different ways, such as the \(ABCDE\) to \(d1q1d3q30\) transformation (see Appendix B) [158]. For a PMSM equipped with a FSCW, the dependency of the electrical variables represented in a synchronously rotating reference frame with the angular position is low due to the low number of pole and slot periodicities. This can be appreciated in Figure 5.1, where the flux linkage waveforms obtained by FE simulations for the designed prototype at rated conditions are shown.

![Figure 5.1: Flux linkage variation with the angular position (\(i_{q1} = \sqrt{2}I_{nom}, i_{d1} = i_{d3} = i_{q3} = 0\))](image)

Therefore, the existence of space harmonics and the dependency of the \(d - q\) flux linkages with the angular position can generally be neglected. Such an approximation is usually acceptable in FSCW PMSMs, but not in integral-slot winding or BLDC machines, where the induced voltage waveform is trapezoidal.

Additionally, if the machine is star connected and the neutral point is not available, the sum of phase currents is forced to zero unless there is an asymmetrical short-circuit fault. This further reduces the number of parameters affecting the flux linkages and therefore, healthy 3-phase PMSMs can be modeled if the relationships between the \(d - q\) flux linkages and currents are known [34], \(\lambda_d = \lambda_d(i_{d}, i_{q}), \lambda_q = \lambda_q(i_{d}, i_{q})\). For this matter, any of the approaches described in the previous subsection for calculating the magnetic field distribution in a machine can be used.

However, even with the previous assumptions, the higher number of state variables in a 5-phase machine
makes things complicated. If instead of coupling the magnetic field and electrical circuit equations, the flux linkages are determined beforehand, having a complete modeling of the current dependencies for a star connected 5-phase machine involves storing four parameters ($\lambda_{d1}$, $\lambda_{q1}$, $\lambda_{d3}$, $\lambda_{q3}$), depending each on four variables ($i_{d1}$, $i_{q1}$, $i_{d3}$, $i_{q3}$).

Alternatively, the simple 5-phase PMSM model proposed in [158] is considered in the present thesis. For the sake of simplicity, it is assumed that the stator flux linkage due to the PMs is composed just of its first and third harmonics:

$$
\begin{align*}
\lambda_{A,m}(\theta_e) &= \lambda_{m1} \sin (1(\theta_e - 0\alpha_m)) + \lambda_{m3} \sin (3(\theta_e - 0\alpha_m)) \\
\lambda_{B,m}(\theta_e) &= \lambda_{m1} \sin (1(\theta_e - 1\alpha_m)) + \lambda_{m3} \sin (3(\theta_e - 1\alpha_m)) \\
\lambda_{C,m}(\theta_e) &= \lambda_{m1} \sin (1(\theta_e - 2\alpha_m)) + \lambda_{m3} \sin (3(\theta_e - 2\alpha_m)) \\
\lambda_{D,m}(\theta_e) &= \lambda_{m1} \sin (1(\theta_e - 3\alpha_m)) + \lambda_{m3} \sin (3(\theta_e - 3\alpha_m)) \\
\lambda_{E,m}(\theta_e) &= \lambda_{m1} \sin (1(\theta_e - 4\alpha_m)) + \lambda_{m3} \sin (3(\theta_e - 4\alpha_m))
\end{align*}
$$

When transformed to the $d1q1d3q30$ reference frame, the PM flux linkage components reflect as constant value parameters. Assuming no space harmonics and cross-coupling terms, the constitutive equations for the machine are:

$$
\begin{align*}
\lambda_{d1} &= \lambda_{m1} + L_{d1}i_{d1} \\
\lambda_{q1} &= L_{q1}i_{q1} \\
\lambda_{d3} &= \lambda_{m3} + L_{d3}i_{d3} \\
\lambda_{q3} &= L_{q3}i_{q3} \\
\lambda_0 &= L_0i_0
\end{align*}
$$

where $L_{d1}$, $L_{q1}$, $L_{d3}$, $L_{q3}$ and $L_0$ are the $d1$-, $q1$-, $d3$-, $q3$- and 0-axis inductances, respectively. If no saturation and cross-coupling effects are considered, variables $L_{d1}$, $L_{q1}$, etc. become constant value parameters and the flux linkage in each axis only depends on the current in that same axis. It is important to notice that, in order to account for possible asymmetrical fault conditions, the constitutive equation for the 0-axis flux linkage has been added to the model proposed by [158].

It has been argued in [21] that, under fault conditions, the effect of saturation and space harmonics is more pronounced due to the unbalance nature of the machine. However, since the aim of this chapter is to develop a model that it is able to capture the main characteristics of a 5-phase machine drive while keeping things simple, the previous machine model with constant parameters has been considered. In fact, machine parameters have been estimated by considering an infinite permeability iron material. In order to have consistent results with the used FE program, loss components other than copper losses are not considered in the constitutive equations, but they are computed afterward. A brief comment on how to account for these losses in the machine equations is given in the next section.

### 5.1.3 More advanced machine models

The simple machine model proposed in the previous subsection can be extended in order to account for more complex phenomena. Specifically, different suggestions on how to implement the following phenomena are given below:

- Inclusion of space harmonics.
- Magnetic saturation.
- Iron losses.
- Frequency dependency.
- Temperature dependency.
Inclusion of space harmonics

In general terms, the flux linkages and the inductances referred to the \( d-q \) reference frame depend on the angle between the rotor and the stator due to the non-perfectly sinusoidal armature MMF distribution and the non-uniform air-gap permeance caused by stator slotting, local saturation, etc. The most straightforward way to account for space harmonics is to include a higher number of harmonics in the flux linkage waveform due to the PMs and to model the \( d-q \) inductances as angle dependent variables:

\[
\begin{align*}
\lambda_{A,m}(\theta_e) &= \sum_{n=1}^{\infty} \lambda_{m,n} \sin(n(\theta_e - 0\alpha_m)) \\
L_{d1}(\theta_e) &= L_{d1,0} + \sum_{n=1}^{\infty} L_{d1,n} \sin(n(\theta_e + \phi_{d,n})) \\
L_{q1}(\theta_e) &= L_{q1,0} + \sum_{n=1}^{\infty} L_{q1,n} \sin(n(\theta_e + \phi_{q,n}))
\end{align*}
\]  

(5.5)

where parameters \( \lambda_{m,n}, L_{d1,n}, L_{q1,n} \), etc. must be determined beforehand; either by a numerical (e.g. FEM) or analytical method \([380, 381]\). Another different approach for including space harmonics is to employ dynamic reluctance models in which the magnetic circuit of the machine is varied for each rotor position \([349]\). This way, the dependency of motor parameters on the mechanical angle of the rotor is inherently accounted for.

Magnetic saturation

In the proposed machine model (equation (5.4)), the flux linkage in each \( d-q \) axis has been assumed to be dependent only on the current flowing in that same axis. In reality, due to magnetic saturation, the flux linkages in an AC machine are a function of the instantaneous currents in all axes. Since iron saturation is a very localized phenomena, its effect on the machine parameters depends a lot on the particular machine design. The effect of the current in one axis affecting the flux linkage in another one is usually termed cross-saturation or cross-coupling magnetization \([382]\).

In 3-phase machines, it is usual to obtain the \( \lambda_d = \lambda_d(i_d, i_q) \) and \( \lambda_q = \lambda_q(i_d, i_q) \) dependencies via experimental tests or FE simulations. For instance, \([382]\) describes an experimental method to determine the flux linkage characteristics for a PMSM in its entire operating range. The method is based on applying square-wave pulses to the machine windings while the rotor is locked. A similar method based on a current decay test in which \( i_d \) or \( i_q \) transitions from a high initial to a zero value is proposed in \([383]\) for the parameter estimation of a synchronous reluctance machine.

Once the flux linkage characteristics have been obtained, they can be stored in look-up tables or be approximated by analytical functions in order to be latter used in the dynamical simulation of the PMSM drive \([384]\). In this sense, the effect of magnetic saturation on an IPM machine is approximated in \([34]\) by making the \( q \)-axis inductance current dependent:

\[
L_q = L_q(i_q) = \begin{cases} 
L_q = C_1 |i_q|^2 & \text{for } L_q \leq L_{q,max} \\
L_q = L_{q,max} & \text{otherwise}
\end{cases}
\]

(5.6)

Current dependency in the \( d \)-axis is not included in the formulation due to the larger effective air-gap along the \( d \)-axis in an IPM machine. It is argued that in a IPM machine, the \( d \)-axis is almost free of saturation as a consequence of the rotor magnet slots behaving like large air-gaps. The same approach with a different formulation is proposed in \([385]\).

The previous approach may only be valid for anisotropic machines with a large difference in the effective air-gap lengths. More generally, \([386]\) reviews different methods to obtain the machine parameters from FE simulations for an IPM synchronous machine and describes three simplified models to represent the flux linkage dependencies:
and likewise for \( \lambda_q (i_d, i_q) \). In [380], saturation in an induction machine has been accounted for by making the air-gap permeance function to be dependent on the position and the level of air-gap flux. The effect of the variable air-gap length reflects in the \( d-q \) flux linkages as second order harmonic terms depending on saturation. The equivalent circuit for the proposed model incorporates one stator and two rotor circuits for each one of the \( d \) and \( q \) axes. Continuing with the same idea, [21] accounts for iron saturation in an induction machine by making the stator mutual inductances dependent on the angular position and the flux level:

\[
\begin{aligned}
    \lambda_d (i_d, i_q) &= (C_1 e^{-A_1 i_q} + C_2 e^{-A_2 i_q}) e^{-\left(C_3 e^{-A_3 i_q} + C_4 e^{-A_4 i_q}\right)} i_d \\
    &+ (C_5 e^{-A_5 i_q} + C_6 e^{-A_6 i_q}) e^{-\left(C_7 e^{-A_7 i_q} + C_8 e^{-A_8 i_q}\right)} i_d \\
\end{aligned}
\]

and likewise for \( \lambda_q (i_d, i_q) \). In [380], saturation in an induction machine has been accounted for by making the air-gap permeance function to be dependent on the position and the level of air-gap flux. The effect of the variable air-gap length reflects in the \( d-q \) flux linkages as second order harmonic terms depending on saturation. The equivalent circuit for the proposed model incorporates one stator and two rotor circuits for each one of the \( d \) and \( q \) axes. Continuing with the same idea, [21] accounts for iron saturation in an induction machine by making the stator mutual inductances dependent on the angular position and the flux level:

\[
\begin{aligned}
    L_{AB} &= L_{BA} = -\frac{L_m}{2} \left( 1 + \sum_{n=1}^{\infty} K_{2n} \sin (2n (\theta_c + \phi_n - 0\alpha_m)) \right) \\
    L_{BC} &= L_{CB} = -\frac{L_m}{2} \left( 1 + \sum_{n=1}^{\infty} K_{2n} \sin (2n (\theta_c + \phi_n - 1\alpha_m)) \right) \\
    L_{CA} &= L_{AC} = -\frac{L_m}{2} \left( 1 + \sum_{n=1}^{\infty} K_{2n} \sin (2n (\theta_c + \phi_n - 2\alpha_m)) \right) \\
\end{aligned}
\]

The mutual inductances between phases are considered as the sum of the first three harmonic components. The magnitudes of the inductance harmonics \( K_{2n} \) and the phase displacements \( \phi_n \) vary with the machine load and must be determined via FE simulations or experimental measurements [21].

Generally speaking, the constant parameters in any of the previously mentioned models must be fitted to the previously obtained data by employing a least squares approach or a multidimensional optimization method. The main advantages of having analytical expressions for the current dependencies of the flux linkages and/or the inductances are that less information needs to be stored and that the derivative terms needed in the voltage equations can be readily obtained.

A different and quite simple method to model the effect of magnetic saturation in an electrical machine is by modifying the associated electrical circuit. This approach has been traditionally used in the modeling of iron saturation in induction [388,389] and synchronous machines [390–392]. It allows to model magnetic saturation as well as iron losses, as it will be later discussed.

Although the previous references are all related to 3-phase machines, similar procedures can be used to account for magnetic saturation in 5-phase PMSMs.

Finally, the last method that can be used to account for magnetic saturation without resorting to the dynamical FE simulation of a PMSM drive is to employ a magnetic circuit model or permeance network approach. Numerous models have been described in the related literature, as stated in the previous section.
Iron losses

As previously commented, one way to account for loss components other than stator copper losses is to compute them once the instantaneous electric variables (currents, voltages, etc.) have been determined. In this approach, the computed machine losses are regarded as being equivalent to an additional load torque and they are subtracted from the electromagnetic torque in order to calculate the actual or effective torque of the machine. This torque is then used in the calculation of the mechanical speed of the PMSM drive (see Figure 5.4). A number of analytical models have been proposed in the related literature to compute the losses in the various machine elements, such as:

- Stator iron [221,222].
- Rotor iron [393].
- Magnets [149,224,225].
- Bearings [226].
- Air friction [232].
- etc.

The previous approach is the one considered in the present thesis.

A different method to account for machine losses in a dynamic simulation is to modify the associated electrical circuit by introducing a dedicated parameter. This approach has been traditionally used in the transient simulation of synchronous [390] and induction machines [198]. For instance, [394–396] model stator iron losses by inserting a resistor \( R_c \) in parallel with the magnetizing branch in both the \( d \)-axis and \( q \)-axis equivalent circuits. This way, machine losses are assumed to be dependent on the air-gap flux linkage. The considered loss model, shown in Figure 5.2, is used in [397] to investigate control algorithms for loss minimization in a PMSM.

![Dynamic PMSM d – q equivalent circuits accounting for iron losses](image)

An equivalent way of including iron losses in the electrical circuit of the machine is by adding a dedicated series resistor [398,399]. In [400] it is shown that both methods are mathematically equivalent, although the parallel model, despite being more intuitive, has the disadvantage of having an increased number of state variables. Additional parameters can be introduced in the equivalent electrical circuit to separate the iron losses among those related to the main flux and the leakage flux [401] or those depending on the voltage and the currents [402].

It must be noted that, by including machine losses in the electrical circuit of the machine, losses are assumed to be dependent of the square of the fundamental frequency. This is usually true for the eddy current losses in the iron laminations, but not for other loss components. Therefore, in order to include other losses in the electrical circuit of the machine, the iron loss resistances must be varied according to machine speed [403].
CHAPTER 5. FAULT ANALYSIS OF A PMSM DRIVE

Frequency dependency

Although parameter dependency with frequency is often neglected when examining the drive’s response to faults, it is common knowledge that the windings in an electrical machine are subject to both skin and proximity effects. These phenomena cause the resistance at high frequencies to be much higher than the DC measured resistance and, at the same time, they cause a decrease in the inductance values. Also, at high frequencies, the effect of turn to turn and turn to ground capacitances start to be significant.

Skin effect and proximity effect are usually considered in high voltage machines that use copper bar windings instead of round stranded conductors. In these machines, the conductor height may be significantly larger than the skin depth at supply frequency and the skin effect has to be definitely addressed. Traditionally, the frequency effects on the resistance and the inductance have been accounted by introducing frequency dependent correction factors [135, 198]. The previous references give analytical expressions for these factors, derived upon a one-dimensional magnetic field analysis and assuming purely sinusoidal currents. A 2D analytical subdomain model for the calculation of eddy-current losses in the bar windings of a PM machine is described in [338].

In low voltage machines using random-wound round stranded conductors, calculating the frequency effects is more complicated. Most commonly, FE simulations have been used to approximate the leakage flux distribution inside the slots and investigate the skin and proximity effects [404, 405]. It has been shown in [406] that FSCW PMSMs can exhibit high AC to DC resistance ratios at high frequencies in spite of using stranded conductors. In [407] an analytical method is presented to model the frequency effects and estimate the AC losses in electrical machines making use of stranded conductors. According to the investigations, transposition and the use of litz wire are two effective ways of reducing the resistance increase due to skin and proximity effects [408]. In [339, 340] proximity losses are predicted by an analytical 2D subdomain machine model. Similar methods could be used to model frequency effects when analyzing PMSM drive faults.

Other authors have proposed different methods in order to incorporate the high frequency effects in the equivalent circuit of the machine [409, 410]; specially in the context of terminal overvoltage simulation when excited by a PWM signal.

Temperature dependency

Thermal dynamics in electrical machines are usually slow, with time constants ranging from dozens of minutes to several hours. Therefore, it is common practice to overlook machine parameter dependency with temperature when analyzing the drive response to faults.

An exception to the above are the winding inter-turn short-circuit faults. In such faults, a substantial amount of heat is generated in the shorted turns, so temperature can locally increase in a short amount of time. In [365] a simple thermal model together with a dynamic reluctance network is employed to investigate the short-circuit withstand capability of a PMSM. A similar study is carried out in [411] with the aid of a more complex lumped-parameter thermal network.

5.2 Model implementation

In the following section, the actual implementation of the PMSM drive model is described. To supply the machine, a 5-phase half bridge inverter has been considered. First, the implementation of the PM machine model is detailed. Then, the same process is carried out for the associated inverter. Finally, the coupling of both models, together with the associated torque and speed controllers, and the building of the integral drive model is discussed.

The drive model has been implemented in the MATLAB/Simulink® environment [277,412]. The PMSM and inverter models have been created making use of the blocks available in the SimPowerSystems™ library [413]. Thus, the electrical connection between elements is represented in a direct manner, as shown in Figure 5.3. Faults can be modeled just by modifying the electrical circuit of the drive. In
the diagram below, green lines represent Simulink signals and red lines electrical connections between elements.

Figure 5.3: Electrical circuit of the 5-phase PMSM drive model

5.2.1 Machine model

This sections specifies the mathematical relationships between the machine variables. The general scheme for the PMSM model is represented in Figure 5.4. The rules that govern the machine behavior are divided among electrical, mechanical and loss equations.

For the modeling of machine losses, a number of analytical methods have been proposed, as discussed in the previous section. Due to time constraints, no loss model besides stator copper losses has been actually implemented in the present work. Although this matter should be addressed in the future, this simplification is not expected to be heavily detrimental for the conclusions derived from this chapter. When experimentally validating the model results with experimental data, losses other than copper losses will be included into the load torque term.

Figure 5.4: Overview of the PMSM model: inputs (green), outputs (red)
Electrical equations

Current equation  For a star-connected electrical machine with an isolated neutral point \( N \), the sum of phase currents is forced to zero:

\[
\sum_{i=1}^{m} i_{ph} = 0 \Rightarrow i_0 = 0 \tag{5.10}
\]

Please note that, in the present thesis, the phase currents \( i_{ph} \) refer to the current flowing through the machine windings. During healthy state conditions, these currents will be equal to the currents flowing through the corresponding inverter branch. However, this may not hold true for certain fault conditions, such as in the case of a single phase to neutral short-circuit. In this case, the branch current in the faulted phase has an additional path to flow and equation (5.10) does not hold anymore.

Voltage equations  The voltage equations for a star-connected \( m \)-phase PMSM are:

\[
\left\{ \begin{array}{l}
u_{phN}(t) = R_{ph} i_{ph}(t) + \frac{d\lambda_{ph}(t)}{dt} \\
ph = A, B, \ldots m
\end{array} \right. \tag{5.11}
\]

After some tedious math, it can be demonstrated that applying equations (5.4) and (5.10) to (5.11) leads to:

\[
\sum_{i=1}^{m} u_{phN} = 0 \Rightarrow u_0 = 0 \tag{5.12}
\]

The previous equation can be used to calculate the phase to neutral voltages \( u_{phN} \) from the phase to phase voltages \( u_{ph-ph} \):

\[
\left\{ \begin{array}{l}
u_{AN} = \frac{1}{5}(4u_{AB} + 3u_{BC} + 2u_{CD} + u_{DE}) \\
u_{BN} = \frac{1}{5}(-u_{AB} + 3u_{BC} + 2u_{CD} + u_{DE}) \\
u_{CN} = \frac{1}{5}(-u_{AB} - 2u_{BC} + 2u_{CD} + u_{DE}) \\
u_{DN} = \frac{1}{5}(-u_{AB} - 2u_{BC} - 3u_{CD} + u_{DE}) \\
u_{EN} = \frac{1}{5}(-u_{AB} - 2u_{BC} - 3u_{CD} - 4u_{DE})
\end{array} \right. \tag{5.13}
\]

Therefore, the model only requires 4 variables as voltage inputs. By applying the \( ABCDE \) to \( d_1q_1d_3q_30 \) transformation to the voltage equations expressed vectorially in the phase reference frame, \( x_{ph} = P^{-1} [x_{dq}] \):

\[
[u_{ph}] = R_{ph} [i_{ph}] + \frac{d[y_{ph}]}{dt} \tag{5.14}
\]

\[
[P^{-1}][u_{dq}] = R_{ph} [P^{-1}] [i_{dq}] + \frac{d}{dt} ([P^{-1}] [\lambda_{dq}]) \tag{5.15}
\]

\[
[u_{dq}] = R_{ph} [i_{dq}] + \omega_e [A] [\lambda_{dq}] + \frac{d[y_{dq}]}{dt} \tag{5.16}
\]
where:

\[
[A] = [P] \frac{d}{dt} ([P^{-1}]) = \begin{bmatrix}
0 & -1 & 0 & 0 & 0 \\
1 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & -3 & 0 \\
0 & 0 & 3 & 0 & 0 \\
0 & 0 & 0 & 0 & 0
\end{bmatrix}
\]

(5.17)

Finally, introducing the constitutive equations in (5.16) and rearranging the terms results in:

\[
\begin{align*}
\frac{di_{d1}}{dt} &= \frac{1}{L_{d1}} (u_{d1} - R_{ph} i_{d1} + \omega_e L_{q1} i_{q1}) \\
\frac{di_{q1}}{dt} &= \frac{1}{L_{q1}} (u_{q1} - R_{ph} i_{q1} - \omega_e \lambda_{m1} - \omega_e L_{d1} i_{d1}) \\
\frac{di_{d3}}{dt} &= \frac{1}{L_{d3}} (u_{d3} - R_{ph} i_{d3} + 3\omega_e L_{q3} i_{q3}) \\
\frac{di_{q3}}{dt} &= \frac{1}{L_{q3}} (u_{q3} - R_{ph} i_{q3} - 3\omega_e \lambda_{m3} - 3\omega_e L_{d3} i_{d3}) \\
\frac{di_{0}}{dt} &= \frac{1}{L_0} (u_0 - R_{ph} i_0)
\end{align*}
\]

(5.18)

Equation (5.18) allows to compute the current components from the input voltages. The calculation process starts with obtaining the phase to neutral voltages from the phase to phase voltages as in (5.13). Then, the Park transform is applied to the phase voltages to get \([u_{dq}]\). The \(d-q\) currents are calculated from the above equation. Finally, the Park transform is reversely applied to obtain the phase currents.

**Torque equation**  The electromagnetic torque delivered by the machine can be deduced from equation (5.16) by conducting an energy balance [293]. Disregarding copper losses, it is:

\[
\frac{5}{2} \left(u_{d1} i_{d1} + u_{q1} i_{q1} + u_{d3} i_{d3} + u_{q3} i_{q3} + u_0 i_0\right) = T \omega_m + \frac{dW_m}{dt}
\]

(5.19)

where \(T\) is the electromagnetic torque and \(W_m\) is the magnetic energy. If the \(d-q\) currents and the rotor position \(\theta_m\) are selected as state variables, equation (5.19) becomes, after some manipulations:

\[
T d\theta_m = \frac{5}{2} p (\lambda_{d1} i_{q1} - \lambda_{q1} i_{d1} + 3\lambda_{d3} i_{q3} - 3\lambda_{q3} i_{d3}) + \frac{5}{2} \left(i_{d1} d\lambda_{d1} + i_{q1} d\lambda_{q1} + i_{d3} d\lambda_{d3} + i_{q3} d\lambda_{q3} + i_0 d\lambda_0\right)
\]

\[
-dW_m
\]

(5.20)

The magnetic energy and the flux linkages can be expressed as a function of the selected independent state variables [293]:

\[
\begin{align*}
\frac{dW_m}{d\theta_m} &= \frac{\partial W_m}{\partial \theta_m} d\theta_m + \frac{\partial W_m}{\partial i_{d1}} di_{d1} + \frac{\partial W_m}{\partial i_{q1}} di_{q1} + \frac{\partial W_m}{\partial i_{d3}} di_{d3} + \frac{\partial W_m}{\partial i_{q3}} di_{q3} + \frac{\partial W_m}{\partial i_0} di_0 \\
\frac{d\lambda_{d1}}{d\theta_m} &= \frac{\partial \lambda_{d1}}{\partial \theta_m} d\theta_m + \frac{\partial \lambda_{d1}}{\partial i_{d1}} di_{d1} + \frac{\partial \lambda_{d1}}{\partial i_{q1}} di_{q1} + \frac{\partial \lambda_{d1}}{\partial i_{d3}} di_{d3} + \frac{\partial \lambda_{d1}}{\partial i_{q3}} di_{q3} + \frac{\partial \lambda_{d1}}{\partial i_0} di_0 \\
\frac{d\lambda_{q1}}{d\theta_m} &= \frac{\partial \lambda_{q1}}{\partial \theta_m} d\theta_m + \frac{\partial \lambda_{q1}}{\partial i_{d1}} di_{d1} + \frac{\partial \lambda_{q1}}{\partial i_{q1}} di_{q1} + \frac{\partial \lambda_{q1}}{\partial i_{d3}} di_{d3} + \frac{\partial \lambda_{q1}}{\partial i_{q3}} di_{q3} + \frac{\partial \lambda_{q1}}{\partial i_0} di_0 \\
\frac{d\lambda_{d3}}{d\theta_m} &= \frac{\partial \lambda_{d3}}{\partial \theta_m} d\theta_m + \frac{\partial \lambda_{d3}}{\partial i_{d1}} di_{d1} + \frac{\partial \lambda_{d3}}{\partial i_{q1}} di_{q1} + \frac{\partial \lambda_{d3}}{\partial i_{d3}} di_{d3} + \frac{\partial \lambda_{d3}}{\partial i_{q3}} di_{q3} + \frac{\partial \lambda_{d3}}{\partial i_0} di_0 \\
\frac{d\lambda_{q3}}{d\theta_m} &= \frac{\partial \lambda_{q3}}{\partial \theta_m} d\theta_m + \frac{\partial \lambda_{q3}}{\partial i_{d1}} di_{d1} + \frac{\partial \lambda_{q3}}{\partial i_{q1}} di_{q1} + \frac{\partial \lambda_{q3}}{\partial i_{d3}} di_{d3} + \frac{\partial \lambda_{q3}}{\partial i_{q3}} di_{q3} + \frac{\partial \lambda_{q3}}{\partial i_0} di_0 \\
\frac{d\lambda_0}{d\theta_m} &= \frac{\partial \lambda_0}{\partial \theta_m} d\theta_m + \frac{\partial \lambda_0}{\partial i_{d1}} di_{d1} + \frac{\partial \lambda_0}{\partial i_{q1}} di_{q1} + \frac{\partial \lambda_0}{\partial i_{d3}} di_{d3} + \frac{\partial \lambda_0}{\partial i_{q3}} di_{q3} + \frac{\partial \lambda_0}{\partial i_0} di_0
\end{align*}
\]

(5.21)
Substituting (5.21) in (5.20), the energy balance becomes:

$$T d\theta_m = \frac{5}{2} p (\lambda_d i_{q1} - \lambda_q i_{d1} + 3\lambda_d i_{q3} - 3\lambda_q i_{d3}) d\theta_m + \frac{\partial W'_m}{\partial \theta_m} d\theta_m + (...) d\theta_m + (...) d\theta_q + ...$$  \hspace{1cm} (5.22)

where:

$$\frac{\partial W'_m}{\partial \theta_m} = \frac{5}{2} (i_{d1} \frac{\partial \lambda_d}{\partial \theta_m} + i_{q1} \frac{\partial \lambda_q}{\partial \theta_m} + i_{d3} \frac{\partial \lambda_d}{\partial \theta_m} + i_{q3} \frac{\partial \lambda_q}{\partial \theta_m} + i_0 \frac{\partial \lambda_0}{\partial \theta_m}) - \frac{\partial W_m}{\partial \theta_m}$$ \hspace{1cm} (5.23)

is the magnetic coenergy derivative with respect to $\theta_m$. Since the state variables are independent of each other, the following equation is deduced for the electromagnetic torque \[293\]:

$$T = \frac{5}{2} p (\lambda_d i_{q1} - \lambda_q i_{d1} + 3\lambda_d i_{q3} - 3\lambda_q i_{d3}) + \frac{\partial W'_m}{\partial \theta_m}$$ \hspace{1cm} (5.24)

At steady state conditions, the magnetic coenergy derivative has a zero average value and therefore, it is commonly omitted in the calculation of the average torque. In fact, this simplification is usually applied in preliminary machine design and when studying control strategies for a PMSM \[293\]. The term $\partial W'_m/\partial \theta_m$, however, is important in the computation of the torque ripple; being one of the main contributors to this phenomenon.

From this point onwards, the coenergy derivative term is neglected and only the electromagnetic torque due to the flux linkage is considered. Introducing (5.4) in (5.24) leads to:

$$T = \frac{5}{2} p (\lambda_m i_{q1} + (L_{d1} - L_{q1}) i_{d1} i_{q1} + 3\lambda_{m3} i_{q3} + 3 (L_{d3} - L_{q3}) i_{d3} i_{q3})$$ \hspace{1cm} (5.25)

The values of the machine electrical parameters considered in the present chapter are collected in Table 5.1. They have been calculated by conducting FE simulations on the designed prototype machine, assuming an infinitely permeable iron material.

| PM flux linkage (1st harm.), $\lambda_{m1}$ | 0.1468 [Wb] | PM flux linkage (3rd harm.), $\lambda_{m3}$ | 0.0002 [Wb] |
| d1-axis inductance, $L_{d1}$ | 19.16 [mH] | q1-axis inductance, $L_{q1}$ | 19.20 [mH] |
| d3-axis inductance, $L_{d3}$ | 19.18 [mH] | q3-axis inductance, $L_{q3}$ | 19.18 [mH] |
| 0-axis inductance, $L_0$ | 19.19 [mH] | Phase resistance, $R_{ph}$ | 0.9695 [Ω] |

Table 5.1: Electrical parameters for the PMSM (assuming $T_w = 85$ °C, $T_{PM} = 75$ °C and infinite permeability of the iron material)

**Mechanical equations**

The mechanical equation for the machine is:

$$J \frac{d\omega_m}{dt} = T - (T_{load} + T_{loss} + B\omega_m)$$ \hspace{1cm} (5.26)

where $J$ is the total rotor plus load inertia, $T_{load}$ the load torque, $T_{loss}$ the torque corresponding to machine losses other than copper losses and $B\omega_m$ the friction torque. In reality, it is expected that friction, along with other loss terms, does not follow a linear variation. When carrying the experimental tests, all the negative torque components will be associated to the load in a global torque load term, $T'_{load}$:

$$J \frac{d\omega_m}{dt} = T - (T'_{load} + B\omega_m)$$ \hspace{1cm} (5.27)
5.2. MODEL IMPLEMENTATION

Modeling of machine faults

Two fault scenarios have been considered for the machine windings: an open-circuit fault and the short-circuit of the whole phase winding. Since winding inter-turn short-circuits are incipient faults that rapidly evolve into other kind of faults [30], they have not been considered in the present thesis. A short-circuit fault between two phase windings inside the machine is expected to be extremely rare owing to the physical separation between phase windings that results from the winding topology. It can, however, occur at the machine terminals due to an improper connection or due to physical damage on the input cables.

Open-circuit faults The modeling of a phase winding open-circuit fault is made just by adding an “infinite” value resistance in series with the corresponding phase winding.

Short-circuit faults The short-circuit between two phases at the machine terminals is modeled simply by adding a branch between the corresponding phases in the electrical diagram of the machine.

In case a whole phase winding in a star connected electrical machine is short-circuited, the branch current has an additional path to flow and equation $\sum_{m} i_{ph} = 0$ does not hold anymore. As a consequence, $\sum_{m} u_{phN} \neq 0$ and the voltage equations must be modified to account for the fault. For instance, if phase $A$ is short-circuited, $u_{AN} = 0$, and the relationship between the phase to phase voltages and the phase to neutral voltages becomes:

$$
\begin{align*}
    u_{AN} &= 0 \\
    u_{BN} &= -u_{AB} \\
    u_{CN} &= -u_{AB} - u_{BC} \\
    u_{DN} &= -u_{AB} - u_{BC} - u_{CD} \\
    u_{EN} &= -u_{AB} - u_{BC} - u_{CD} - u_{DE}
\end{align*}
$$

(5.28)

similar equations are derived considering the short-circuit fault of other phases. By applying equation (5.28) instead of (5.13), the short-circuit of phase $A$ can be readily considered.

5.2.2 Inverter model

The half bridge inverter model has been implemented by using an ideal DC voltage source, a capacitor branch and 10 ideal power semiconductor switches with an antiparallel diode. The switches are characterized by having a negligible internal resistance and infinite snubber resistance and capacitance values. The switch model can be easily extended to incorporate more detailed device characteristics, such as: internal resistance and inductance during on state, diode forward voltage, current fall time and tail time, snubber circuit characteristics, etc. No driver circuit for the power switches is considered.

The machine supply voltages are controlled by selecting the appropriate gate signals that drive the switches. A sinusoidal PWM strategy has been chosen for most of the simulations. In this strategy, the width of the pulses is computed by comparing the reference signals $u_{phO}^*$ with a carrier sawtooth wave.

When the reference signals are higher than the carrier signal, the PWM signals are in their high state. Otherwise, they are in their low state. The implemented PWM strategy is schematically shown in Figure 5.5.

The width of the pulse expressed as a fraction of the switching period $T_s$ is termed duty cycle $D$ and is obtained as:

$$
D_{ph} = \frac{1}{2} + \frac{u_{phO}^* [n]}{U_{DC}}, \quad D_{ph} \in [0, 1]
$$

(5.30)
As long as the voltage magnitude does not lie in the over-modulation region, the resulting PWM voltage supply waveform has a first harmonic of the same amplitude as the reference signal. The small angular displacement that is observable between the reference voltage in Figure 5.5a and the resulting first harmonic of the phase to neutral waveform in Figure 5.5d is due to the low switching frequency in the example shown above. In the upcoming drive simulations a switching frequency of 4 kHz has been considered.

Modeling of inverter faults

Inverter faults are modeled just by modifying the corresponding switch logical input. For instance, in the case of a permanent opening of a power switch, a 0 signal is permanently commanded to the gate of the faulted switch. Likewise, in the case of a permanent closing of a switch, a 1 signal is commanded to the switch gate. In this last case, it is assumed that the inverter is equipped with the necessary hardware protection to detect the fault and automatically open the other switch in the same leg in order to avoid the short-circuiting of the DC bus.

5.2.3 Integrated drive model

The integral drive model is the result of electrically coupling the machine and inverter models as represented in Figure 5.3. Additionally, a current loop and a speed loop have been added in order to simulate the closed loop operation of the machine. The complete drive model diagram is shown in Figure 5.6. A standard vector control strategy has been used to control the currents. In the diagram below, the reference values for the variables are distinguished by the subscript $^\ast$. 

---

Figure 5.5: Sinusoidal PWM voltage supply

(a) Reference signal (blue) and sampled signal (red)

(b) Duty cycle (blue) and carrier waveform (red)

(c) PWM voltage supply, $u_{AO}$

(d) Resulting phase to neutral voltage, $u_{AN}$
5.2. MODEL IMPLEMENTATION

Measurements

The inverter and machine models provide the control system with the measurement of the currents and the rotor position, respectively. As it would be done in a real drive, the phase currents are measured at the inverter output. It is important to notice that, in case of a short-circuit fault, the branch currents and the currents flowing through the machine windings may be different, depending on the affected phases.

For the measurement of the currents and the rotor position, ideal, non-saturating sensing elements have been considered. No delay is assumed between the input and the output of a sensor. For the measurement of the branch currents, four sensors corresponding to phases A, B, C and D have been considered. The fifth branch current is calculated considering the sum of the branch currents is zero:

$$i_{E, \text{branch}} = i_{A, \text{branch}} + i_{B, \text{branch}} + i_{C, \text{branch}} + i_{D, \text{branch}}$$  \hspace{1cm} (5.31)

The branch currents are latter transformed to the $d-q$ reference frame in order to serve as inputs to the current controller.

Current control loop

The machine is operated using a standard vector control strategy in the $d-q$ reference frame. Another option would have been to control the currents referred to the $\alpha-\beta$ reference frame [414] (see Appendix G). The current control is achieved by using PI controllers with a feed-forward voltage compensation term to improve the system response. A scheme of the controllers for the $i_{d1}$ and $i_{q1}$ currents is shown in Figure 5.7. The control of currents $i_{d2}$ and $i_{q2}$ is achieved in a similar fashion. Since the machine is star connected, no control is implemented for current $i_0$.

The $d-q$ voltage references generated by the current controllers are transformed back to the fixed stator reference frame and introduced as inputs of the PWM voltage supply strategy. The parameters of the current controllers considered in the present chapter are collected in Table 5.2. The values for the proportional and integral gains have been calculated in order for the dynamic response of the control system to have a natural frequency of $\omega_n = 1382$ rad/s with a damping coefficient of $\zeta = 0.707$. All four PI current controllers use the same parameters.

<table>
<thead>
<tr>
<th>Proportional term, $K_{P,i}$</th>
<th>36.53</th>
</tr>
</thead>
<tbody>
<tr>
<td>Integral term, $K_{I,i}$</td>
<td>36656</td>
</tr>
</tbody>
</table>

Table 5.2: Parameters of the PI current controllers
5.3 Simulation results

The following section presents the simulation results obtained with the aid of the proposed PMSM drive model. In addition to checking the healthy state operation of the machine, a number of fault scenarios have been simulated. Specifically, the following machine and inverter faults have been considered:

- Winding open-circuit faults: single phase, two adjacent phases and two non-adjacent phases.
- Terminal short-circuit faults: single phase, two adjacent phases and two non-adjacent phases.
- Permanent opening of an inverter switch: single switch and single leg faults.
- Permanent closing of a single inverter switch.

Other possible faults in a PMSM drive are so severe (e.g. short-circuiting the DC bus via the high side and low side transistors of the same leg) that they demand immediate remedial actions to be taken in order to prevent the equipment from being damaged [25]. This is usually accomplished by adding protection circuits that monitor the saturation voltage across the transistors. When the voltage drop exceeds a certain threshold, the fault is detected and the inverter is immediately shut down, leading to an uncontrolled operation of the drive. The sudden shutdown of PMSM drives at high speeds has been thoroughly discussed in [385], for instance, with a special focus on IPM machines with a large saliency ratio.

Since the aforementioned protection circuits are a standard feature in inverter drives nowadays and due to the fact that the possibility of multiple faults (e.g. open-circuit and short-circuit fault) happening at the same time is deemed remote [25], no other faults have been considered in the present thesis.

Considering that the speed behavior of a PMSM drive depends on the load characteristics (load inertia, torque/speed curve, etc.), for the sake of simplicity and in order to obtain results as general as possible, an infinite inertia load has been considered. Thus, it has been assumed that the speed of the drive remains constant even after a fault. This assumption is common in the fault analysis of PMSM drives [34,99,415]. Even if this does not correspond to a real case, the electrical time constant of a machine drive is usually many times smaller than the mechanical time constant and the electrical steady state is reached in a short time. Therefore, it is expected that the main conclusions drawn from this section are still valid in a real case scenario.
In the simulations, no current limitation has been considered for the current controllers. The only imposed limitation is that of the available DC bus voltage ($U_{DC} = 300$ V). The simulations have been conducted for the drive operating at rated conditions, that is, at a speed of 600 rpm and a load of 27.06 N·m. Since to study the behavior of the machine under flux-weakening operation is beyond the scope of the present thesis, no higher speeds have been considered.

The obtained simulation results are presented in the following subsections.

### 5.3.1 Healthy drive operation

The first conducted simulation is that of the PMSM drive operating with no faults at rated conditions. The most significant electrical variables for the drive operating at steady state conditions are shown in Figure 5.8. It can be appreciated that the drive is able to accurately follow the current references in the $d-q$ reference frame and deliver the commanded torque. Though the PWM supply strategy introduces a rich harmonic content in the phase to neutral and $d-q$ referred voltages, the voltage harmonics are effectively filtered by the high phase inductance, leading to almost purely sinusoidal currents. The torque ripple introduced by the inverter switching is negligible.

Figure 5.8: Steady state operating curves during healthy drive operation (constant speed of 600 rpm)
5.3.2 Winding open-circuit faults

Open-circuit of a single phase

The simulation results for the opening of a phase winding are collected in Figure 5.9. In the simulation, phase A has been open-circuited at a time corresponding to the fifth electrical period. For the sake of clarity, only the current and torque waveforms are displayed.

In addition to the current in the faulted phase dropping to zero, the main consequence of the opening of a phase is that the current references cannot be adequately followed anymore. From the current controllers perspective, the open-circuit fault modifies the systems inner behavior, so the drive is not able to adequately follow the $i^*_q$ current reference while eliminating the currents in the $d_1$, $d_3$ and $q_3$ axes. As a result, an increased torque ripple is obtained with an important second order harmonic. In order to maintain the average torque, the RMS currents in the healthy phases increase.

The phase currents and the corresponding electromagnetic torque once the steady state has been reached are shown in Figure 5.10. It can be appreciated that, in order to deliver the same mean torque, the currents in the phases adjacent to the faulted phase increase, whereas the currents in the non-adjacent phases more or less maintain their pre-fault RMS values. The high torque ripple and the fact that the dominating torque harmonic under open-circuit conditions is the second harmonic can be easily spotted as well. Depending on the application such a high ripple may be heavily detrimental for the operation of the system and should be dealt with.

![Figure 5.9: Drive’s response to a single phase open-circuit (constant speed of 600 rpm)](image-url)
5.3. SIMULATION RESULTS

Figure 5.10: Steady state operating curves during single phase open-circuit (constant speed of 600 rpm)

**Open-circuit of two adjacent phases**

In this case, the open-circuit of phases C and D has been considered. The drive response to the sudden open-circuit of the two phases is shown in Figure 5.11, whereas the operating curves once the steady state has been reached are presented in Figure 5.12.

Figure 5.11: Drive’s response to the open-circuit of two adjacent phases (constant speed of 600 rpm)

The inability of the current controller to follow the references is more evident in this case, seeing how the ripple in the \(d-q\) currents is higher. The resulting phase currents hardly resemble a sinusoidal waveform anymore and the torque oscillates almost between zero and its maximal value. The dominant torque harmonic under fault conditions is again the second order one. It can be appreciated that, even if the controller reacts to the fault and tries to compensate for it by increasing the healthy phase currents, it cannot maintain the average torque. The reason for this is that, for the considered drive system, the maximum inverter output voltage is reached and the current controllers saturate. As a consequence, the drive cannot deliver the commanded torque and the mean torque value is greatly reduced.
A better controllability for the currents can be achieved by increasing the DC bus voltage. For example, Figure 5.12 shows the steady state current and torque waveforms for two different bus voltage values: $U_{DC} = 300$ V and $U_{DC} = 375$ V (25% increase). As a result of the increased available voltage, the phase currents can increase and the resulting average torque is closer to the demanded torque. The downside to the increased torque is a substantial increase in the machine copper losses; specially in the phases that are adjacent to the faulted ones.

**Open-circuit of two non-adjacent phases**

The simulation results for the open-circuit of two non-adjacent phases are collected below. The faulted phases in this case are phases $B$ and $E$. The transient response of the PMSM drive is displayed in Figure 5.13, whereas the steady state curves are presented in Figure 5.14.

The drive’s response to the open-circuit of two non-adjacent phases is similar to the response in case the opened phases are adjacent; although a quite better response is obtained in terms of average torque. The current controller is saturated and the average torque decays with respect to the healthy state, but the reduction in average torque is lower than in the previous case. In order to alleviate the fault, the current primarily increases in the healthy phase in-between the faulted phases. The torque ripple is comparable to the fault scenario where the open-circuitued phases are adjacent.
Conclusions for winding open-circuit faults

The numerical comparison between the three considered winding open-circuit simulations is presented in Table 5.3. The table shows the results for the average $d - q$ currents and voltages, the mean torque $T_{\text{mean}}$, the torque ripple $\Delta T = T_{\text{max}} - T_{\text{min}}$ and the average copper losses in a period $P_{\text{Cu}}$. In order to ease the comparison with the normal operation of the drive, numerical results for the healthy state are given as well. In all the cases, a constant DC bus voltage of $U_{\text{DC}} = 300$ V has been considered.

The results indicate that the drive can, up to a point, react to the loss of one or several faults by shaping the currents in the healthy phases and increasing their magnitudes. However, it is evident by examining the mean currents in the $d - q$ reference frame that, under a fault scenario, the current controllers fail to adequately follow the current commands. This translates into a change of the resulting phase to neutral voltages, that is specially noticeable when two phases are open-circuited. The resulting voltages have a
significant third harmonic component.

The low voltage margin of the considered PMSM drive leads to an easy saturation of the current controllers and thus, to a reduction in the torque capability when two phases are open-circuited. The fault scenario in which the average torque is further reduced (or in which a greater margin for the increase of currents in order to maintain the mean torque should be provided) is in the open-circuit of two adjacent phases. Generally speaking, the prize to pay for trying to maintain the torque is an increase in the machine losses. The resulting phase currents are limited and not dangerous in terms of the current limit of the inverter switches or in terms of the applied demagnetizing field. A transient thermal analysis should be conducted in order to check for how long can the faulted states be maintained. Another choice is to place a limit on the post-fault currents and allow a derating of the average torque in order to avoid the increase in machine losses.

Under all the considered open-circuit fault conditions, the torque ripple is significant and of the same order of magnitude as the mean torque. This pulsating torque may be an issue in applications demanding a constant torque response (e.g. position control) and should be remedied by modifying the control strategy.

### 5.3.3 Terminal short-circuit faults

**Short-circuit of a single phase**

The simulation results for the short-circuit of a whole phase winding are presented next. In this case, the terminal of the faulted phase is connected to the neutral point of the star connection. Figure 5.15 shows the drive’s response when phase A is short-circuited in its entirety at a time corresponding to the fifth electrical period. In the simulations, it has been assumed that the switching on the corresponding inverter leg is maintained. Since they do not provide much meaningful information, voltage waveforms are not included in the figure.

As a consequence of the short-circuit, the current in the faulted phase initially increases but then it decays due to the fact that it is limited by the high phase inductance. The current references cannot be adequately followed and a high ripple is introduced in the \( d-q \) currents. This ripple is reflected in the resulting electromagnetic torque, which has a significant second order harmonic. The waveforms of the resulting phase currents and electromagnetic torque are more easily appreciated in Figure 5.16, where the operating curves once the steady state has been reached are shown. The measured branch current in the faulted phase is included for comparison as well.
5.3. SIMULATION RESULTS

![Simulation Results](image)

**Figure 5.15:** Drive’s transient response to the short-circuit of a whole phase winding (phase to neutral) (constant speed of 600 rpm)

In observing the figures, it is evident how there is a significant difference between the measured branch current and the current flowing through the faulted phase windings. The RMS values of the steady state currents are close to those when the drive is operating under healthy conditions, but the drive is unable to maintain the average torque. In this case, the reduction in the average torque is not due to saturation of the current controllers, as increasing the DC bus voltage does not result in a higher mean torque being delivered. The reason has more to do with the fact that the fault modifies the plant’s behavior in control terms.
Short-circuit of two adjacent phases

An internal short-circuit between two distinct phases is very unlikely to happen in the considered machine as the chosen winding topology provides for physical isolation between phases. It could however happen at the connections side, where all cables are held together, or in the terminal box due to an improper phase connection. Owing to this, the simulated fault considers the sudden connection of the terminals of two adjacent phases. Since such a fault will likely cause a DC bus short-circuit through the high side switch of one phase and the low side switch of the other phase, in the simulations it has been assumed that the fault is immediately detected by the protection hardware of the inverter and that the inverter legs corresponding to the faulted phases are automatically opened. The switching continues in the remaining three healthy phases.

Figure 5.17 shows the drive’s response to the sudden short-circuit of the terminals of phases $C$ and $D$. Compared to the short-circuit of a single phase, the present fault is more severe in terms of mean torque reduction and increased current magnitudes. The resulting $d - q$ currents largely deviate from the commanded reference values and there is a significant demagnetizing field being applied periodically as a result of $i_{d1}$ and $i_{d3}$ reaching large negative levels.

Figure 5.17: Drive’s response to the short-circuit of two adjacent phases at the machine terminals (constant speed of 600 rpm)

The steady state current and electromagnetic torque waveforms are displayed in Figure 5.18. The phase current waveforms for the healthy phases are very close to those resulting from the open-circuit of two adjacent phases. In fact, the situation in both fault scenarios is very similar: the control of two phases is lost and the current controllers modify the applied voltages in an attempt to deliver the commanded torque with the healthy phases. The difference between the two scenarios lies in that, in the open-circuit case, the currents in the uncontrolled phases are zero, whereas in the short-circuit case they are time-varying variables. As it occurred in the open-circuit fault of two adjacent phases, the current controllers are saturated. Although not displayed in the graphs, the average output torque can be increased by elevating the DC bus voltage.
5.3. SIMULATION RESULTS

Figure 5.18: Steady state operating curves at the short-circuit of two adjacent phases (constant speed of 600 rpm)

Short-circuit of two non-adjacent phases

In this case, it is considered that a short-circuit happens between the terminals of phases B and E. As it has been done in the previous fault scenario, it is assumed that the drive protects itself against shoot-through faults by opening the legs corresponding to the faulted phases. The drive’s transient response to the fault and the resulting steady state operation curves are shown in Figures 5.19 and 5.20, respectively.

Figure 5.19: Drive’s response to the short-circuit of two non-adjacent phases at the machine terminals (constant speed of 600 rpm)

The results are very similar as those obtained in the previous case. The torque waveform exhibits a significant ripple, being the second harmonic the dominant one. Although there is a decrease in the torque capability, the mean torque value is higher than in the case where the faulted phases are adjacent. In trying to counter the fault, it is the current magnitude in the healthy phase in-between the faulted phases the one that increases most.
Conclusions for terminal short-circuit faults

The comparison between the three considered short-circuit simulations is presented in Table 5.4, together with the data for the healthy drive. In all the cases, the same bus voltage \( U_{DC} = 300 \) V has been considered.

<table>
<thead>
<tr>
<th></th>
<th>Healthy</th>
<th>Single phase short-circuit</th>
<th>Two adjacent phases SC</th>
<th>Two non-adjacent phases SC</th>
</tr>
</thead>
<tbody>
<tr>
<td>( i_{d1,\text{mean}} ) [A]</td>
<td>0.000</td>
<td>-1.250</td>
<td>0.379</td>
<td>-2.240</td>
</tr>
<tr>
<td>( i_{q1,\text{mean}} ) [A]</td>
<td>6.703</td>
<td>4.866</td>
<td>3.616</td>
<td>5.502</td>
</tr>
<tr>
<td>( i_{d3,\text{mean}} ) [A]</td>
<td>0.000</td>
<td>-0.017</td>
<td>0.275</td>
<td>0.501</td>
</tr>
<tr>
<td>( i_{q3,\text{mean}} ) [A]</td>
<td>0.000</td>
<td>0.000</td>
<td>-0.856</td>
<td>-0.147</td>
</tr>
<tr>
<td>( u_{d1,\text{mean}} ) [V]</td>
<td>-90.02</td>
<td>-66.52</td>
<td>-47.75</td>
<td>-74.58</td>
</tr>
<tr>
<td>( u_{q1,\text{mean}} ) [V]</td>
<td>109.5</td>
<td>89.41</td>
<td>109.97</td>
<td>77.74</td>
</tr>
<tr>
<td>( u_{d3,\text{mean}} ) [V]</td>
<td>1.614</td>
<td>0.453</td>
<td>34.95</td>
<td>7.754</td>
</tr>
<tr>
<td>( u_{q3,\text{mean}} ) [V]</td>
<td>0.802</td>
<td>0.721</td>
<td>11.16</td>
<td>20.10</td>
</tr>
<tr>
<td>( T_{\text{mean}} ) [N·m]</td>
<td>27.06</td>
<td>19.65</td>
<td>14.58</td>
<td>22.22</td>
</tr>
<tr>
<td>( \Delta T ) [N·m]</td>
<td>0.668</td>
<td>19.77</td>
<td>19.53</td>
<td>39.11</td>
</tr>
<tr>
<td>( P_{Cu} ) [W]</td>
<td>109.0</td>
<td>108.9</td>
<td>116.4</td>
<td>197.8</td>
</tr>
</tbody>
</table>

Table 5.4: Simulation results for short-circuit faults (steady state)

As it occurred with the open-circuit faults, the short-circuit of two phases leads to a significant third harmonic component of the phase to neutral voltages that can be appreciated by the values of \( u_{d3,\text{mean}} \) and \( u_{q3,\text{mean}} \). The current controllers fail to follow the commands and this results in a decrease of the mean torque and an increase of the torque ripple in all the short-circuit scenarios. The average torque reduction is more severe in case two adjacent phases are short-circuited, whereas the highest torque pulsation occurs for the short-circuit fault between two non-adjacent phases. Machine losses increase significantly in this last case, meaning that the faulty state is not susceptible of being maintained for a long time.

In all the considered cases notable demagnetizing armature fields are generated due to the large negative \( i_{d1} \) and \( i_{d3} \) current values. It has been checked via FE simulations that these demagnetizing field levels do not pose a threat for the designed machine. However, if the drive where to operate at higher speeds (i.e. under flux weakening conditions) the demagnetizing currents would probably be higher and this matter should be further studied.
5.3. SIMULATION RESULTS

5.3.4 Inverter faults

Opening of a single transistor

The first considered inverter fault is the permanent opening of a power semiconductor switch. This fault can happen as a result of damage to this element or due to a problem with the logic signal that controls the transistor gate. In this case, the current in the faulted phase is forced to zero during half of the electrical period, unless the machine speed is high enough that the flow of current through the free-wheeling diode of the faulted switch is allowed. In the case under examination, the drive’s response to the permanent opening of the high side transistor of phase $A$ is displayed in Figure 5.21.

![Figure 5.21: Drive’s response to the opening of a single switch (constant speed of 600 rpm)](image)

As in other cases, the drive cannot fully follow the commanded current references and a significant torque ripple appears, though there is no decrease in the average torque value. Since the faulted phase can contribute to the torque during half a period, the dominant torque harmonic in this case is the first order one. In order to maintain the torque, current magnitudes are primarily increased in the phases adjacent to the faulted phase. Although, as a consequence of the fault, negative $d_1$- and $d_3$-axes currents are introduced in the machine, their magnitude is lower than in other fault scenarios. For the considered machine, it is one of the least dangerous faults in terms of increased losses and demagnetizing fields; though, the torque ripple is still high.

The steady state operation curves for the PMSM drive are shown in Figure 5.22.
CHAPTER 5. FAULT ANALYSIS OF A PMSM DRIVE

Figure 5.22: Steady state operating curves at the permanent opening of a single inverter switch (constant speed of 600 rpm)

Opening of an inverter leg

The permanent opening of both transistors in the same leg can occur as a consequence of the control logic commanding the switching of the gates being damaged. This results in the faulted phase being connected to the DC bus only through the free-wheeling diodes that protect the power semiconductor switches.

For the considered machine and, under rated load and speed conditions, the simulation results for the permanent opening of both transistors in the same leg are exactly the same as those obtained for the winding open-circuit of a single phase. Thus, the results will not be shown here. However, it should be noticed that, generally speaking, the open-circuit of a phase winding and the opening of the corresponding inverter leg are conceptually two different faults. For higher speeds, the induced peak back-EMF can become higher than the DC bus voltage and allow the current in the faulted phase to flow through the free-wheeling diodes. In this latter case, energy is returned from the electrical machine to the DC bus. The operation of 3-phase PMSM drives with an opened inverter leg at high speeds has been studied, for instance, in [25].

Since the designed prototype is not meant to be operated at speeds higher than 600 rpm, no further simulations have been conducted. This fault scenario should be further considered, however, if the machine was to be operated in the flux weakening region.

Closing of a single transistor

In this case, the permanent closing of a single transistor is considered. In order to avoid a DC bus shoot-through it is assumed that the protection circuit of the inverter responds by immediately opening the opposing transistor in the same leg. The drive’s transient response to the permanent closing of the high side transistor of phase A is shown in Figure 5.23.

As it can be appreciated in the figure, the considered fault is one of the most severe ones. DC current components appear in all the phases; positive for the faulted phase and negative for the healthy phases if the faulted phase is permanently connected to the high side of the DC bus. In case the low side transistor fails, the signs of the DC current components are reversed. The value of the DC component in the faulted phase is mainly limited by the stator resistance and thus, the fault can result in very high current values [25]. These high currents pose a risk to the machine in terms of current withstand capability of the power semiconductors, PM demagnetization and winding overheating. The extremely high and reversing torque values can also damage the machine bearings, the shaft, the mechanical load and the coupling element.

This is a catastrophic drive fault that calls for immediate detection and remedial action to be taken; most commonly by shutting down the inverter (opening of the remaining switches) [25]. The steady state current and electromagnetic torque waveforms are displayed in Figure 5.24.
5.3. SIMULATION RESULTS

![Graphs showing phase currents, d-q currents, and electromagnetic torque in response to closing a single switch.](image)

Figure 5.23: Drive’s response to the closing of a single switch (constant speed of 600 rpm)

![Graphs showing steady state currents and electromagnetic torque at permanent closing of a single inverter switch.](image)

Figure 5.24: Steady state operating curves at the permanent closing of a single inverter switch (constant speed of 600 rpm)

Conclusions for inverter faults

The numerical results for the simulated faults are collected in Table 5.5. In order not to repeat data, the results for the permanent opening of both transistors in the same leg have not been added. Although many other inverter faults could have been simulated, with combinations of various opened and closed switches, it is assumed that in a fault-tolerant drive the necessary means are provided to detect the faults and place remedial actions, so the chances of one fault causing another and having multiple faults at the same time are slim. Therefore, only the most probable inverter faults have been considered.

Without doubt, the most serious drive fault is that of a single transistor remaining permanently closed. Machine losses increase by several orders of magnitude and large demagnetizing fields are applied. Interestingly enough, the main values of the $d-q$ currents do not deviate from the commanded references as much as one could expect and the mean torque is more or less maintained. However, there are large variations in the currents and the torque that are not appreciated from their mean values.
CHAPTER 5. FAULT ANALYSIS OF A PMSM DRIVE

<table>
<thead>
<tr>
<th></th>
<th>Healthy</th>
<th>Single switch permanent open</th>
<th>Single switch permanent close</th>
</tr>
</thead>
<tbody>
<tr>
<td>$i_{d1, \text{mean}}$ [A]</td>
<td>0.000</td>
<td>−0.204</td>
<td>0.588</td>
</tr>
<tr>
<td>$i_{q1, \text{mean}}$ [A]</td>
<td>6.703</td>
<td>6.868</td>
<td>5.716</td>
</tr>
<tr>
<td>$i_{d3, \text{mean}}$ [A]</td>
<td>0.000</td>
<td>−0.080</td>
<td>0.529</td>
</tr>
<tr>
<td>$i_{q3, \text{mean}}$ [A]</td>
<td>0.000</td>
<td>−0.191</td>
<td>−0.027</td>
</tr>
<tr>
<td>$u_{d1, \text{mean}}$ [V]</td>
<td>−90.02</td>
<td>−94.12</td>
<td>−72.81</td>
</tr>
<tr>
<td>$u_{q1, \text{mean}}$ [V]</td>
<td>109.5</td>
<td>104.8</td>
<td>114.7</td>
</tr>
<tr>
<td>$u_{d3, \text{mean}}$ [V]</td>
<td>1.614</td>
<td>6.915</td>
<td>12.61</td>
</tr>
<tr>
<td>$u_{q3, \text{mean}}$ [V]</td>
<td>0.802</td>
<td>−3.170</td>
<td>21.85</td>
</tr>
<tr>
<td>$T_{\text{mean}}$ [N·m]</td>
<td>27.06</td>
<td>27.72</td>
<td>23.03</td>
</tr>
<tr>
<td>$\Delta T$ [N·m]</td>
<td>0.608</td>
<td>26.39</td>
<td>377.3</td>
</tr>
<tr>
<td>$P_{Cu}$ [W]</td>
<td>109.0</td>
<td>142.3</td>
<td>14443</td>
</tr>
</tbody>
</table>

Table 5.5: Simulation results for inverter faults (steady state)

Of all the considered faults, the permanent opening of a single inverter transistor is the less serious of them all. The mean torque capability of the can be maintained at the expense of increasing machine losses. Again, in order to determine for how long could the fault be sustained without limiting the torque or modifying the control law, a transient thermal analysis should be conducted.

5.4 Finite Element validation

In order to check the validity of the proposed PMSM model, a number of FE simulations have been conducted. Since the FEM considers the actual geometry of the machine, both permeance and armature MMF harmonics are inherently taken into account in the simulations. This way, the pertinence of the simplifications the model has been developed on can be examined. In the FE simulations, an infinite permeability iron material and a zero voltage drop in the power semiconductor switches have been considered.

For the sake of simplicity, the simulations have been conducted assuming a constant motor speed, even after the occurrence of a fault. It has also been assumed that the applied voltage references remain unchanged in case of a fault. These two simplifications are not detrimental for checking the validity of the model and help reducing the complexity of the simulations.

Since simulating the machine with the FE software and a sinusoidal PWM voltage supply would be extremely time consuming, in order to reduce the needed step size and reduce the simulation time, the drive has been simulated with a square-wave voltage supply. Figure 5.25 represents the phase to ground and phase to neutral voltages during normal operation considered in the validation of the model.

![Figure 5.25: Square-wave PWM voltage supply for FE validation](image)

This can be thought of as the most simple form of PWM. The width of the pulse $\alpha$ has been selected so
that a first harmonic corresponding to the rated phase voltage has been obtained:

\[ \alpha = \arcsin \left( \frac{1}{2} + \frac{\pi}{\sqrt{8}} \frac{U_1}{U_{DC}} \right) \]  
(5.32)

Thus, the maximum attainable voltage with this commutation strategy is: \( U_{1,\text{max}} = \sqrt{2} U_{DC} \). For the considered PMSM drive: \( U_1 = 98.91 \, \text{V}_{\text{RMS}}, \, U_{DC} = 300 \, \text{V} \) and \( \alpha = 1.048 \, \text{rad} \). The low number of commutations per electrical period causes a rich low frequency harmonic content of the voltage waveform. This translates into a rich harmonic content of the phases currents and an increased torque ripple.

In the following subsections, the results for the FE validation are collected.

### 5.4.1 Healthy drive operation

The most significant electrical variables for the PMSM drive under healthy operating conditions are shown in Figure 5.26. The figure presents the results obtained both with the proposed analytical model and with the commercial FE software [228]. The used software does not allow specifying initial machine currents, but these are initialized either to 0 or from a magnetostatic computation. Therefore, in order to reach the steady state operation of the machine, a number of electrical periods must be simulated. The results shown in Figure 5.26 correspond to an electrical period at steady state conditions.

As expected, the current harmonic content and the torque ripple are higher than in the sinusoidal PWM supply case.

As it is easily appreciated, both methods lead to fairly similar results; with the advantage that the analytical model is thousands of times faster to simulate. Since the same voltage supply is specified in both methods, there is no difference in the voltage magnitudes. The small differences in the current waveforms are due to some extent to the simplifications undertaken in the developing of the model, but also due to meshing and discretization errors in the FE simulations. In observing the figure, it can be appreciated that the phase currents calculated with the FEM are not equal for all the five phases; fact that does not make much sense, since the conditions for all the phases are the same. When the transient simulation step size is reduced in the FE simulations, the comparative between both methods turns better. The small differences in the phase currents spread to the torque waveform.

The main results for the conducted simulations are collected in Table 5.6. The good agreement between both simulation methods help build confidence in the validity of the proposed model.

<table>
<thead>
<tr>
<th></th>
<th>RMS</th>
<th>RMS, (_{\text{p}})</th>
<th>mean</th>
<th>peak</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phase currents, (i_{\text{ph}}) [A] (mean)</td>
<td>Analytical model</td>
<td>4.908</td>
<td>4.740</td>
<td>–</td>
</tr>
<tr>
<td></td>
<td>FE model (64 points/period)</td>
<td>5.096</td>
<td>4.933</td>
<td>–</td>
</tr>
<tr>
<td></td>
<td>FE model (128 points/period)</td>
<td>4.976</td>
<td>4.806</td>
<td>–</td>
</tr>
<tr>
<td>Electromagnetic torque, (T) [N-m]</td>
<td>Analytical model</td>
<td>27.08</td>
<td>–</td>
<td>27.07</td>
</tr>
<tr>
<td></td>
<td>FE model (64 points/period)</td>
<td>28.14</td>
<td>–</td>
<td>28.12</td>
</tr>
<tr>
<td></td>
<td>FE model (128 points/period)</td>
<td>27.44</td>
<td>–</td>
<td>27.43</td>
</tr>
</tbody>
</table>

Table 5.6: Analytical model and FE comparison during healthy drive operation

### 5.4.2 Winding open-circuit faults

In the analysis of winding open-circuit faults, the three cases previously considered have been studied: the open-circuit of a single phase, the open-circuit of two adjacent phases and the fault of two non-adjacent phases. The steady state results for the phase currents in this three cases are collected in Figure 5.27. The FE simulations have been conducted considering a fixed simulation step size corresponding to 64 points per electrical period. Better results are expected for a smaller step size. The results for the electromagnetic torque are given in Table 5.7. Again, the results of the analytical model are in good agreement with the results obtained by the FE simulations.
Figure 5.26: Steady-state operating curves during healthy drive operation: analytical model (solid line), FEM (marks)

Table 5.7: Electromagnetic torque comparison at winding open-circuit conditions (analytical model vs FE)
5.4. **FINITE ELEMENT VALIDATION**

5.4.3 **Terminal short-circuit faults**

The cases considered in the validation of the terminal short-circuit models are: the short-circuit of a single phase (short-circuit between phase terminal and neutral point), the short-circuit between the terminals of two adjacent phases and the fault between the terminals of two non-adjacent phases. Again, in these two last cases, it has been assumed that the inverter’s protection circuit can instantaneously detect the fault and that the legs corresponding to the faulted phases are turned off in order to avoid a bus shoot-through.

The steady state results for the phase currents are shown in Figure 5.28, whereas the electromagnetic torque calculations are collected in Table 5.8. A good agreement between both calculation methods can be observed as well in observing the obtained results.

<table>
<thead>
<tr>
<th></th>
<th>Analytical</th>
<th>FE</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$T_{mean}$</td>
<td>$T_{max}$</td>
</tr>
<tr>
<td>Single phase open-circuit</td>
<td>27.08</td>
<td>28.50</td>
</tr>
<tr>
<td>Two adjacent phases open-circuit</td>
<td>10.97</td>
<td>23.37</td>
</tr>
<tr>
<td>Two non-adjacent phases open-circuit</td>
<td>14.89</td>
<td>31.81</td>
</tr>
</tbody>
</table>

Note: All values given in [N·m]

Table 5.8: Electromagnetic torque comparison at winding open-circuit conditions (analytical model vs FE)
5.4.4 Inverter faults

The cases considered in the validation of the drive model are the same as the ones studied previously: the permanent opening of a single switch, the permanent opening of both switches in the same leg and the permanent closing of a single switch. As it has been done in previous simulations, in the case of a permanent closing of a switch, it is assumed that the other switch in the same leg is automatically opened to avoid damaging inverter.

The steady state results for the simulations are collected in Figure 5.29 and Table 5.9. As in previous cases, the results of the analytical model are in good agreement with those obtained by the FEM. In the case of a permanent opening of an inverter leg, the ability of the current to flow through the free-wheeling diodes of the faulted leg can be appreciated in 5.29b. It is thus demonstrated that this fault is different from the open-circuit of a phase winding. The amount of current flowing from the machine to the inverter in this case is small.

<table>
<thead>
<tr>
<th>Inverter Fault Condition</th>
<th>Analytical $T_{\text{mean}}$</th>
<th>Analytical $T_{\text{max}}$</th>
<th>Analytical $T_{\text{min}}$</th>
<th>FEM $T_{\text{mean}}$</th>
<th>FEM $T_{\text{max}}$</th>
<th>FEM $T_{\text{min}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permanent opening of a single switch</td>
<td>26.27</td>
<td>41.09</td>
<td>10.14</td>
<td>27.25</td>
<td>42.10</td>
<td>9.89</td>
</tr>
<tr>
<td>Permanent opening of both switches in the same leg</td>
<td>19.86</td>
<td>28.05</td>
<td>10.14</td>
<td>20.76</td>
<td>30.68</td>
<td>9.89</td>
</tr>
<tr>
<td>Permanent closing of a single switch</td>
<td>24.97</td>
<td>265.3</td>
<td>$-238.4$</td>
<td>26.25</td>
<td>266.4</td>
<td>$-238.8$</td>
</tr>
</tbody>
</table>

Note: All values given in [N·m]

Table 5.9: Electromagnetic torque comparison at inverter fault conditions (analytical model vs FEM)
5.5 Conclusions

A 5-phase analytical PMSM drive model suitable for fault analysis has been developed. The model, while simple and fast to simulate, retains the main machine characteristics and is able to accurately predict the behavior of the machine both under healthy and faulty conditions when no iron saturation is considered. The proposed model is general and can account for many different PMSM drive faults. Additionally, the model can be adapted to different control strategies and supply conditions. Different strategies on how to extend the model in order to account for iron saturation and other complex phenomena have been suggested. The validity of the model has been checked with a number of FE simulations; obtaining a good agreement between both simulation methods under all the considered fault scenarios.

Using the developed model as a tool, different fault conditions at rated load and speed have been investigated for the fault-tolerant PMSM drive developed in the present thesis. As a conclusion of the study, the permanent closing of a single inverter switch is the most severe drive fault for the designed prototype. The rest of faults can, up to a point, be accommodated by the current controllers. However, large torque ripples and increased machine losses result from the fault conditions. In order to ensure the safety of the drive, appropriate remedial actions should be taken.

It is expected that, in subsequent work, the developed model serves as a tool to identify fault signatures and test post-fault remedial strategies.
Chapter 6

Fault-Tolerant Operation of a PMSM Drive

In addition to being able to withstand and diagnosis possible faults, a fault-tolerant drive requires that appropriate remedial actions are taken after a fault in order to isolate it, eliminate the hazard and maintain the uninterrupted operation of the drive with the minimum possible performance degradation. In the present chapter, the fault-tolerant operation of the drive composed by the designed PMSM prototype and a 5-phase half-bridge inverter is investigated. Each possible fault scenario is separately studied. For each fault, a literature review on previously proposed remedial actions is presented. Then, the implementation of a simple modified control strategy is discussed and the dynamic fault response of the considered PMSM drive is evaluated with the aid of the model developed on the previous chapter. Finally, the conclusions for the chapter are stated.

6.1 Post-fault actuation in fault-tolerant PMSM drives

The very first step to be accomplished by a fault-tolerant drive subject to a fault is to quickly detect it and perform a fault diagnosis. A comprehensive diagnosis involves the detection of the fault, the determination of its location and the assessment of the fault severity [30]. Numerous works have dealt with the detection of faults in electrical machines in the past. In fact, the amount of published material is huge. A review from 1999, [416], accounts for more than 350 references related to the fault detection and diagnosis of electrical machines, including journal papers, conferences, workshops, books, etc. This can give an idea of the amount of work dedicated to the detection of faults in electrical machines. The detection of inter-turn short circuit faults in induction motors has received most of the attention [30,417–419]. Most of the previously conducted work has been devoted to the condition monitoring of electrical machines [29,59,420] and thus, to the detection of incipient faults like bearing damage and inter-turn short-circuit faults. However, many of the proposed methods can be used in the fault detection of sudden faults as well, like winding open-circuit faults [62,164] and inverter faults [42,43,421,422].

The basic requirements for an adequate diagnosis technique to be implemented in a fault-tolerant drive are that the technique can perform online and that it detects the fault as quickly as possible. Additional desirable features are that it can discriminate among different faults, it does not lead to false alarms and it is not affected by variations in the load conditions, electromagnetic noise or changes in the voltage supply. As they tend to be application specific, the diagnosis method should not rely on threshold values or, at least, the threshold values used should be as universal as possible in order to not depend on the machine’s rated power, operating load and speed conditions, machine type, etc. [423]. The technique should be simple to implement and have low computational requirements. In addition to this, methods that can diagnosis the fault without requiring extra sensors should be preferred.

Owing to the aforementioned requirements, the selection and implementation of a proper diagnosis technique is not easy task and research continues to be done. Generally speaking, no specific technique exists
that can adequately deal with the various possible faults and thus, a combination of different diagnosis methods is usually required [17].

In the present thesis, due to hardware limitations and time constraints, the diagnosis of faults in PMSM drives has not been investigated. It has been assumed that, for each fault, a proper fault diagnostic technique can be implemented in order to detect and locate the fault within a reasonable amount of time and thus, allow for the necessary remedial actions after the fault to be taken. The implementation of a truly fault-tolerant PMSM drive requires that the different faults are adequately diagnosed. Therefore, this matter should be definitely addressed in the future.

Once a fault has been detected, the control system must ensure that the fault does not propagate and negatively affect the voltage supply or the output torque [62]. Therefore, some remedial action has to be taken in order to allow the continued operation of the drive with a minimum impact on its performance. In addition to the diagnosis of the fault, the main steps to be taken by the control system of a fault-tolerant PMSM drive are [20]:

- Isolation of the fault.
- Hardware reconfiguration.
- Software reconfiguration.

Isolating the fault prevents that the fault propagates to other machine components, whereas the reconfiguration of the drive’s hardware is generally aimed at recovering degrees of freedom for machine supply that are lost with the occurrence of the fault. Finally, the objective of the software reconfiguration is to compute suitable current references that minimize the parasitic effects arising from the fault, such as an increase in the torque ripple or the machine losses. If the fault is too severe, as in the case of a permanent closing of an inverter switch, the drive may not operate any further and a safe shut down may be necessary in order to avoid further damage to the system [20].

It must be noted that the specific actions to be taken after a fault are dependent on the drive’s characteristics (machine type, inverter configuration, load and speed conditions, etc.) and the type of fault. For example, not all inverter topologies may allow a reconfiguration of their hardware, whereas the optimal current references to be computed after the fault depend on whether each phase can be independently controlled or not. In this sense, a full H-bridge inverter offers and additional degree of freedom over a half-bridge inverter that can be used to achieve a better drive performance under fault conditions [119]. Depending on the number of phases, different number of faults can be accommodated, which call for different post-fault remedial actions actions to be taken. Finally, the optimal control strategy to be applied after the fault depends on the objective to be achieved. Some applications demand that the machine maintains the same torque capability as in the pre-fault state [62, 69], whereas others focus on eliminating the torque ripple arising from the fault conditions [123, 185]. Generally speaking, it is sought that the fault can be thermally sustained for a long time, meaning that enough provision in terms of temperature increase must be made when designing the machine or that the machine losses operating at fault conditions must be limited to their pre-fault values. According to [21], since operation at fault conditions represents an abnormal state, a reduction in the drive’s efficiency should be a secondary concern as long as the fault can be thermally accommodated. As it will be latter demonstrated, the limits dictated by the current withstand capability of the power semiconductor switches and the available DC bus voltage may impose serious limits to the performance attainable by a PMSM drive operating at fault conditions, as well.

6.2 Remedial strategies

In the following section, the remedial strategies aimed at guaranteeing the fault-tolerance of a PMSM drive are investigated. Specifically, the strategies designed for a 5-phase PMSM supplied by a half-bridge inverter are considered. In particular, the operation of the drive under the following faults is investigated:

- Winding open-circuit faults.
  - Single phase.
6.2. REMEDIAL STRATEGIES

- Two adjacent phases.
- Two non-adjacent phases.

- Terminal short-circuit faults.
  - Single phase.
  - Two adjacent phases.
  - Two non-adjacent phases.

- Inverter faults.
  - Permanent opening of a single inverter switch.
  - Permanent closing of a single inverter switch.

Each fault is separately studied. For each one of the considered faults, previous work regarding the fault-tolerant operation of PMSM drives is reviewed. Since the considered PMSM is supplied by a half-bridge inverter, only a brief discussion regarding machine supply from a full H-bridge inverter is given. Moreover, modified inverter configurations that allow a reconfiguration of the hardware are not investigated in the present chapter. Hence, only methods that involve a software reconfiguration are addressed. In addition to reviewing previously proposed techniques, the application of a control strategy aimed at minimizing the stator copper losses while maintaining the main harmonic of the air-gap MMF waveform is proposed. The general procedure, together with the equations to compute the current references for a number of fault scenarios, is described in Appendix F.

As they are incipient faults that rapidly evolve into other kind of faults [30], inter-turn short-circuit faults have not been considered in the present thesis. As a brief comment, in [16] it has been suggested that the inverter legs corresponding to the faulted phase are switched off in order to isolate the fault, although this would inevitably put the winding isolation at risk due to the high short-circuit current in the shorted turn. Alternatively, some authors have proposed to short-circuit the faulted phase at its terminals in order the MMF of the shorted turn to be spread to the whole phase winding and thus, be limited by the whole phase inductance [62, 66]. In this case, it is necessary that the machine is designed with a high inductance and be supplied by a full H-bridge inverter if some post-fault operation capability must remain. The validity of this procedure has been questioned in [173] for large bar-wound machines, in which the amount of armature flux linkage seen by each conductor heavily depends on its position within the slot. In this case, a more proper method can be to inject a certain amount of current into the faulted winding in order to limit the short-circuit current in the faulted turn. The precise amount current must be calculated for each particular design and fault condition, although, if the position of the faulted turn is not known, a general procedure can be to inject the rated current at 90° lagging with respect to the phase EMF [173]. Another method to limit the fault current, based on operating at flux weakening conditions, has been proposed in [22].

6.2.1 Winding open-circuit faults

Open-circuit of a single phase

The open-circuit fault of a single phase is one of the most studied fault scenarios in the related literature. A number of different remedial strategies have been proposed involving both the hardware and the software reconfiguration of the drive. In the following subsection, besides reviewing previous work on the matter, the application of the strategy described in Appendix F in the case of a 5-phase PMSM drive is evaluated. The proposed remedial strategy is compared to one of the most common modified control strategies and the advantages of using a full H-bridge inverter instead of a half-bridge inverter are discussed. Finally, remarks regarding the actual implementation of any modified control strategy are summarized.

Fault isolation Fault isolation may not be a major concern in the case of winding open-circuit faults, although it is always recommended that the faulted phase is isolated from the DC bus by turning off
the power switches in the corresponding inverter leg [16]. In this sense, it must be noted that a half-bridge inverter supplied machine cannot be entirely isolated after a fault since the phases always remain connected, either by a common neutral point or a delta connection. A full H-bridge is advantageous in this aspect, as the faulted phase can be totally isolated from the DC bus by opening the associated semiconductor switches.

Previously proposed modified control strategies Previously proposed methods regarding operation with winding open-circuit faults fall mainly into two categories: methods that compute the current references to preserve the main harmonic of the air-gap MMF waveform and methods that employ an optimization algorithm to compute optimal instantaneous current references.

The first category of methods, aimed at preserving the main harmonic of the air-gap MMF, has been proposed both for the fault-tolerant operation of PMSM [123, 279] and induction machine drives [31, 96, 165, 424]. The main difference among the distinct references lies in the number of considered phases. The proposed control strategy consists of varying the magnitude and angle of the healthy phase currents phasors in order to maintain the main harmonic of the air-gap MMF with respect to the pre-fault state. In multiphase systems ($m > 3$) in which a single phase has been lost, infinite possibilities arise regarding the choice of phasors. Most usually, the healthy current references are chosen so that the current magnitudes in all the phases are equal [165, 424]. Subsequently, this method is referred to as the “equal currents method”. The application of this control strategy in the fault-tolerant operation of a 5-phase PMSM drive is further investigated on a latter point.

The advantages of the equal currents method are that it is simple to implement, it requires a low number of computations and that the derived formulas are valid for machines of any size or power rating. For machines designed with a sinusoidal back-EMF, the computed current references lead to a smooth torque with almost no ripple, unless the machine operates under significant iron saturation conditions. However, since the calculations are usually carried on a phase representation of the currents, the current references become angle dependent variables. Moreover, the method does not naturally account for the limits imposed by the current withstand capability of the power semiconductor switches or by the available DC bus voltage. Furthermore, since the healthy phase currents are chosen to have the same magnitude, the method cannot be thought of as being optimal in the sense of machine loss or torque ripple reduction. Also, this approach may not be valid for machines with non-sinusoidal back-EMF waveforms [23].

Another family of methods regarding the fault-tolerant operation with winding open-circuit faults involves finding suitable current references by employing an optimization algorithm. In this sense, a number of authors have proposed different algorithms to minimize the stator copper losses while imposing a number of constraints, including those arising by the fault conditions [23, 171, 172, 185, 425]. Since in this methods it is usually imposed that the instantaneous torque equals the demanded torque, the torque ripple can be eliminated even at fault conditions. In all the previous cases, the current references are calculated for each rotor position. Hence, the current references can be considered truly optimal for the specific objective to be achieved. The torque ripple is minimized, machine losses can be kept at minimum and the constraints imposed by the current and voltage limits can be easily addressed; making the methods valid both for the constant-torque and constant-power regions [172, 185]. Moreover, the proposed formulas can be readily applied to different phase number machines and the algorithms can cater for terminal short-circuit faults, in addition to winding open-circuits. Most of the previous references consider the supply from a full H-bridge inverter and hence, independent control of each phase current. The additional constraint imposed by having the machine supplied by a half-bridge inverter is considered in [23], for instance.

However, as the optimal current references depend on the rotor position, the aforementioned methods require that the references are pre-calculated and stored as angle dependent variables in look-up tables. In order to deal with different kind of faults and faults in different phases, a large amount of data may be needed to be stored. Additionally, the methods require that the contribution of each phase current to the electromagnetic torque is perfectly modeled and thus, they may face difficulties when trying to address the effect of iron saturation. For instance, [426] calculates the optimal current references that maximize the torque to copper losses, addressing the effects of iron saturation and reluctance torque, as well as the current limits imposed by the inverter components. However, the method requires that the torque versus position and current characteristic is perfectly known beforehand, which demands extensive simulation and/or experimentation. Furthermore, the computation of optimal currents that produce the
6.2. REMEDIAL STRATEGIES

demanded torque at any time instant can result in relatively large time harmonics in the phase current waveforms. This, coupled to the large space harmonic content of the air-gap MMF distribution in a FSCW, can lead to a significant rotor loss increase; specially in SPM machines [171]. The effects of the fault-tolerant operation with the aforementioned optimal current control strategies on the rotor losses of a SPM machine have been discussed, for instance, in [425].

Alternatively to the previously discussed methods, an analytical method to compute the current references in a faulty 5-phase PMSMs is proposed in [114,123,427]. The method, aimed at minimizing the torque ripple in post-fault operation, is based on choosing the healthy current phasor angles so that specific order harmonics of the electromagnetic torque are eliminated. Three different cases are considered. In the first one, only the fundamental harmonic of the flux density distribution is considered and purely sinusoidal current references are chosen in order to eliminate the second order harmonic of the electromagnetic torque. In the second case, the previously computed phasor references are modified by considering higher order flux density harmonics. Finally, in the third case, third harmonic current references are introduced in order to eliminate both the second order and fourth order torque harmonics. By considering just the fundamental harmonic of the flux density distribution, the equations proposed by [123] lead to the same current references obtained with the equal currents method.

The original method proposed in [123] for machines supplied by a half-bridge inverter is extended in [114] to account for full H-bridge inverter supplied machines. A comparison between the fault-operation of two twin machines, one with a double-layer FSCW and the other with a single-layer FSCW, operated both with the control strategy proposed in [123], is given in [427]. Experimental results show that both motors exhibit a smooth and suitably high torque when the proposed current references are adopted, even operating with faults. A similar approach for a 5-phase PM motor with a trapezoidal back-EMF waveform is described in [174].

As advantages, the proposed formulas are general and valid for any 5-phase PMSM; regardless of its size or electrical parameters. The formulas are quite simple and easy to implement and do not require a large amount of data to be stored as in the methods based on an optimization of the phase currents. However, the formulas proposed are only valid for 5-phase PMSMs. Moreover, as with the previously discussed optimization algorithms, the method requires that the contribution of each phase current to the electromagnetic torque is perfectly modeled and, therefore, it may be difficult to apply under iron saturation conditions. Finally, the constraints imposed by the current withstand capability of the semiconductor switches and the available DC bus voltage have not been addressed.

In addition to PMSMs, some remedial strategies have been proposed for the operation of BLDC machines subject to open-circuit faults as well [428]. For instance, [169] proposes modified switching patterns for BLDC motors operating in the 180° conducting mode. Specific switching patterns, aimed at reducing the torque ripple, are proposed in the case of 3-, 4- and 5-phase machines supplied by a full H-bridge inverter. Different patterns are proposed as well for winding open-circuit faults and inverter open switch faults. In any case, since for a \( m \)-phase BLDC machine, inverter switching is limited to 2\( m \) commutations per cycle, the controlability of the phase currents is poor and the resulting currents and torque waveforms have a significant harmonic content, leading to an increase in the acoustic noise.

In the present thesis, a simple, yet general, modified control strategy is proposed. The method, based on the work by [414], is based on minimizing stator copper losses while maintaining the main harmonic of the air-gap MMF waveform. The generalities of the method are outlined in Appendix F. Instead of choosing the current references so that the healthy phase currents have the same magnitude as in the equal currents method [424], the proposed method selects the current references that preserve the main MMF harmonic while minimizing the stator copper losses. Also, since sinusoidal current references are considered, the optimization algorithm must not be applied for each rotor position, saving computation time and reducing data storage requirements. This method can be thought of a simple version of the optimization algorithms previously mentioned when only the first harmonic torque component is considered. By referring the phase currents to the \( \alpha - \beta \) reference frame, constant relationships between the current components are derived in the case of an open-circuit fault. Hence, the amount of data to be pre-calculated and stored in order to be used in the fault-operation of the drive is minimum. In the following, the operation of the considered PMSM drive with the proposed control strategy is thoroughly discussed.
Fault-tolerant operation with copper loss minimization  The current references to be applied in order to minimize the stator copper losses while maintaining the main harmonic of the armature MMF have been derived in Appendix F. In the case of an AC machine supplied by a half-bridge inverter, as long as an additional path for the zero sequence to follow is not provided, there is no control over the zero sequence current. Therefore, the strategy proposed below involves applying the current references computed from equation (F.12). In the case phase A is open-circuited: \( i_A(t) = i_0(t) = 0 \), and thus, the suitable current references to be followed are:

\[
\begin{align*}
\hat{i}_{\alpha_1} &= \chi \cdot i_{\alpha_1,h} \\
\hat{i}_{\beta_1} &= \chi \cdot i_{\beta_1,h} \\
\hat{i}_{\alpha_3} &= -\chi \cdot i_{\alpha_1,h} \\
\hat{i}_{\beta_3} &= 0 \\
i_0 &= 0 
\end{align*}
\]  

(6.1)

By considering the previous current references with no torque derating (\( \chi = 1 \)), the operating curves displayed in Figure 6.1 are obtained. The figure shows the variation of the main electric variables at steady state and under rated speed and load conditions, together with some reference values (peak or mean) in the pre-fault state (in dash-dot). The curves have been analytically computed by considering the linear machine parameters described in Table 5.1. Hence, no iron saturation and no losses besides stator copper losses have been considered in the computations. The voltage limit displayed in Figure 6.1d corresponds to the voltage limit of a sinusoidal-PWM switching strategy in the linear range (no overmodulation, \( u_{1,max} = \frac{1}{2}U_{DC} \)).

In observing Figure 6.1, it can be concluded that the proposed control strategy is successful in maintaining the mean torque value with a negligible ripple. In order to retain the main armature MMF harmonic, the current increases primarily in the phases adjacent to the faulted phase. Such an increase in the phase currents does not come without incurring a penalty: the phase to neutral voltages in some of the phases also increase and so do stator copper losses. A further effect of the modification of the control law is that, although the current references corresponding to the first harmonic are equal to those of the healthy state \((i_d1, i_q1)\), the third harmonic current references \((i_d3, i_q3)\) experience high and fast fluctuations. These fluctuations are likely to cause an increase in the rotor losses with respect to the pre-fault state and may even be dangerous to the PMs, as both \(i_d3\) and \(i_q3\) currents produced demagnetizing magnetic fields. These last issues are further discussed in a latter section. The tiny torque ripple that is observable in the torque waveform is caused by the higher order harmonics of the PM flux linkage that interact with the higher order current components. In any case, the quality of the resulting torque waveform is remarkable.

Fault-tolerant operation with equal currents  The same analysis as in the previous point has been conducted for the modified control strategy based on keeping the same current magnitudes in the healthy phases [123, 414, 424]. In this case, the introduced current references are:

\[
\begin{align*}
i_A &= 0 \\
i_B &= 1.382\sqrt{2}I \cos (\theta_e + \psi - 1\alpha_m + \pi/5) \\
i_C &= 1.382\sqrt{2}I \cos (\theta_e + \psi - 2\alpha_m) \\
i_D &= 1.382\sqrt{2}I \cos (\theta_e + \psi - 3\alpha_m) \\
i_E &= 1.382\sqrt{2}I \cos (\theta_e + \psi - 4\alpha_m - \pi/5)
\end{align*}
\]  

(6.2)

That is, for a single faulted phase, the strategy consists on scaling all the currents by a factor and shifting the currents in the phases adjacent to the faulted phase towards it. The main objective of modifying the control strategy is the same as in the previous case: to preserve the main harmonic of the air-gap MMF. The difference between both strategies lies in the waveform of the higher order current components. As the results obtained with both methods are very similar, the steady state operating curves for the strategy based on keeping the same current magnitudes in the healthy phases are not presented. A proper comparison between both methods is given in a latter subsection by discussing the results obtained from conducting FE simulations on the designed prototype machine.
Fault-tolerant operation with supply from a full H-bridge inverter  In addition to allowing the faulted phase to be totally isolated, the use of a full H-bridge inverter grants additional degrees of freedom for the current control over the use of a half-bridge inverter. Specifically, by controlling each phase current individually, the zero sequence current can be used in order to minimize the stator copper losses. In this case, by employing the method described in Appendix F, the current references are computed from equation (F.10). In the case phase A is open-circuited, $i_A(t) = 0$:  

Figure 6.1: Steady state operating curves for open-circuit fault with modified control strategy aimed at minimizing stator copper losses (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating ($\chi = 1$)
factor of \( \chi \) machine temperatures below a safety limit. As it can be appreciated in the table, by choosing a derating a second way of dealing with the additional losses is to employ a derating factor in order to keep the derated and if in the design stage the loss increase under fault conditions has not been contemplated, winding and PM temperatures:

\[
\begin{align*}
    i_{\alpha_1} &= \chi \cdot i_{\alpha_1, h} \\
    i_{\beta_1} &= \chi \cdot i_{\beta_1, h} \\
    i_{\alpha_3} &= -2/3 \cdot \chi \cdot i_{\alpha_1, h} \\
    i_{\beta_3} &= 0 \\
    i_0 &= -\sqrt{2}/3 \cdot \chi \cdot i_{\alpha_1, h}
\end{align*}
\]

(6.3)

The steady state operating curves obtained by applying the previous current references with no torque derating \( (\chi = 1) \) are shown in Figure 6.2. As in the previous cases, the considered control strategy is able to deliver the mean torque with a negligible ripple. The main difference with respect to the supply from a half-bridge inverter lies in the waveform of the zero sequence current that is not forced to be zero anymore. This leads to a reduction in the copper losses and in the peak phase to neutral voltage, but not in the peak current values. Interestingly enough, when a full H-bridge inverter is used, the current magnitudes increase most in the phases non-adjacent to the faulted one. The higher order harmonic current references are also lower than in the case of supply from a half-bridge inverter. Therefore, the supply from a full H-bridge inverter is advantageous in a number of ways. In the next section a more detailed comparison between both supply conditions is made.

**FE simulation results**  In the following section, a comparison among the cases studied above is presented based on the results obtained by a number of FE simulations. For each modified control strategy, different torque derating factors \( (\chi) \) have been introduced in order to preserve one of the following variables: the voltage limit imposed by the DC bus \((U_{DC} = 300 \text{ V})\), the peak current in the inverter power switches, the stator copper losses or the mean electromagnetic torque.

In all the cases, a constant drive speed of 600 rpm has been considered. It has been assumed that the current references are perfectly followed and thus, the phase currents are purely sinusoidal. The current references and the torque derating factors have been calculated from the linear machine parameters collected in Table 5.1. Specifically, the following \( d-q \) axis current references have been considered for the machine operating at no fault conditions: \( i_d = 6.703 \text{ A}, \ i_q = 0 \text{ A}, \ i_{\alpha_1} = i_{\alpha_3} = i_{\beta_3} = 0 \text{ A} \). It must be noted that, in order to address the effect of iron saturation in the drive’s response, the FE simulations have been conducted by considering the magnetization curve of the iron material used in the prototype (see Figure 4.9). This fact makes that the rated electromagnetic torque \((T = 27.06 \text{ N-m})\) is not delivered with the considered current references even at no fault conditions.

The stator phase resistance and the magnets remanence have been estimated by assuming the following winding and PM temperatures: \( T_u = 85 \text{ °C}, \ T_{PM} = 75 \text{ °C} \). Iron losses in the machine have been computed by employing the modified Bertotti formulas discussed in Chapter 4 (see equation (4.12)). The loss coefficients have been adjusted for the loss curves given by the supplier \((k_h = 187.0, \ k_{cc} = -0.022, \ k_{exc} = 2.803)\).

The FE simulation results for the control strategy aimed at minimizing the stator copper losses while being supplied from a half-bridge inverter are collected in Table 6.1.

Like the analytically calculated operating curves in Figure 6.1 suggested, the results from the FE simulations demonstrate that the designed machine prototype is able to deliver the demanded torque and power under fault conditions (same torque case, \( \chi = 1.000 \)), with a moderate penalty in terms of electromagnetic losses (increase of 36.92% with respect to the pre-fault state). Such an increase in the machine losses can be reflected in two different ways. If the application the fault-tolerant drive is intended for demands that the rated torque and power are delivered under fault conditions, this objective must be addressed when designing the machine and provide for the corresponding temperature increase due to the additional losses. It should be noticed that, proportionally, the highest loss increase with respect to the pre-fault state occurs in the machine windings (increase of 50.41%), whereas the loss increase in the iron parts and in the PMs is lower than a 7%. If the fault-tolerant application allows the torque to be derated and if in the design stage the loss increase under fault conditions has not been contemplated, a second way of dealing with the additional losses is to employ a derating factor in order to keep the machine temperatures below a safety limit. As it can be appreciated in the table, by choosing a derating factor of \( \chi = 0.816 \), the stator copper losses can be made to be equal to those of the pre-fault state.
6.2. REMEDIAL STRATEGIES

The decrease in the torque capability is approximately that of the derating factor (torque ratio of 0.803). The slight difference in values is due to the effect of non-linearities in the magnetization curve of the iron lamination material. The derating factor aimed at maintaining the copper losses can be analytically computed and is valid for any machine operating under any conditions. On the other hand, in order to ensure that the total machine losses are preserved, the required derating factor ought to be calculated for each particular machine and torque/speed conditions. However, since for the designed prototype, stator copper losses are the dominant loss component and the main additional losses under fault conditions are generated within the machine windings, there is little difference between trying to preserve the stator copper losses or the total electromagnetic losses. In any case, it must be noted that the increase in copper losses is not the same in all the phases. The faulted machine does not generate copper losses and thus, the temperature distribution in the machine will not be uniform. A complete thermal model of the machine that addressed each phase separately would be necessary in order to have a proper estimation.
of the temperature variation within the machine at fault conditions.

In addition to the increase in machine losses, other factors regarding the operation of the machine under fault conditions like the peak current flowing through the power semiconductor switches and the voltage limit imposed by the DC bus must be addressed. This last issue deserves special attention in the present case as, as it has been previously commented on Chapter 5, the considered PMSM drive has been designed with a tiny voltage margin. In Table 6.1 the results obtained by imposing the necessary derating factors in order to keep the post-fault voltage (\(\chi = 0.504\)) and current (\(\chi = 0.681\)) magnitudes are shown as well. The considered voltage limit is that of a sinusoidal-PWM switching strategy operating in the linear region (no overmodulation, \(u_{1,\text{max}} = \frac{1}{2}U_{\text{DC}}\)). As it can be appreciated from the values of the necessary derating factors, the aforementioned conditions impose serious limitations to the mean torque deliverable by the drive. In particular, the highest restriction is that of the voltage limit that imposes that the torque capability of the drive is reduced to a 51.99\% of its pre-fault value.

The results included in the last row of Table 6.1 indicate that the introduction of non-zero currents in the \(d_2\) and \(q_2\)-axis of the machine lead to demagnetizing magnetic fields that reduce the flux density levels in the PMs. However, for the considered prototype, and for the investigated fault scenario, the flux densities in the PMs are well over the demagnetization limit and thus, no risks regarding PM demagnetization are expected from operation with a single phase open-circuit fault.

Next, the FE simulation results obtained for the control strategy aimed at equalizing the current magnitudes and the results for the supply from a full H-bridge inverter are presented. Specifically, the simulation results for the control strategy with equal current magnitudes are collected in Table 6.2, whereas the results for the supply from a full H-bridge inverter are shown in Table 6.3.

### Table 6.1: FE simulation results for single phase winding open-circuit fault with modified control and different derating factors. Control aimed at minimizing stator copper losses with supply from a half-bridge inverter

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ((\chi = 0.504))</th>
<th>Same current ((\chi = 0.681))</th>
<th>Same copper losses ((\chi = 0.816))</th>
<th>Same torque ((\chi = 1.000))</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak branch current ([A])</td>
<td>4.959</td>
<td>6.703</td>
<td>8.027</td>
<td>9.838</td>
<td>6.703</td>
</tr>
<tr>
<td>Peak voltage, (u_{ph,N,\text{max}}) ([V])</td>
<td>149.9</td>
<td>170.3</td>
<td>186.3</td>
<td>208.4</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, (T_{\text{mean}}) ([N\cdot m])</td>
<td>13.44</td>
<td>18.04</td>
<td>21.43</td>
<td>25.85</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, (\Delta T) ([N\cdot m])</td>
<td>0.269</td>
<td>0.842</td>
<td>1.510</td>
<td>2.762</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, (P_{Cu}) ([W])</td>
<td>41.62</td>
<td>76.03</td>
<td>109.0</td>
<td>163.8</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, (P_{Fe,t}) ([W])</td>
<td>25.64</td>
<td>27.57</td>
<td>29.19</td>
<td>31.53</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, (P_{Fe,y,u}) ([W])</td>
<td>6.890</td>
<td>8.142</td>
<td>9.246</td>
<td>10.91</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, (P_{Fe,y,r}) ([W])</td>
<td>0.213</td>
<td>0.324</td>
<td>0.415</td>
<td>0.546</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, (P_{PM}) ([W])</td>
<td>0.148</td>
<td>0.266</td>
<td>0.375</td>
<td>0.549</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, (P_{\text{loss}}) ([W])</td>
<td>74.51</td>
<td>112.3</td>
<td>148.3</td>
<td>207.3</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density ([T])</td>
<td>0.964</td>
<td>0.941</td>
<td>0.923</td>
<td>0.898</td>
<td>0.959</td>
</tr>
</tbody>
</table>

### Table 6.2: FE simulation results for single phase winding open-circuit fault with modified control and different derating factors. Control with equal current magnitudes and supply from a half-bridge inverter

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ((\chi = 0.510))</th>
<th>Same current ((\chi = 0.724))</th>
<th>Same copper losses ((\chi = 0.809))</th>
<th>Same torque ((\chi = 1.000))</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak branch current ([A])</td>
<td>4.725</td>
<td>6.703</td>
<td>7.490</td>
<td>9.264</td>
<td>6.703</td>
</tr>
<tr>
<td>Peak voltage, (u_{ph,N,\text{max}}) ([V])</td>
<td>149.7</td>
<td>173.4</td>
<td>183.0</td>
<td>204.7</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, (T_{\text{mean}}) ([N\cdot m])</td>
<td>13.59</td>
<td>19.14</td>
<td>21.26</td>
<td>25.80</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, (\Delta T) ([N\cdot m])</td>
<td>0.260</td>
<td>0.934</td>
<td>1.347</td>
<td>2.794</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, (P_{Cu}) ([W])</td>
<td>43.39</td>
<td>87.34</td>
<td>109.0</td>
<td>166.8</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, (P_{Fe,t}) ([W])</td>
<td>25.74</td>
<td>28.18</td>
<td>29.21</td>
<td>31.50</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, (P_{Fe,y,u}) ([W])</td>
<td>6.959</td>
<td>8.546</td>
<td>9.258</td>
<td>10.95</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, (P_{Fe,y,r}) ([W])</td>
<td>0.220</td>
<td>0.358</td>
<td>0.417</td>
<td>0.553</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, (P_{PM}) ([W])</td>
<td>0.155</td>
<td>0.305</td>
<td>0.376</td>
<td>0.556</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, (P_{\text{loss}}) ([W])</td>
<td>76.47</td>
<td>124.7</td>
<td>148.3</td>
<td>210.4</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density ([T])</td>
<td>0.967</td>
<td>0.941</td>
<td>0.930</td>
<td>0.906</td>
<td>0.959</td>
</tr>
</tbody>
</table>
6.2. REMEDIAL STRATEGIES

The results for the equal currents method are slightly better than the copper loss minimization method in terms of necessary voltage and current overrating, but slightly worse regarding machine losses. The differences in mean torque capability, torque ripple and PM demagnetization figures are insignificant. The main advantage of using the equal currents method over the copper loss minimization method lies in that, with the former method, machine losses are equally distributed among the remaining healthy phases and thus, a more balance post-fault operating condition is reached. Anyhow, the differences are not that big and other factors such as the ease of implementation may pay a more significant role in the choice of one control strategy over the other [414].

In the case of supply from a full H-bridge inverter, the biggest advantages with respect to the supply from a half-bridge inverter are the reduction in the stator copper losses and the necessary voltage overrating. When no torque derating is considered \((\chi = 1.000)\), machine losses operating at open-circuit conditions can be reduced by a 8.83% by switching from a half-bridge inverter to a full H-bridge inverter. Similarly, the deliverable electromagnetic torque when the voltage limit set by the DC bus is considered can be increased by a 29.02%. The use of a full H-bridge inverter offers no advantages regarding the necessary current overrating of the power semiconductor switches. Due to the lower magnitude third harmonic current components, rotor losses and PM demagnetization effects are slightly lower when a full H-bridge inverter is considered.

As a general conclusion for the reviewed methods, the 3 control strategies described above fulfill the requirements of producing the desirable electromagnetic torque with little degradation of the machine operating figures. A higher torque ripple than in the pre-fault state is obtained in all the three cases as a result of the high and fast fluctuations of the third harmonic current components \((i_d, i_q)\). Anyhow, for the considered machine prototype, the resulting torque waveform quality is more than acceptable with any of the discussed methods. Although the third harmonic current components produce demagnetizing fields within the machine, the flux density values in the PMs are well over the safety limits and no risks are expected for the machine operating with single phase open-circuit faults. The large air-gap of the designed machine prototype helps reduce the rotor loss increase due to the higher order harmonic current components.

Finally, in order to demonstrate the adequacy of the proposed fault-tolerant control strategy, the obtained FE simulation results are compared to those obtained when no remedial action after a fault is taken. The considered phase current waveforms are the ones obtained in Chapter 5 when a standard \(d - q\) current vector control is applied (see Figure 5.10a). The results comparing both post-fault scenarios, together with the data corresponding to the healthy state, are collected in Table 6.4.

Based on the obtained results, it can be established that the main difference between using a \(d - q\) current control without modifying the control law and using the proposed control strategy lies in the higher torque pulsations at open-circuit conditions in the former case. Higher peak currents and phase to neutral voltages are required when the control strategy is not modified. This operation mode leads to lower copper losses but higher stator iron and rotor losses. It must be noted that, although the total rotor
loss increase in this case is low (0.239 W), the percentage increase is significant (22.98%); meaning that this may be an issue for other machine designs. In any case, for the considered machine prototype, the main reason that justifies the adoption of a modified post-fault control strategy is the vast improvement in the torque waveform quality that is obtained over the standard $d-q$ vector control with no modification.

**Dynamic response of the PMSM drive** In order to check the adequate post-fault operation of the PMSM drive with the proposed modified control strategy, the dynamic response of the drive has been evaluated with the aid of the analytical model developed in Chapter 5. In the analysis, the same assumptions and simplifications as in the previous chapter have been considered. Specifically, no iron saturation and no losses besides stator copper losses have been contemplated. The simulation has been conducted assuming an infinite inertia load and thus, a constant drive speed of 600 rpm even after the fault. The constant machine parameters introduced in the model are those collected in Table 5.1.

Instead of the standard current vector control in the synchronously rotating $d-q$ reference frame, a current control loop in the $\alpha-\beta$ axes has been employed. The current control has been implemented by using PI controllers with anti-windup control plus a feed-forward voltage compensation term to improve the system’s response (see Appendix G). In particular, the control structure shown in Figure G.1 has been considered. The voltage limit set for the anti-windup control is that of the maximum output voltage for a sinusoidal PWM strategy in the linear range: $u_{1,max} = 1/2 \cdot U_{DC}$ (no overmodulation). No current limitation has been considered for the current controllers. Since the current control is aimed to operate in both healthy and faulty conditions, the PI control associated to the zero sequence current has been implemented as well.

The most notable difference between the current control in the $d-q$ reference frame and the control in the $\alpha-\beta$ reference frame is that, when an $\alpha-\beta$ control is used, a higher bandwidth for the current controllers must be selected [414]. The reason for this is that, for a healthy PMSM operating at steady state conditions, the current references referred to the $d$- and $q$-axes are constant value parameters, whereas if referred to the $\alpha-\beta$ reference frame, they reflect as sinusoidally varying temporal variables. The parameters for the current controllers considered in the present chapter are collected in Table 6.5.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>No control modification</th>
<th>Copper loss minimization</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak branch current [A]</td>
<td>11.72</td>
<td>9.838</td>
<td>6.703</td>
</tr>
<tr>
<td>Peak voltage, $u_{ph,N,max}$ [V]</td>
<td>238.1</td>
<td>208.4</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, $T_{mean}$ [N·m]</td>
<td>26.25</td>
<td>25.85</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, $\Delta T$ [N·m]</td>
<td>22.80</td>
<td>2.762</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, $P_{Cu}$ [W]</td>
<td>154.3</td>
<td>163.8</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, $P_{Fe,t}$ [W]</td>
<td>34.69</td>
<td>31.53</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, $P_{Fe,y,s}$ [W]</td>
<td>12.13</td>
<td>10.91</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, $P_{Fe,y,r}$ [W]</td>
<td>0.618</td>
<td>0.546</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, $P_{PM}$ [W]</td>
<td>0.661</td>
<td>0.549</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, $P_{loss}$ [W]</td>
<td>202.4</td>
<td>207.3</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density [T]</td>
<td>0.896</td>
<td>0.898</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Table 6.5: Parameters of the PI current controllers. Current control in the $\alpha-\beta$ reference frame

The simulation results for the dynamic response of the PMSM drive after a single phase open-circuit fault are displayed in Figure 6.3. In the simulation, phase A has been open-circuit at a time corresponding
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to the 5th electrical period, whereas the modified control strategy has been applied from the 10th period onwards. Anticipating the voltage overrating required by the PMSM drive in order to deliver the rated torque under fault conditions, a DC bus voltage of 420 V has been considered. This corresponds approximately to twice the peak voltage obtained in the analytical calculation of the steady state operating characteristics in Figure 6.1. No torque derating has been considered in the computation of the current references ($\chi = 1$). For the sake of clarity, only the phase current and torque waveforms are displayed.

Figure 6.3: Drive’s transient response to a single phase open-circuit with modified control strategy aimed at minimizing stator copper losses and supply from a half-bridge inverter (constant speed of 600 rpm)

The previous figure clearly demonstrates the importance of the available DC bus voltage in the fault-tolerant operation of the drive. In case the bus voltage is kept at $U_{DC} = 300$ V, the mean torque delivered after the control strategy is updated is of 16.65 N·m. This value approximately corresponds to the torque to be obtained analytically if the derating factor for the voltage limit to be enforced was considered ($\chi = 0.504$). It must be noted that the drive’s response after the fault is different in the present case from when a $d-q$ control was used (see Chapter 5). By comparing the results in Figures 5.9 and 6.3, it can be stated that the $d-q$ current control performs better in terms of after fault torque capability, at the expense of increasing the current magnitudes in the healthy phases. Anyhow, the previous figure evidences the adequacy of modifying the control strategy after the occurrence of a fault.

Additional remarks In addition to the limits imposed by the available DC bus voltage, the current withstand capability of the power semiconductor switches and the increase in machine losses at fault conditions, other aspects must be addressed when selecting a suitable post-fault control strategy. The proposed current control strategy, aimed at minimizing stator copper losses, is specially well suited to be implemented by referring the phase currents to the $\alpha-\beta$ reference frame. Please note that, if the computed current references were referred to the synchronous $d-q$ reference frame, as it is usually done in the current vector control of PMSM drives, the third harmonic current references would vary rapidly and thus, would require a higher bandwidth current controller [414].

The number of computations and the amount of pre-calculated data needed by the proposed control strategy is low. If the machine is supplied by a half-bridge inverter, only the linear coefficients upon which the $i_{a3}$ and $i_{b3}$ current references depend must be stored. These can be easily pre-calculated and stored in look-up tables depending on the type of fault and on the faulted phase index.

Finally, although it has been demonstrated that the supply from a full H-bridge inverter can be advantageous in a number of ways, in the forthcoming fault scenarios, in order not to lengthen the chapter excessively and to stick to the drive configuration physically available for experimentation, only the supply from a half-bridge inverter has been considered. It must be addressed, though, that the proposed method for computing suitable current references aimed at minimizing stator copper losses is general and valid for the different supply conditions.
Open-circuit of two adjacent phases

In the following section the fault-tolerant operation of a 5-phase PMSM subject to the open-circuit fault of two adjacent phases is studied. The same scheme as in the previous section is followed, although the operation with a full H-bridge inverter is not thoroughly addressed.

Fault isolation Like in the case of a single phase open-circuit fault, the most straightforward way to isolate the faulted phase windings from the DC bus is by permanently opening the corresponding power switches.

Previously proposed modified control strategies Most of the methods proposed for the fault-tolerant operation of PMSM drives with single phase winding open-circuits have been applied with a higher number of faulted phases. For instance, [165] proposes different current references for 5-phase machines under the loss of one, two or three phases. In all the cases, the objective is to preserve the main harmonic of the MMF waveform. In [23], an algorithm to compute the optimal current references is proposed and it is employed with the loss of one or two phases. Finally, the analytical methods proposed in [123, 174] have been used to control machines with two phases open-circuited as well; either adjacent or non-adjacent.

It is important to note that, for a PMSM supplied by a half-bridge inverter in the case of loss of two phases, only one way of maintaining the main MMF harmonic is possible. From the 5 degrees of freedom for machine supply that a 5-phase system allows, two are lost due to the fault conditions, another one is eliminated due to the star connection and the remaining two are used in the preservation of the main MMF harmonic, both in magnitude and angle. Therefore, the healthy phase current waveforms are totally determined and no strategy aimed at equalizing the currents in the healthy phases while preserving the MMF is possible. Owing to this, the solutions proposed by [165, 279, 424] for the fault-tolerant operation of a machine with two open-circuited phases are equivalent to the current references computed by the method considered in the present thesis.

Fault-tolerant operation with preservation of the main air-gap MMF harmonic The current references to be applied in order to preserve the main air-gap MMF harmonic in the case of loss of two phase currents are given in Appendix F. In this case, since the machine is supplied by a half-bridge inverter, $i_0(t) = 0$ and the current references are computed by solving the linear system given by equations (F.19), (F.24) and (F.25). For instance, in case phases $C$ and $D$ are open-circuited in a 5-phase machine:

$$
\begin{align*}
  i_{\alpha_1} &= \chi \cdot i_{\alpha_1,h} \\
  i_{\beta_1} &= \chi \cdot i_{\beta_1,h} \\
  i_{\alpha_3} &= 2.6180 \cdot \chi \cdot i_{\alpha_1,h} \\
  i_{\beta_3} &= -0.6180 \cdot \chi \cdot i_{\beta_1,h} \\
  i_0 &= 0
\end{align*}
$$

By considering the previous current references with no torque derating ($\chi = 1$), the steady state operating curves displayed in Figure 6.4 are analytically calculated for the designed machine prototype.

In observing Figure 6.4, it can be concluded that, in order to maintain the mean torque value, a high overrating is required in terms of phase to neutral voltages, peak healthy phase currents and stator copper losses. The preservation of the main MMF harmonic is achieved at the expense of greatly increasing the current magnitudes in the healthy phases; mainly in the phase electrically most separated to the faulted phases. By comparing the magnitudes of the healthy phase currents, it can be demonstrated that the proposed method is equivalent in this case to the methods discussed in [165, 424] and to the method proposed by [123] when only the first harmonic of the flux density distribution is considered. The resulting third harmonic current components have even higher magnitudes than the first harmonic current references. As a consequence, the resulting torque ripple is higher than in the case of a single phase open-circuit. The influence of these high and pulsating third harmonic currents on the machine losses and on the PM flux density levels is checked in the next point.
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(a) Phase currents (solid), healthy state peak value (dash-dot)

(b) α – β currents (solid), healthy state peak value (dash-dot)

(c) d – q currents

(d) Phase to neutral voltages (solid), \( u_{1,\text{max}} = \frac{1}{2}U_{DC} \) (dash-dot)

(e) Electromagnetic torque (solid), healthy state mean value (dash-dot)

(f) Stator copper losses (solid), healthy state mean value (dash-dot)

Figure 6.4: Steady state operating curves for two adjacent phases open-circuit fault with modified control strategy aimed at maintaining the main air-gap MMF harmonic (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating (\( \chi = 1 \)).

**FE simulation results**  The FE simulation results obtained for the proposed control strategy with different derating factors in the case of a two adjacent phases open-circuit fault are collected in Table 6.6. In the FE simulations, the same conditions as in the case of a single phase open-circuit fault have been considered.

As a conclusion of the conducted FE simulations, very high voltage and current overratings must be provided in order for the PMSM drive to be able to deliver the rated torque with a two adjacent phases open-circuit fault. In fact, due to the iron saturation, the mean electromagnetic torque is significantly lower than in the pre-fault case even if no torque derating is considered (\( \chi = 1.000 \)). A further effect of the non-linearities in the magnetization curve of the material is that the torque ripple increases considerably compared to the analytically calculated value. Although the stator iron losses only increase marginally at fault conditions, the percent increase in rotor losses is noteworthy (27.21%). Moreover, stator copper
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Table 6.6: FE simulation results for two adjacent phases winding open-circuit fault with modified control and different derating factors. Control aimed at preserving the main air-gap MMF harmonic with supply from a half-bridge inverter

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ( (\chi = 0.247) ) [V]</th>
<th>Same current ( (\chi = 0.276) ) [A]</th>
<th>Same copper losses ( (\chi = 0.465) ) [W]</th>
<th>Same torque ( (\chi = 1.000) ) [N.m]</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak branch current</td>
<td>5.999</td>
<td>6.703</td>
<td>11.27</td>
<td>24.25</td>
<td>6.703</td>
</tr>
<tr>
<td>Peak voltage, ( u_{phN,\text{max}} )</td>
<td>146.9</td>
<td>152.6</td>
<td>189.6</td>
<td>337.4</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, ( T_{\text{mean}} ) [N.m]</td>
<td>6.579</td>
<td>7.336</td>
<td>11.72</td>
<td>19.82</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, ( \Delta T ) [N.m]</td>
<td>0.297</td>
<td>0.404</td>
<td>2.408</td>
<td>12.79</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, ( P_{Cu} ) [W]</td>
<td>30.94</td>
<td>38.63</td>
<td>109.2</td>
<td>505.7</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, ( P_{Fe,t} ) [W]</td>
<td>24.42</td>
<td>24.75</td>
<td>26.61</td>
<td>32.59</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, ( P_{Fe,y,s} ) [W]</td>
<td>6.126</td>
<td>6.322</td>
<td>7.607</td>
<td>12.16</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, ( P_{Fe,y,r} ) [W]</td>
<td>0.129</td>
<td>0.150</td>
<td>0.293</td>
<td>0.693</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, ( P_{PM} ) [W]</td>
<td>0.059</td>
<td>0.073</td>
<td>0.187</td>
<td>0.630</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, ( P_{\text{loss}} ) [W]</td>
<td>61.68</td>
<td>69.92</td>
<td>143.9</td>
<td>551.7</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density</td>
<td>0.954</td>
<td>0.946</td>
<td>0.894</td>
<td>0.787</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Losses are dramatically increased unless quite a large torque derating is considered. A further effect of the large phase current magnitudes is that the PM flux density levels are substantially reduced, although, for the designed prototype, they remain well over the demagnetization limit value.

Serious restrictions are placed on the drive’s torque capability if the voltages, currents or losses are limited to their pre-fault values. In case the limit imposed by the available DC bus voltage is considered \( (U_{DC} = 300 \text{ V}) \), the mean torque falls to a 24.65% of its pre-fault value, whereas if the current magnitudes are preserved, the drop in torque capability is that of a 27.49%. Therefore, it is very unlikely that a PMSM drive can be made fault-tolerant to the open-circuit fault of two adjacent phases without derating its torque capability and without incurring in huge penalties in terms of size and cost.

Although not shown in here, vast improvements can be obtained by switching the machine supply from a half-bridge inverter to a full H-bridge one. Just as a means of example, the peak current value when no derating is considered \( (\chi = 1) \) is reduced from 24.25 A to 14.07 A; whereas the peak voltage drops by a 36.44%. Likewise, a high reduction in the copper losses can be obtained when a full H-bridge is used; reducing the losses by ratio of 0.376 compared to the supply from a half-bridge.

Dynamic response of the PMSM drive As in the previous case, the transient response of the fault-tolerant PMSM drive after the occurrence of the fault has been checked with the aid of the developed drive model. The simulation results for the dynamic response of the drive are collected in Figure 6.5. In this case, a DC bus voltage of 700 V has been considered. In the simulation phases \( C \) and \( D \) have been open-circuited at a time corresponding to the \( 5^{th} \) electrical period. The modified control strategy aimed at maintaining the main air-gap MMF harmonic is introduced at a time corresponding to the \( 10^{th} \) electrical period.

The simulation results confirm the vast improvement in the quality of the resulting torque waveform when the control strategy is modified. Such an improvement is obtained for both a DC bus voltage of 300 V and 700 V. However, in the former case, the drive is not able to deliver the rated torque and a high ripple results from the saturation of the current controllers.
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Figure 6.5: Drive’s transient response to the open-circuit of two adjacent phases with modified control strategy aimed at maintaining the main air-gap MMF harmonic (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating ($\chi = 1$)

**Open-circuit of two non-adjacent phases**

The last open-circuit fault considered is the winding open-circuit of two non-adjacent phases. As with previous fault scenarios, the present subsection examines the fault-tolerant operation of the drive with the current references derived in Appendix F. Then, a FE validation is conducted and the dynamic response of the drive is checked with the help of the developed PMSM drive model. As previous work regarding operation with two faulted phases has already been covered in the previous subsection, it is not further addressed here.

**Fault-tolerant operation with copper loss minimization** In this case, it is assumed that the open-circuit fault involves phases B and E. Therefore, the current references, computed by the formulas in Appendix F for a 5-phase machine, are:

$$
\begin{align*}
    i_{\alpha_1} &= \chi \cdot i_{\alpha_1,h} \\
    i_{\beta_1} &= \chi \cdot i_{\beta_1,h} \\
    i_{\alpha_3} &= 0.3820 \cdot \chi \cdot i_{\alpha_1,h} \\
    i_{\beta_3} &= 1.6180 \cdot \chi \cdot i_{\beta_1,h} \\
    i_0 &= 0
\end{align*}
$$

(6.5)

The steady state operating curves obtained by applying the previous current references with no torque derating ($\chi = 1$) are shown in Figure 6.6. As in the previous case, in order to maintain the main harmonic of the air-gap MMF, the current magnitudes in the healthy phases must increase significantly with respect to their pre-fault values. However, the necessary current and voltage overratings to deliver the rated torque are much lower than in the case the faulted phases are adjacent. In the present case, the fault conditions are not as bad in terms of third harmonic current magnitudes and, hence, the resulting torque ripple and stator copper losses are lower. Again, the phase current waveforms and the relationship between the current magnitudes are equal to those proposed by [165] or [123] when only the first harmonic of the flux density distribution is considered.
CHAPTER 6. FAULT-TOLERANT OPERATION OF A PMSM DRIVE

(a) Phase currents (solid), healthy state peak value (dash-dot)

(b) $\alpha - \beta$ currents (solid), healthy state peak value (dash-dot)

(c) $d - q$ currents

(d) Phase to neutral voltages (solid), $u_{1,\text{max}} = \frac{1}{2}U_{DC}$ (dash-dot)

(e) Electromagnetic torque (solid), healthy state mean value (dash-dot)

(f) Stator copper losses (solid), healthy state mean value (dash-dot)

Figure 6.6: Steady state operating curves for two non-adjacent phases open-circuit fault with modified control strategy aimed at maintaining the main air-gap MMF harmonic (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating ($\chi = 1$)

**FE simulation results**  In the same way as for previous cases, a number of FE simulations have been conducted in order to evaluate the performance of the designed PMSM drive with the proposed control strategy. The simulation results for different derating factors are collected in Table 6.7.

Compared to the case in which the faulted phases are adjacent, the present fault scenario is less detrimental to the operation of the drive in all respects. Lower current and voltage overratings are required in order to deliver the mean torque, increase in machine losses is far lower and there is also a lower mean torque decrease due to iron saturation. Additionally, the faulted conditions lead to a lower torque ripple and to higher PM flux density levels. Anyhow, the percent increase in machine losses is still high and large provisions must be made in terms of available voltage and current withstand capability for the drive to deliver the rated torque at fault conditions. When no torque derating is considered ($\chi = 1.000$), the percent increase in stator losses is of 101.1%, whereas rotor losses increase by a 8.85%. As with previous
6.2. REMEDIAL STRATEGIES

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit</th>
<th>Same current</th>
<th>Same copper losses</th>
<th>Same torque</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>((\chi = 0.315))</td>
<td>((\chi = 0.447))</td>
<td>((\chi = 0.649))</td>
<td>((\chi = 1.000))</td>
<td></td>
</tr>
<tr>
<td>Peak voltage, (u_{ph,N,\text{max}}) [V]</td>
<td>149.2</td>
<td>172.9</td>
<td>209.6</td>
<td>276.0</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, (T_{\text{mean}}) [N\cdot m]</td>
<td>8.379</td>
<td>11.78</td>
<td>16.69</td>
<td>24.18</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, (\Delta T) [N\cdot m]</td>
<td>0.260</td>
<td>0.974</td>
<td>2.940</td>
<td>7.339</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, (P_{Cu}) [W]</td>
<td>25.62</td>
<td>51.68</td>
<td>108.7</td>
<td>258.4</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, (P_{Fe,t}) [W]</td>
<td>24.35</td>
<td>25.56</td>
<td>27.83</td>
<td>32.39</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, (P_{Fe,y,s}) [W]</td>
<td>6.998</td>
<td>6.876</td>
<td>8.375</td>
<td>11.53</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, (P_{Fe,y,r}) [W]</td>
<td>0.125</td>
<td>0.202</td>
<td>0.333</td>
<td>0.579</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, (P_{PM}) [W]</td>
<td>0.066</td>
<td>0.128</td>
<td>0.257</td>
<td>0.553</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, (P_{\text{loss}}) [W]</td>
<td>56.25</td>
<td>84.45</td>
<td>145.5</td>
<td>303.4</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density [T]</td>
<td>0.964</td>
<td>0.937</td>
<td>0.895</td>
<td>0.821</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Table 6.7: FE simulation results for two non-adjacent phases winding open-circuit fault with modified control and different derating factors. Control aimed at preserving the main air-gap MMF harmonic with supply from a half-bridge inverter.

Dynamic response of the PMSM drive

The dynamic response of the PMSM drive after the proposed control strategy is applied is displayed in Figure 6.7. In this case, a DC bus voltage of 550 V has been considered. The open-circuit of phases B and E has been contemplated in the simulation. The simulation demonstrates that a significant improvement in the torque waveform is achieved by modifying the control strategy in an attempt to preserve the main MMF harmonic. As expected, the voltage required to deliver the rated torque increases with respect to the pre-fault state.

![Figure 6.7: Drive’s transient response to the open-circuit of two non-adjacent phases with modified control strategy aimed at maintaining the main air-gap MMF harmonic (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating (\(\chi = 1\))](image)

Conclusions for winding open-circuit faults

The fault-tolerant operation of a 5-phase PMSM subject to winding open-circuit faults has been investigated. In addition to conducting a small literature review on proposed remedial actions to mitigate the parasitic effects caused by these faults, the application of a modified control strategy aimed at minimizing the stator copper losses has been discussed. A number of simulations using a commercial FE software and the drive model developed in the present thesis have demonstrated the capability of the proposed method to deliver the demanded torque under different fault conditions. However, in order to maintain the mean torque with respect to the pre-fault state, large provisions must be generally made in terms of available DC bus voltage, current withstand capability of the power semiconductor switches and machine loss increase.
The fault scenario involving the loss of two adjacent phases is the most detrimental to the performance of the drive. The dramatic negative consequences derived from the fault can be alleviated up to a point by using a full H-bridge inverter instead of a half-bridge one [114].

It must be noted that, when no voltage overrating is considered ($U_{DC} = 300$ V), a higher torque capability after the fault remains when a standard $d-q$ current vector control with no post-fault modification is used, as the results from Chapter 5 suggest (see Table 5.3). The increase in copper losses in this case, however, is significant and the resulting torque waveform is pulsating, with a dominant second order harmonic component. For the considered machine prototype, the main reason that justifies the adoption of a modified post-fault control strategy after the fault is the vast improvement that can be obtained in the torque waveform quality.

6.2.2 Terminal short-circuit faults

The fault-tolerant operation of a 5-phase PMSM drive under short-circuit faults is investigated next. Specifically, the drive’s response after the short-circuit of a whole phase winding and the short-circuit between the terminals of two phases, either adjacent or non-adjacent, is examined.

Short-circuit of a single phase

In addition to reviewing previous work regarding the fault-tolerant operation of an AC drive subject to this kind of fault, the application of the modified control strategy aimed at minimizing the stator copper losses is evaluated in the following subsection. For such a strategy, two distinct cases are considered: in the first one, the faulted phase is partially isolated by shutting down the corresponding inverter leg, whereas, in the second one, the neutral point voltage is controlled in an attempt to improve the drive’s performance. Both methods are compared to another technique proposed in the related literature. Finally, the applicability of the considered remedial strategies is checked by conducting a number of FE simulations and by analyzing the dynamic response of the PMSM drive when the control strategy is modified.

Fault isolation If a half-bridge inverter is considered, the faulted phase can be partially isolated by shutting down the corresponding inverter leg, forcing the branch current in the faulted phase to be zero. The fault cannot be truly isolated unless a full H-bridge is used. In any case, for a single phase terminal short-circuit fault, the controlability of the faulted phase current is lost. In such a case, the current is only limited by the phase resistance and the inductance. In the case of a fault-tolerant PMSM, it is assumed that the phase self-inductance is high enough to limit the faulted phase current magnitude to that of the rated current.

As it will be latter discussed, another approach when dealing with single phase terminal short-circuit faults is to leave the faulted phase connected and use the switching in the inverter leg to control the neutral point voltage. This way, control over the zero sequence current is achieved and an additional degree of freedom for current control is gained.

Previously proposed modified control strategies Generally speaking, the methods applied in the fault-tolerant operation of PMSM drives with winding open-circuit faults can be used in the post-fault control of terminal short-circuit faults as well. In this case, instead of having one or more phases in which the current is forced to zero, the faulted phase currents are driven by the PM flux linkage and, hence, are time dependent variables that depend on the motor parameters and on the drive’s speed. In any case, the ability to control the currents in the faulted phases is highly reduced and suitable current references for the remaining healthy phases must be computed in order to reduce the parasitic effects arising from the fault.

The previously discussed methods that compute the current references by solving an optimization algorithm have been already employed in the case of terminal short-circuit faults [171, 425]. The operation with terminal short-circuit faults in both the constant torque and constant power regions has been addressed in [172, 185]. The proposed algorithms compute the optimal currents that minimize the stator
copper losses while delivering the demanded torque by considering the short-circuit as a non-controllable, additional constraint. This approach is totally reasonable for fault-tolerant FSCW machines, for which the mutual inductance between phases tends to be negligible. Although the short-circuit current cannot be controlled by varying the current references in the healthy phases, it is limited by the high phase self-inductance and therefore, does not pose a risk to the operation of the machine. An alternative method is proposed in [429] for machines with non-negligible mutual inductances between phases. By computing suitable current references in the healthy phases, the flux linked by the faulted phase is kept constant and hence, the short-circuit current is mitigated.

Finally, reference [123] proposes analytical formulas for the calculation of the current references in 5-phase drives in the case of a terminal short-circuit fault. Only the supply from a half-bridge is considered. The current references are chosen so that the second order harmonic of the electromagnetic torque is eliminated. Unlike in the case of open-circuit faults, the injection of third harmonic currents is not contemplated. The healthy phase current phasors are chosen so that the faulted system remains symmetrical and the currents in opposing phases have the same magnitude.

In the following subsections, the method based on minimizing the stator copper losses while maintaining the main harmonic of the air-gap MMF waveform is applied to the post-fault operation of the machine with a terminal short-circuit fault. Two cases are contemplated depending on whether the voltage of the neutral connection can be controlled or not. Additionally, a method similar to the one proposed in [123], based on selecting symmetrical healthy phase current references, is discussed.

Copper loss minimization with partial isolation of the faulted phase

The drive condition where the faulted phase is isolated by shutting down the corresponding inverter leg is considered. In this case, the related branch current, as seen by the inverter, is forced to zero, which causes the zero sequence current to be a function of the short-circuit current in the faulted phase. If the faulted phase is denoted by index $k$:

$$
\begin{align*}
[i_{ph}]_k &= i_k(t) \\
[i_{branch}]_k &= 0
\end{align*}
$$

and, hence:

$$
\sum_{l=1}^{m} [i_{branch}]_l = 0 \Rightarrow \sum_{l=1}^{m} [i_{branch}]_l = \sum_{l\neq k}^{m} [i_{ph}]_l = 0 \Rightarrow \sum_{l=1}^{m} [i_{ph}]_l = i_k(t)
$$

From this last equation and by considering the definition of the zero sequence current given in Appendix B, it turns out that:

$$
i_0 = \frac{\sqrt{2}}{m} \sum_{l=1}^{m} [i_{ph}]_l = \frac{\sqrt{2}}{m} i_k(t)
$$

The current references to be followed, in this case, are the same as those considered in the open-circuit fault of a single phase winding. The only difference between both scenarios is that, in the present case, a compensation term corresponding to the short-circuit current in the faulted phase is added to the third order harmonic and zero sequence current references. In the case phase $A$ in a 5-phase machine is short-circuited ($k = 1$):

$$
\begin{align*}
i_{\alpha_1} &= \chi \cdot i_{\alpha_1,h} \\
i_{\beta_1} &= \chi \cdot i_{\beta_1,h} \\
i_{\alpha_3} &= -\chi \cdot i_{\alpha_1,h} + 4/5 \cdot i_A(t) \\
i_{\beta_3} &= 0 \\
i_0 &= \sqrt{2}/5 \cdot i_A(t)
\end{align*}
$$
CHAPTER 6. FAULT-TOLERANT OPERATION OF A PMSM DRIVE

Assuming the mutual inductances between phases are negligible and denoting the phase self-inductance by \( L_{ph} \), the short-circuit current in the faulted phase can be analytically calculated from the voltage equation:

\[
\begin{align*}
    u_{kN} &= R_{ph}i_k + L_{ph} \frac{di_k}{dt} + e_k(t) = 0 \quad (6.10)
\end{align*}
\]

If only the first harmonic of the PM flux linkage is considered:

\[
\begin{align*}
    e_k(t) &= \omega \lambda_{m1} \cos(\theta_e - (k - 1) \alpha_m) \quad (6.11)
\end{align*}
\]

and:

\[
\begin{align*}
    i_k(t) &= -\frac{\omega \lambda_{m1}}{Z_{sc}} \cos(\theta_e - (k - 1) \alpha_m - \phi_{sc}) \quad (6.12)
\end{align*}
\]

where \( Z_{sc} = \sqrt{R_{ph}^2 + (L_{ph}\omega_e)^2} \) and \( \tan \phi_{sc} = L_{ph}\omega_e/R_{ph} \). The current in the faulted phase is, therefore, solely a function of the speed and the machine parameters. Hence, the current references to be introduced and the derating factors needed to limit the voltage, the branch currents or the machine losses depend all on the operating conditions of the drive. In any case, the linear coefficients that relate the third harmonic and the first harmonic current references are constant and equal those introduced for the open-circuit fault of a single phase.

By considering the previous current references with no torque derating (\( \chi = 1 \)), the operating curves displayed in Figure 6.8 are obtained. The figure shows the variation of the main electric variables at steady state and under rated speed and load conditions, together with the pre-fault state reference values (in dash-dot). A constant speed of 600 rpm is considered in the analytical calculation of the electrical variables.

As in the previously considered cases, the proposed control strategy is successful in maintaining the mean torque capability with a small torque ripple. The preservation of the main MMF harmonic is achieved by increasing the current magnitudes in all the healthy phase, although the resulting phase currents are not symmetrical. The voltage, current and loss overratings needed to deliver the rated torque are moderately low. The pulsating third order current components are significant, but not as much as in the case of an open-circuit fault involving the loss of two adjacent phases with no torque derating. Owing to this, it is expected that the increase in rotor losses and the related PM demagnetization effects are not harmful for the considered prototype machine. These two aspects are further investigated in a latter subsection with the help of a FEA.

As explained in Chapter 4, the fact that the short-circuit current in the faulted phase is a bit higher than the rated current is due to the fact that the prototype machine was designed assuming a magnet temperature of \( T_{PM} = 100 \) °C. Results obtained in latter thermal simulations suggest that this temperature, operating at steady state healthy machine conditions, may be closer to 75 °C; meaning that the magnet remanence and, thus, the PM flux linkage and the short-circuit current magnitude will be higher than initially considered. Anyhow, the percent increase in the current magnitude for the faulted machine is not that high (13.61%), so the prototype design is deemed acceptable; at least for the present fault conditions.

As a general conclusion, it can be stated that, although the fault scenario is worse than the open-circuit of a single phase in terms of increased copper loss, it is not such a dramatic fault as the open-circuit fault of two adjacent phases.

Copper loss minimization without isolation of the faulted phase

If the neutral point of the star connection is allowed to remain connected to the inverter leg corresponding to the faulted phase, a path for the faulted branch current to flow is provided and control over the zero sequence current is gained. This additional degree of control can be used in an attempt to reduce the stator copper losses. In this case, the current references must be computed from equation (F.10). The expressions are the same as
6.2. REMEDIAL STRATEGIES

Figure 6.8: Steady state operating curves for single phase short-circuit fault with partial fault isolation. Modified control strategy aimed at minimizing stator copper losses with no torque derating ($\chi = 1$) for the open-circuit fault of a single phase winding with supply from a full H-bridge inverter, with the exception that, in the present case, an additional term depending on the short-circuit current must be added. For a 5-phase machine with phase $A$ short-circuited:

$$\begin{cases}
i_{\alpha_1} = \chi \cdot i_{\alpha_1,h} \\
i_{\beta_1} = \chi \cdot i_{\beta_1,h} \\
i_{\alpha_3} = -\frac{2}{3} \cdot \chi \cdot i_{\alpha_1,h} + 2/3 \cdot i_A(t) \\
i_{\beta_3} = 0 \\
i_0 = -\sqrt{2}/3 \cdot \chi \cdot i_{\alpha_1,h} + \sqrt{2}/3 \cdot i_A(t)
\end{cases} \quad (6.13)$$

By adopting the previous current references and assuming no torque derating ($\chi = 1$), the operating curves displayed in Figure 6.9 are analytically derived for the considered prototype machine.
CHAPTER 6. FAULT-TOLERANT OPERATION OF A PMSM DRIVE

Figure 6.9: Steady state operating curves for single phase short-circuit fault with no fault isolation. Modified control strategy aimed at minimizing stator copper losses with no torque derating (χ = 1)

Compared to the previous case in which the faulted phase was partially isolated, the present control strategy achieves an overall reduction of the healthy phase current magnitudes by allowing a substantial current to flow through the corresponding inverter leg. The reduction in the current magnitudes is easiest appreciated by comparing the current components referred to the α – β reference frame. Thus, the third order harmonic current components, \( i_d^3 \) and \( i_q^3 \), have a lower amplitude than in the previous case, which translates into a reduction of the torque ripple. On the contrary, in order to preserve the main harmonic of the rotating MMF, more bus voltage is required, meaning that the reduction in the necessary current overrating is achieved at the expense of increasing the voltage overrating. As one may expect, the additional degree of freedom gained from the control of the zero sequence current is useful in reducing the stator copper losses.

A more detailed comparison between the operation modes presented above, aimed at minimizing stator copper losses, with and without isolation of the faulted phase, is presented in a latter point.
Fault-tolerant operation with symmetrical currents A final modified control strategy in which the healthy phase currents remain symmetrical two-by-two after the fault is presented. In this case, it is assumed that the faulted phase is isolated from the DC bus by turning off the corresponding inverter switches. The method, closely similar to that proposed by [123], is based on the imposition that the sum of currents in the healthy phases is zero and that the healthy currents are symmetrical with respect to the faulted phase. Hence, the current magnitudes in opposing phases are chosen to be equal, whereas the currents of adjacent phases are allowed to be different. In case the terminal short-circuit fault of phase A is considered, the post-fault current references to be followed are:

\[
\begin{align*}
    i_A(t) &= -\omega_e \lambda_{m1}/Z_{sc} \cdot \cos(\theta_e - \alpha_m - \phi_c) \\
    i_B(t) &= B \cos(\theta_e + \psi + \phi_b) \\
    i_C(t) &= C \cos(\theta_e + \psi + \phi_c) \\
    i_D(t) &= -B \cos(\theta_e + \psi + \phi_b) \\
    i_E(t) &= -C \cos(\theta_e + \psi + \phi_c)
\end{align*}
\]

where parameters \( B, C, \phi_b \) and \( \phi_c \) are calculated in order to fulfill some additional constraint. Instead of imposing that the resulting second order torque harmonic is zero as in [123], here the preservation of the main air-gap MMF harmonic after the fault is sought. If only the first current harmonic is considered, the pre-fault currents referred to the \( \alpha - \beta \) reference frame are:

\[
\begin{align*}
    i_{\alpha_1}(t) &= \sqrt{2} I_1 \cos(\theta_e + \psi) \\
    i_{\beta_1}(t) &= \sqrt{2} I_1 \sin(\theta_e + \psi) \\
    i_{\alpha_3}(t) &= 0 \\
    i_{\beta_3}(t) &= 0 \\
    i_0(t) &= 0
\end{align*}
\]

Therefore, by referring the current references in equation (6.14) to the \( \alpha - \beta \) reference frame and equating the terms in the \( i_{\alpha_1} \) and \( i_{\beta_1} \) components (plus a derating factor \( \chi \)), the following system is derived for variables \( B, C, \phi_b \) and \( \phi_c \):

\[
\begin{align*}
    B \cos \phi_b &= \sqrt{10}/2 \cdot \chi \cdot I_1 + \sqrt{5}/5 \cdot \omega_e \lambda_{m1}/Z_{sc} \cdot \cos(\phi_{sc} + \psi) \\
    B \sin \phi_b &= -\sqrt{5}/2 \left( \sqrt{5} + \sqrt{5} - \sqrt{5} - \sqrt{5} \right) \cdot \chi \cdot I_1 - \sqrt{5}/5 \cdot \omega_e \lambda_{m1}/Z_{sc} \cdot \sin(\phi_{sc} + \psi) \\
    C \cos \phi_c &= -\sqrt{10}/2 \cdot \chi \cdot I_1 - \sqrt{5}/5 \cdot \omega_e \lambda_{m1}/Z_{sc} \cdot \cos(\phi_{sc} + \psi) \\
    C \sin \phi_c &= -\sqrt{5}/2 \left( \sqrt{5} + \sqrt{5} - \sqrt{5} - \sqrt{5} \right) \cdot \chi \cdot I_1 + \sqrt{5}/5 \cdot \omega_e \lambda_{m1}/Z_{sc} \cdot \sin(\phi_{sc} + \psi)
\end{align*}
\]

Such a system can be readily solved numerically. Once the values for the variables \( B, C, \phi_b \) and \( \phi_c \) are known, the current references in the healthy phases can be computed from equation (6.14). For instance, Table 6.8 collects the values for the previous parameters when the prototype machine operating at 600 rpm is considered. The derating factors have been analytically calculated in each case in order to ensure the corresponding goals (voltage limit, same current, etc.).

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ((\chi = 0.219))</th>
<th>Same current ((\chi = 0.429))</th>
<th>Same copper losses ((\chi = 0.570))</th>
<th>Same torque ((\chi = 1.000))</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current magnitude, ( B ) [A]</td>
<td>4.972</td>
<td>6.703</td>
<td>7.925</td>
<td>11.76</td>
</tr>
<tr>
<td>Current magnitude, ( C ) [A]</td>
<td>2.909</td>
<td>3.625</td>
<td>4.532</td>
<td>8.008</td>
</tr>
<tr>
<td>Current angle, ( \phi_b ) [rad]</td>
<td>-1.180</td>
<td>-1.028</td>
<td>-0.964</td>
<td>-0.852</td>
</tr>
<tr>
<td>Current angle, ( \phi_c ) [rad]</td>
<td>2.279</td>
<td>2.843</td>
<td>3.076</td>
<td>-2.884</td>
</tr>
</tbody>
</table>

Table 6.8: Healthy phase current parameters for control strategy with symmetrical currents

The resulting steady state operating curves obtained with the present modified control strategy are very similar to those obtained with the strategy aimed at minimizing the stator copper losses when the faulted
phase is isolated from the DC bus and no derating is considered. Therefore, the curves for the different electrical variables are not shown here. The fundamental difference between both operation modes is that, with the former method, the \( i_{d_3} \) current component is zero all over the period, whereas in the present case it varies sinusoidally with a small amplitude. The rest of current components referred to the \( \alpha - \beta \) reference frame are exactly the same. As a consequence, stator copper losses are slightly higher when the healthy phase currents are chosen to be symmetrical with respect to the faulted phase. There is a reduction in the peak current value, but the voltage required to maintain the main MMF harmonic is a bit higher. The waveforms for the third order harmonic currents, \( i_{d_3} \) and \( i_{q_3} \), are very similar in both cases. A more detailed comparison between both control strategies is given in the next subsection by discussing the results obtained from conducting FE simulations on the designed prototype machine.

**FE simulation results** In the following section, a comparison among the 3 methods discussed above is presented based on the results obtained by a number of FE simulations. As with other fault scenarios, for each modified control strategy, different torque derating factors (\( \chi \)) have been introduced in order to preserve one of the following variables: the voltage limit imposed by the DC bus (\( U_{DC} = 300 \text{ V} \)), the peak current in the inverter legs, the stator copper losses or the mean electromagnetic torque. The derating factors and the current references have been analytically calculated for the linear machine parameters stated in Chapter 5. It must be noted that, since the short-circuit current in the faulted phase depends on the fundamental frequency, the derating factors and the current references to be introduced, are a function of the speed of the machine.

The FE simulation results for the control strategy aimed at minimizing the stator copper losses in the case the inverter leg corresponding to the faulted phase is shut down are collected in Table 6.1; whereas the results for that same strategy when the neutral point is allowed to remain connected to DC bus are shown in Table 6.2. Finally, the results for the fault-tolerant operation with symmetrical currents are displayed in Table 6.3.

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ((\chi = 0.339))</th>
<th>Same current ((\chi = 0.392))</th>
<th>Same copper losses ((\chi = 0.575))</th>
<th>Same torque ((\chi = 1.000))</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak branch current ([\text{A}])</td>
<td>6.227</td>
<td>6.704</td>
<td>8.412</td>
<td>12.47</td>
<td>6.703</td>
</tr>
<tr>
<td>Peak voltage, (u_{ph,N,max} [\text{V}])</td>
<td>149.6</td>
<td>152.1</td>
<td>162.1</td>
<td>201.4</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, (T_{mean} [\text{N-m}])</td>
<td>9.007</td>
<td>10.38</td>
<td>15.15</td>
<td>25.21</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, (\Delta T [\text{N-m}])</td>
<td>0.729</td>
<td>0.736</td>
<td>0.792</td>
<td>2.786</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, (P_{c,1} [\text{W}])</td>
<td>71.76</td>
<td>78.38</td>
<td>108.8</td>
<td>221.5</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, (P_{c,t,1} [\text{W}])</td>
<td>26.64</td>
<td>26.99</td>
<td>28.45</td>
<td>32.31</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, (P_{c,y,1} [\text{W}])</td>
<td>7.401</td>
<td>7.630</td>
<td>8.616</td>
<td>11.62</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, (P_{r,1} [\text{W}])</td>
<td>0.217</td>
<td>0.239</td>
<td>0.332</td>
<td>0.592</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, (P_{r,m} [\text{W}])</td>
<td>0.115</td>
<td>0.137</td>
<td>0.237</td>
<td>0.576</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, (P_{loss} [\text{W}])</td>
<td>106.1</td>
<td>113.4</td>
<td>146.4</td>
<td>266.6</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density ([\text{T}])</td>
<td>0.910</td>
<td>0.911</td>
<td>0.915</td>
<td>0.867</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Table 6.9: FE simulation results for single phase short-circuit fault with modified control and different derating factors. Control aimed at minimizing stator copper losses with partial isolation of the faulted phase.

As it can be appreciated from the previous tables, when no torque derating is considered (\( \chi = 1.000 \)), all three methods are capable of delivering the rated torque with quite a low ripple. The slight decrease in the mean torque value with respect to the pre-fault state is due to saturation of the iron lamination material. Although the reduction in mean torque is similar in the three cases, the increase in torque ripple amplitude due to local iron saturation is highest when the strategy with symmetrical currents is chosen. The current and voltage overratings required to deliver the rated torque are similar in all the cases. The peak current flowing through the power semiconductor switches is lowest with the symmetrical currents method, but the required voltage is highest. The methods that contemplate the partial isolation of the faulted phase by turning off the corresponding inverter leg perform similarly in terms of increased machine losses. If the faulted phase is not isolated and the zero sequence current can be controlled, a modest reduction in machine losses can be achieved (9.34%). As in the fault scenarios involving the open-circuit of one or more phases, the main increase in machine losses at fault conditions comes from a substantial
6.2. REMEDIAL STRATEGIES

Table 6.10: FE simulation results for single phase short-circuit fault with modified control and different derating factors. Control aimed at minimizing stator copper losses with no fault isolation.

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ((x = 0.219))</th>
<th>Same current ((x = 0.429))</th>
<th>Same copper losses ((x = 0.570))</th>
<th>Same torque ((x = 1.000))</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak voltage, (u_{ph,N_{max}}) [V]</td>
<td>147.6</td>
<td>159.4</td>
<td>176.8</td>
<td>207.0</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, (T_{mean}) [N-m]</td>
<td>1.157</td>
<td>9.463</td>
<td>16.46</td>
<td>25.13</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, (\Delta T) [N-m]</td>
<td>1.073</td>
<td>1.016</td>
<td>1.094</td>
<td>2.864</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, (P_{Cu}) [W]</td>
<td>47.23</td>
<td>67.62</td>
<td>108.8</td>
<td>198.3</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, (P_{Fe,t}) [W]</td>
<td>24.86</td>
<td>25.95</td>
<td>27.93</td>
<td>31.34</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, (P_{Fe,y,s}) [W]</td>
<td>6.407</td>
<td>7.109</td>
<td>8.466</td>
<td>11.03</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, (P_{Fe,y,r}) [W]</td>
<td>0.125</td>
<td>0.209</td>
<td>0.342</td>
<td>0.563</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, (P_{PM}) [W]</td>
<td>0.040</td>
<td>0.111</td>
<td>0.254</td>
<td>0.541</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, (P_{loss}) [W]</td>
<td>78.67</td>
<td>101.0</td>
<td>145.8</td>
<td>241.7</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density [T]</td>
<td>0.905</td>
<td>0.911</td>
<td>0.916</td>
<td>0.896</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Table 6.11: FE simulation results for single phase short-circuit fault with modified control and different derating factors. Control with symmetrical currents and partial isolation of the faulted phase.

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Voltage limit ((x = 0.219))</th>
<th>Same current ((x = 0.429))</th>
<th>Same copper losses ((x = 0.570))</th>
<th>Same torque ((x = 1.000))</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak voltage, (u_{ph,N_{max}}) [V]</td>
<td>147.6</td>
<td>159.4</td>
<td>176.8</td>
<td>207.0</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, (T_{mean}) [N-m]</td>
<td>5.814</td>
<td>11.32</td>
<td>14.94</td>
<td>25.09</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, (\Delta T) [N-m]</td>
<td>0.736</td>
<td>0.793</td>
<td>1.104</td>
<td>3.494</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, (P_{Cu}) [W]</td>
<td>60.04</td>
<td>84.23</td>
<td>108.8</td>
<td>224.5</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, (P_{Fe,t}) [W]</td>
<td>25.94</td>
<td>27.16</td>
<td>28.24</td>
<td>32.08</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, (P_{Fe,y,s}) [W]</td>
<td>6.909</td>
<td>7.769</td>
<td>8.520</td>
<td>11.54</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, (P_{Fe,y,r}) [W]</td>
<td>0.175</td>
<td>0.257</td>
<td>0.329</td>
<td>0.594</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, (P_{PM}) [W]</td>
<td>0.076</td>
<td>0.155</td>
<td>0.235</td>
<td>0.581</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, (P_{loss}) [W]</td>
<td>95.20</td>
<td>119.6</td>
<td>146.1</td>
<td>269.3</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density [T]</td>
<td>0.908</td>
<td>0.913</td>
<td>0.916</td>
<td>0.874</td>
<td>0.959</td>
</tr>
</tbody>
</table>

increase in the stator copper losses. With the first method, the overall stator losses increase by a 76.52% compared to the pre-fault state, whereas the rotor losses increase by a 12.31%. It must be noted that the temperature increase in each phase will be different with any of the proposed control strategies, as no method results in equal magnitude currents. In order to accurately predict the post-fault temperature distribution in the machine, a complete thermal model that accounted for each phase separately should be employed.

As a general conclusion for the reviewed methods, the three control strategies described above fulfill the requirement of producing the desirable electromagnetic torque with little degradation on the machine operating figures. Since the demagnetizing effects of the third harmonic currents over the PMs are low in all three cases, it can be stated that, as long as the increased losses can be effectively evacuated, any of the proposed operation modes can be safely applied in the fault-tolerant operation of the considered PMSM drive at rated speed conditions. In terms of machine loss increase, the method aimed at minimizing stator copper losses with control over the zero sequence current performs best. However, if strict constraints on the available DC bus voltage and the peak branch currents are placed, it leads to the largest torque derating among the three methods.

When it comes to the ease of implementation, the methods aimed at minimizing the stator copper losses may be advantageous with respect to the symmetrical currents method in terms of a reduced number of computations. The linear coefficients that relate the \( (i_{a3}, i_{b3}) \) and \( (i_{a3}, i_{b3}) \) current components can be easily stored in look-up tables depending on the faulted phase index, with the advantage that they do not depend on the drive's speed. Furthermore, since these coefficients are the same as the ones to be used in case a single phase open-circuit is considered, the data storage requirements for the PMSM drive to be tolerant to multiple faults are lower. On the contrary, the method proposed by [123], based
on choosing symmetrical current references, requires that the amplitude and angle of the currents in the healthy phases are calculated for each operating speed, making the amount of data to be pre-calculated and stored much greater. Furthermore, based on the derived current reference equations, the symmetrical currents method is only readily applicable to 5-phase machines, whereas the method aimed at minimizing the copper losses can be easily adapted to other phase number drives just by considering the corresponding transformation matrix [T] in equations (F.10) and (F.12); (see Appendices B and F).

Finally, in order to demonstrate the adequacy of the reviewed control strategies, the method aimed at minimizing the stator copper losses with partial fault isolation is compared to the steady state operation of the PMSM drive in case no remedial action after the short-circuit fault is taken. In the latter case, the phase current waveforms are considered to be equal to those obtained in Chapter 5 when $d - q$ current vector control is applied (see Figure 5.16a). The results comparing both post-fault scenarios, together with the data corresponding to the pre-fault state, are collected in Table 6.12. Since in the case when no remedial action is taken machine losses remain fairly constant with respect to the pre-fault state, in order to have a more sound comparison between both operation modes, a torque derating factor is considered for the modified control strategy to maintain the stator copper losses ($\chi = 0.816$).

<table>
<thead>
<tr>
<th></th>
<th>No remedial action</th>
<th>Copper loss minimization</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak voltage, $u_{ph,N,max}$ [V]</td>
<td>287.1</td>
<td>162.1</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, $T_{mean}$ [N-m]</td>
<td>19.29</td>
<td>15.15</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, $\Delta T$ [N-m]</td>
<td>19.44</td>
<td>0.792</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, $P_{C_u}$ [W]</td>
<td>105.4</td>
<td>108.8</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, $P_{Fe,t}$ [W]</td>
<td>26.81</td>
<td>28.45</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, $P_{Fe,y,rs}$ [W]</td>
<td>8.962</td>
<td>8.616</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, $P_{Fe,r}$ [W]</td>
<td>0.454</td>
<td>0.352</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, $P_{Mag}$ [W]</td>
<td>0.429</td>
<td>0.237</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, $P_{loss}$ [W]</td>
<td>142.1</td>
<td>146.4</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density [T]</td>
<td>0.920</td>
<td>0.915</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Table 6.12: FE simulation results for single phase terminal short-circuit fault when no remedial action is taken. Supply from a half-bridge inverter

The results indicate that, although both operation modes perform fairly similarly with regards to machine losses, the adoption of the proposed modified control strategy can remarkably improve the quality of the electromagnetic torque waveform. Although the torque capability is lessened by a 21.46%, the high torque pulsations almost disappear. Moreover, the peak voltages and currents are lower when the control strategy is varied. It must be noted that the currents when no remedial actions after the fault are taken have been computed and latter introduced in the FE software by assuming an infinite iron permeability and a DC bus voltage of $U_{DC} = 300$ V. By examining the resulting peak phase to neutral voltages, it can be concluded that considering the real magnetization curve of the iron material can lead to substantial variations in the operating curves at fault conditions. It is therefore expected that the real PMSM prototype would behave differently as considered when no remedial action after the fault was taken. In any case, the obtained results indicate that a substantial improvement can be achieved when the proposed post-fault control strategy is adopted.

**Dynamic response of the PMSM drive** As in previous cases, the adequacy of implementing the method aimed at minimizing the stator copper losses while partially isolating the faulted phase has been evaluated with the aid of the developed analytical model. The drive’s dynamic response is displayed in Figure 6.10. In the simulations, a short-circuit fault is introduced in phase $A$ at the fifth electrical period, whereas the remedial actions take place at a time corresponding to the 15$^{th}$ period. The dynamic simulations have been conducted for a DC bus voltage of 300 V and 500 V. The DC bus voltage has been increased over the theoretical value calculated analytically (see Figure 6.8) in order to improve the dynamic response of the system after the change in the control strategy.

The simulation results indicate that modifying the control strategy may be detrimental in terms of the deliverable mean torque if not enough voltage for the post-fault operation of the drive is provided.
6.2. REMEDIAL STRATEGIES

Figure 6.10: Drive’s transient response to the terminal short-circuit of a single phase with modified control strategy aimed at minimizing the stator copper losses (constant speed of 600 rpm). Supply from a half-bridge inverter with partial fault isolation and no torque derating ($\chi = 1$).

Furthermore, the considered PMSM drive is not able to deliver the rated torque after the fault, even if the voltage is substantially increased. The torque ripple resulting from the drive operating with the proposed modified control strategy is also higher than for the previous fault scenarios. The reason for these phenomena is that the current references $i_{\alpha_1}^*$ and $i_{\alpha_3}^*$ are not adequately followed after the change in the control law; most probably due to saturation or an insufficient bandwidth of the current controllers. Anyhow, the torque ripple is clearly diminished when the proposed modified control strategy is adopted and, as long as enough voltage is provided, the mean torque capability of the drive is also increased with respect to the post-fault scenario when no remedial action is taken.

The dynamic response of the PMSM drive when the same method with no fault isolation is used is shown in Figure 6.11. Although the present strategy leads to lower copper losses, the drive’s response is worse in terms of the resulting electromagnetic torque and, thus, it is discarded in favor of the previous method.

Figure 6.11: Drive’s transient response to the terminal short-circuit of a single phase with modified control strategy aimed at minimizing the stator copper losses (constant speed of 600 rpm). Supply from a half-bridge inverter with no fault isolation and no torque derating ($\chi = 1$).
Short-circuit of two adjacent phases

The fault-tolerant operation of the PMSM drive subject to a terminal short-circuit between two adjacent phases is investigated next. Since a sudden connection between the terminals of two phases would result in a DC bus shoot-through, it is assumed that the fault is immediately detected by the inverter’s protection hardware, leading to the automatic opening of the corresponding inverter legs. Therefore, the fault is automatically isolated from the DC bus by the protection equipment of the drive system and no further action has to be considered to isolate the fault.

Like in the case of terminal short-circuit faults involving the loss of a single phase, a number of different control strategy approaches can be used in the present fault scenario. However, the terminal short-circuit between two adjacent phases has received little attention in the related literature. With minimum effort, any of the previously reviewed methods could be easily applied in the post-fault operation of a PMSM drive subject to this kind of fault. In the following section, the control strategy derived in Appendix F aimed at preserving the main harmonic of the air-gap MMF is considered.

Fault-tolerant operation with preservation of the main air-gap MMF harmonic A short-circuit fault between two machine terminals involves the loss of control over two phase currents. If the faulted phases are denoted by indexes \(k\) and \(j\), the short-circuit currents can be analytically calculated by equating the phase to neutral voltages and realizing that \(i_k(t) = -i_j(t)\). If a negligible mutual inductance between phases is assumed:

\[
\begin{align*}
    u_{kN} &= R_{ph}i_k + L_{ph}\frac{di_k}{dt} + e_k(t) = R_{ph}i_j + L_{ph}\frac{di_j}{dt} + e_j(t) = u_{jN} \\
    R_{ph}i_k + L_{ph}\frac{di_k}{dt} + \frac{1}{2}(e_k(t) - e_j(t)) &= 0
\end{align*}
\]  

and, if only the first harmonic of the PM flux linkage is considered:

\[
\begin{align*}
    e_k(t) &= \omega_e \lambda_{m1} \cos(\theta_e - (k-1)\alpha_m) \\
    e_j(t) &= \omega_e \lambda_{m1} \cos(\theta_e - (j-1)\alpha_m)
\end{align*}
\]

Thus:

\[
\begin{align*}
    i_k(t) &= -\frac{\omega_e \lambda_{m1}}{Z_{sc}} \sin(\phi_k) \sin(\theta_e - (k-1)\alpha_m + \phi_k - \phi_{sc}) \\
    i_j(t) &= \frac{\omega_e \lambda_{m1}}{Z_{sc}} \sin(\phi_k) \sin(\theta_e - (k-1)\alpha_m + \phi_k - \phi_{sc})
\end{align*}
\]

where \(\phi_k = (k-j)\alpha_m/2\). For the considered fault scenario, no path for the zero sequence current to flow is provided \(i_0 = 0\) and, hence, the current references to be introduced in order to preserve the main MMF harmonic are computed by solving the linear system given by equations (F.19), (F.24) and (F.25). In case a 5-phase machine with a terminal short-circuit between phases \(C\) and \(D\) \((k = 3, j = 4)\) is considered:

\[
\begin{align*}
    i_{\alpha1} &= \chi \cdot i_{\alpha1,h} \\
    i_{\beta1} &= \chi \cdot i_{\beta1,h} \\
    i_{\alpha2} &= 2.6180 \cdot \chi \cdot i_{\alpha1,h} \\
    i_{\beta2} &= -0.6180 \cdot \chi \cdot i_{\beta1,h} + 1.0515 \cdot i_C(t)
\end{align*}
\]

By employing the current references computed from equations (6.20) and (6.21) and assuming no torque derating \((\chi = 1)\), the operating curves displayed in Figure 6.12 can be analytically calculated for the prototype machine. The results obtained closely resemble those in the case the open-circuit of two adjacent phases is considered. In order to maintain the drive’s torque capability, large currents flow in
the healthy phases and a large overrating is required in terms of phase to neutral voltages and stator copper losses. The resulting third harmonic current components have even higher magnitudes than the first harmonic current references, which leads to a noticeable increase in the torque ripple magnitude.

\[ \theta_e \text{ [rad]} \]

\[ i_{ph} \text{ [A]} \]

\[ 0 \quad \frac{\pi}{2} \quad \pi \quad \frac{3\pi}{2} \quad 2\pi \]

\[ -30 \quad -20 \quad -10 \quad 0 \quad 10 \quad 20 \quad 30 \]

\( i_{A0}, i_{B0}, i_{C0}, i_{D0}, i_{E0} \)

\( \theta_e \) 

\( \text{Phase currents (solid), healthy state peak value (dash-dot)} \)

\( \theta_e \) 

\( \text{\( \alpha - \beta \) currents (solid), healthy state peak value (dash-dot)} \)

\( \theta_e \) 

\( i_{dq} \text{ [A]} \)

\[ 0 \quad \frac{\pi}{2} \quad \pi \quad \frac{3\pi}{2} \quad 2\pi \]

\[ -30 \quad -20 \quad -10 \quad 0 \quad 10 \quad 20 \quad 30 \]

\( i_{d1}, i_{q1}, i_{d3}, i_{q3} \)

\( \theta_e \) 

\( \text{d - q currents} \)

\( \theta_e \) 

\( u_{phN} \text{ [V]} \)

\[ 0 \quad \frac{\pi}{2} \quad \pi \quad \frac{3\pi}{2} \quad 2\pi \]

\[ 200 \quad 400 \quad 600 \quad 800 \quad 1000 \]

\( u_{AN}, u_{BN}, u_{CN}, u_{DN}, u_{EN} \)

\( \theta_e \) 

\( \text{Phase to neutral voltages (solid), } u_{1,max} = \frac{1}{2}U_{DC} \) 

\( \theta_e \) 

\( T \text{ [N.m]} \)

\[ 0 \quad \frac{\pi}{2} \quad \pi \quad \frac{3\pi}{2} \quad 2\pi \]

\[ 25 \quad 250 \quad 500 \quad 750 \quad 1000 \]

\( \theta_e \) 

\( P_{Cu} \text{ [W]} \)

\[ 0 \quad \frac{\pi}{2} \quad \pi \quad \frac{3\pi}{2} \quad 2\pi \]

\[ 0 \quad 250 \quad 500 \quad 750 \quad 1000 \]

\( \theta_e \) 

\( \text{Electromagnetic torque (solid), healthy state mean value (dash-dot)} \)

\( \theta_e \) 

\( \text{Stator copper losses (solid), healthy state mean value (dash-dot)} \)

Figure 6.12: Steady state operating curves for terminal short-circuit between two adjacent phases. Modified control strategy aimed at maintaining the main air-gap MMF harmonic with no torque derating (\( \chi = 1 \))

**FE simulation results** In the same way as for previous fault scenarios, the performance of the designed PMSM drive under the considered post-fault operation mode has been evaluated via a number of FE simulations. The simulation results for different derating factors are collected in Table 6.13.

The simulation results demonstrate that the proposed control strategy is successful in preserving the torque capability of the drive. Interestingly enough, non-linearities in the magnetization curve of the iron material cause the main torque to be enhanced when no derating factor is considered (\( \chi = 1.000 \)). However, a significant torque ripple results from operating at rated torque and the current and voltage...
Electromagnetic torque

\[
T = N \cdot m
\]

\[\text{Phase currents (} U_{DC} = 700 \text{ V)}
\]

Figure 6.13: Drive’s transient response to the terminal short-circuit between two adjacent phases with modified control strategy aimed at maintaining the main air-gap MMF harmonic (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating (\(\chi = 1\)).

Table 6.13: FE simulation results for two adjacent phases short-circuit fault with modified control and different derating factors. Control aimed at minimizing stator copper losses with supply from a half-bridge inverter

Dynamic response of the PMSM drive  

The dynamic response of the considered PMSM drive after a terminal short-circuit fault between two adjacent phases and the modification of the control strategy is shown in Figure 6.13. In the simulations, a terminal short-circuit fault between phases \(C\) and \(D\) has been contemplated with a DC bus voltage supply of 700 V.

Compared to the case where the current control was implemented in the \(d - q\) reference frame (see Figure 5.17), the present fault condition is worse in terms of mean torque capability. Until the proposed remedial actions are taken, the current control in the \(a - \beta\) reference frame delivers a lower torque that also becomes negative during parts of the electrical cycle. As in previous fault scenarios, the simulation results demonstrate that vast improvements in the torque waveform can be obtained by modifying the current references after the fault; specially if sufficient voltage is provided for the drive to operate after the fault. Like in the case of a single phase terminal short-circuit fault, the dynamic results indicate that more stringent requirements are placed on the current controllers when trying to follow the proposed
current references than in the case of open-circuit faults. Anyhow, as long as enough voltage for the post-fault operation is provided, a high mean torque with quite a small ripple can be achieved.

**Short-circuit of two non-adjacent phases**

The last short-circuit fault that is considered is the terminal short-circuit between two non-adjacent phases. The same structure as with the previous fault scenarios is considered. Like in the case of two adjacent phases short-circuit, it is assumed that that the fault is immediately detected by the inverter’s protection hardware, leading to the automatic opening of the corresponding inverter legs. In the following section, the use of the control strategy aimed at preserving the main air-gap MMF harmonic is proposed.

**Fault-tolerant operation with preservation of the main air-gap MMF harmonic**

In this case, it is assumed that the short-circuit fault involves phases $B$ and $E$ ($k = 2$, $j = 5$). Therefore, the current references, computed by the formulas in Appendix F for a 5-phase machine, are:

\[
\begin{align*}
    i_{\alpha_1} &= \chi \cdot i_{\alpha_1,h} \\
    i_{\beta_1} &= \chi \cdot i_{\beta_1,h} \\
    i_{\alpha_3} &= 0 \\
    i_{\beta_3} &= 1.6180 \cdot \chi \cdot i_{\beta_1,h} - 1.7013 \cdot i_B(t)
\end{align*}
\]  

The steady state operating curves obtained by applying the previous current references with no torque derating ($\chi = 1$) are shown in Figure 6.14. The analytically calculated results are quite similar to those obtained when the short-circuited phases are adjacent. The magnitude of the short-circuit currents is higher, but the increase in healthy phase currents is somewhat lower than in the previous case. As a result, the increase in copper losses is a bit lower than in the adjacent phases case. Both operation modes lead to similar voltage requirements and third order harmonic current magnitudes.

**FE simulation results**

In the same way as for other fault scenarios, the performance of the designed PMSM drive with the proposed control strategy has been evaluated via a FEA. However, unlike in previous cases, the analytical calculations suggest that is not possible to comply with the voltage limit imposed by the DC bus (while having purely sinusoidal phase currents) or maintain the peak current magnitudes or the stator copper losses with respect to their pre-fault values by introducing suitable torque derating factors. Therefore, only the case for which $\chi = 1$ has been simulated. The simulation results for such conditions are collected in Table 6.14.

<table>
<thead>
<tr>
<th>Derating factor</th>
<th>Same torque ($\chi = 1.000$)</th>
<th>Pre-fault</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak branch current [A]</td>
<td>9.264</td>
<td>6.703</td>
</tr>
<tr>
<td>Peak voltage, $u_{ph,N,max}$ [V]</td>
<td>319.2</td>
<td>142.5</td>
</tr>
<tr>
<td>Mean torque, $T_{mean}$ [N-m]</td>
<td>21.27</td>
<td>26.69</td>
</tr>
<tr>
<td>Torque ripple, $\Delta T$ [N-m]</td>
<td>7.346</td>
<td>0.126</td>
</tr>
<tr>
<td>Stator copper loss, $P_{Cu}$ [W]</td>
<td>468.0</td>
<td>108.9</td>
</tr>
<tr>
<td>Stator teeth loss, $P_{Fe,t}$ [W]</td>
<td>30.97</td>
<td>31.15</td>
</tr>
<tr>
<td>Stator yoke loss, $P_{Fe,y,s}$ [W]</td>
<td>11.76</td>
<td>10.32</td>
</tr>
<tr>
<td>Rotor iron loss, $P_{Fe,y,r}$ [W]</td>
<td>0.644</td>
<td>0.512</td>
</tr>
<tr>
<td>Magnet loss, $P_{PM}$ [W]</td>
<td>0.594</td>
<td>0.528</td>
</tr>
<tr>
<td>Total electrom. loss, $P_{loss}$ [W]</td>
<td>512.0</td>
<td>151.4</td>
</tr>
<tr>
<td>Minimum PM flux density [T]</td>
<td>0.788</td>
<td>0.959</td>
</tr>
</tbody>
</table>

Table 6.14: FE simulation results for two non-adjacent phases short-circuit fault with modified control and different derating factors. Control aimed at minimizing stator copper losses with supply from a half-bridge inverter.

Compared to the case in which the short-circuited phases are adjacent, the present post-fault operation mode leads to quite similar results. The required current and voltage overratings are comparable, although
CHAPTER 6. FAULT-TOLERANT OPERATION OF A PMSM DRIVE

(a) Phase currents (solid), healthy state peak value (dash-dot)

(b) $\alpha - \beta$ currents (solid), healthy state peak value (dash-dot)

(c) $d - q$ currents

(d) Phase to neutral voltages (solid), $u_{\text{1, max}} = \frac{1}{2}U_{\text{DC}}$ (dash-dot)

(e) Electromagnetic torque (solid), healthy state mean value (dash-dot)

(f) Stator copper losses (solid), healthy state mean value (dash-dot)

Figure 6.14: Steady state operating curves for terminal short-circuit between two non-adjacent phases. Modified control strategy aimed at maintaining the main air-gap MMF harmonic with no torque derating ($\chi = 1$)

the increase in machine losses is lower; specially in the stator windings and in the rotor. Stator losses increase by a 239.6% compared to the pre-fault state, whereas rotor losses increase by a factor of 1.190 for the particular machine considered in the present thesis. The saturation of the iron material causes a noticeable reduction of the mean torque and a significant torque ripple. In terms of PM demagnetization, the present post-fault operation mode leads to the highest demagnetization fields among the short-circuit fault scenarios studied up to this point. However, the flux density levels in the PMs are well beyond the demagnetization limit for the magnets used in the fabrication of the prototype (limit of 0.6 T at 150 °C [273]).

Dynamic response of the PMSM drive  The dynamic response of the PMSM drive after the proposed control strategy is applied is displayed in Figure 6.15. In this case, a DC bus voltage of 650 V has
been considered for a terminal short-circuit between phases B and E. Like in the previous fault scenario, the PMSM drive performs poorly after the fault until the control strategy is modified. In this case, a significant breaking torque is produced during part of each electrical period. If the DC bus voltage is not sufficiently high, this breaking torque is not eliminated after the change in the control law. The best performance is achieved when the control law is changed and enough supply voltage is available. In this case, the rated mean torque can be delivered with quite a small ripple.

Figure 6.15: Drive’s transient response to the terminal short-circuit between two non-adjacent phases with modified control strategy aimed at maintaining the main air-gap MMF harmonic (constant speed of 600 rpm). Supply from a half-bridge inverter with no torque derating ($\chi = 1$)

Conclusions for terminal short-circuit faults

In the present section, the capability of the considered PMSM drive to operate under different terminal short-circuit faults has been demonstrated. In addition to conducting a small literature review on previously proposed remedial actions, the application of a general modified control strategy aimed at minimizing the stator copper losses has been discussed. Generally speaking, in order to maintain the mean torque with respect to the pre-fault state, large provisions must be made in terms of available DC bus voltage, power semiconductor switches current withstand capability and machine loss increase. Compared to winding open-circuit faults, terminal short-circuit faults lead to higher machine losses. However, if a derating of the mean torque is allowed and enough voltage and current are provided, the drive can be uninterruptedly operated by delivering a significant amount of the rated torque with a small ripple. An exception to this is the case where the short-circuit fault is produced between two non-adjacent phases. In such a scenario, it is not possible to preserve the shape of the air-gap MMF waveform without incurring extra losses.

Additionally, the results from a number of dynamic simulations conducted with the help of the analytical PMSM drive model developed in the present thesis indicate that the current references cannot be as adequately followed in the case of terminal short-circuit faults as in the case of winding open-circuits. This is thought to be due to a higher voltage and current controller’s bandwidth requirement in the case of short-circuit faults; although this aspect should be further investigated in the future.

6.2.3 Inverter faults

Inverter related faults can be broadly classified among open-circuit faults, short-circuit faults and intermittent gate-misfiring faults [42]. In the present subsection, only faults involving the permanent opening or the permanent closing of an inverter switch are addressed. Since nowadays inverters are usually equipped with a number of protection functions [18], it is assumed that the inverter protection system is able to detect the different faults and trigger the necessary remedial actions. Therefore, fault scenarios involving the loss of more than switch have been considered rare and are not studied in the following subsections.
CHAPTER 6. FAULT-TOLERANT OPERATION OF A PMSM DRIVE

Opening of a single transistor

The most straightforward way of dealing with a transistor open-circuit fault is by opening the corresponding switch in the same leg in order to isolate the faulted phase from the DC bus. In this case, the drive can be operated in the same fashion as with a single phase winding open-circuit fault and any of the previously discussed methods to compute the suitable current references may be used. Likewise, the same procedure can be applied in the case of an open-circuit fault involving more than one transistor. For a fault-tolerant \( m \)-phase drive supplied by a half-bridge inverter, the drive is able to accommodate the permanent loss of \( m - 3 \) switches belonging to different phases without too harsh penalties if the proper torque derating factor is introduced.

A second way of dealing with a single transistor open-circuit fault is by letting the electrical current to flow through the faulted phase during half of the electrical cycle. This method has been proposed for instance in [169] for the post-fault operation of BLDC machines operating in the 180° conducting mode. As they compute the optimal healthy phase currents for every rotor position, the optimization algorithms proposed in [23, 171, 172, 185, 425] could be easily adapted to the present fault scenario, although this operation mode has not been addressed in the previous references. The methods aimed at preserving the main harmonic of the air-gap MMF [165,424] (including the one discussed in Appendix F) or the methods aimed at eliminating specific torque harmonics [114, 123, 174] may face difficulties when computing the suitable current references since the current in the faulted phase is not composed just of a single harmonic.

In the present thesis, the first strategy of dealing with transistor open-circuit faults is considering. By opening the corresponding switch in the same leg, the machine is rendered to a state similar to that of a winding open-circuit and the control strategy aimed at minimizing the stator copper losses while preserving the main harmonic of the air-gap MMF can be used.

Dynamic response of the PMSM drive  As with previous fault scenarios, the transient response of PMSM drive for the considered remedial action is evaluated with the aid of the developed drive model. The simulation results for the drive’s dynamic response when a permanent opening of a single transistor occurs are collected in Figure 6.16. In the simulations, the high side switch of phase \( A \) is open-circuited from the fifth period onwards, whereas the remedial actions are assumed to take place at a time corresponding to the 10\(^{th}\) electrical period.

![Phase currents (\( U_{DC} = 420 \) V)](image)

![Electromagnetic torque](image)

Figure 6.16: Drive’s transient response to the opening of a single switch with modified control strategy aimed at minimizing stator copper losses (constant speed of 600 rpm). Supply from a half-bridge inverter

The simulations prove that, by switching off the opposing switch in the faulted leg and by applying the modified control strategy proposed for winding open-circuit faults, the pulsations in the electromagnetic torque waveform can be eliminated. In the case of \( U_{DC} = 300 \) V, the mean torque delivered by the drive is reduced in a 18.97\%, but the torque oscillations are significantly lower. It can be appreciated, how, if enough DC bus voltage is provided, the drive can deliver the rated torque with a negligible ripple by adopting the proposed modified control strategy.
6.2. REMEDIAL STRATEGIES

Closing of a single transistor

As demonstrated in Chapter 5, the short-circuit of a power semiconductor switch is a dangerous fault scenario that requires immediate action to be taken in order to protect the drive, the machine and the associated mechanical load (shaft, coupling, load, etc.). In this kind of faults, very high currents may flow through both the machine windings and the inverter legs, leading to a fast heating of the aforementioned elements, intense PM demagnetization fields and a high pulsation of the electromagnetic torque. This is a catastrophic drive fault that calls for immediate remedial action to be taken in order to drive the system to a safe state.

Some strategies have been proposed in the related literature to allow AC drives to operate under the occurrence of switch short-circuit faults, e.g. [19,97,98,100,430]. However, the previous methods require that the inverter topology is modified in order to allow the drive hardware to be reconfigured after the fault.

In the following, two possible remedial actions against the short-circuit fault of a power semiconductor switch in a half-bridge inverter are considered. The first one consists on preventing machine operation by turning off all the healthy transistors in the drive [25]. The second one, less intuitive, consist on causing a symmetrical short-circuit fault at the machine terminals by closing the transistors in the same side as the faulted one [62]. In order to avoid a DC bus shoot-through, the remaining power switches must be turned off. In this case, the machine is rendered to a symmetrical state, in which the phase currents are limited by the high phase self-inductance.

Dynamic response of the PMSM drive  The aforementioned remedial strategies are compared by checking the dynamic response of the PMSM drive under both operation modes. In the first case, all the healthy transistors are switched off after the detection of the fault, which is assumed to happen within the first electrical period after the fault. In the second case, the high side switches are closed and the low side switches are opened in order render the system to a symmetrical state. The simulation results for both scenarios are shown in Figures 6.17 and 6.18, respectively. In both cases, a DC bus voltage of 300 V is considered.

![Phase currents](image1.png)

(a) Phase currents

![Electromagnetic torque](image2.png)

(b) Electromagnetic torque

Figure 6.17: Drive’s transient response to the closing of a single switch when the inverter is switched off (constant speed of 600 rpm). Supply from a half-bridge inverter

In the case the inverter is switched off, the free-wheeling diodes associated to each transistor provide a path for the currents in the healthy phases to flow during part of the electrical cycle. As a result, the current in the faulted phase retains a high DC component that is mainly limited by the phase resistance. The high short-circuit current is likely to produce damages to the machine windings and/or the inverter equipment if the drive is not quickly stopped. The high pulsating torque that is obtained may also jeopardize the integrity of the mechanical load and the elements in between (shaft, bearings, coupling, etc.). In this case, it might be necessary to incorporate TRIACs (TRIode for Alternating Current) or any other safety element that allows the disconnection of the machine from the corresponding inverter leg when this type of fault occurs [19].
In the second case, by imposing the same terminal voltage to all the phases, the drive is lead to a symmetrical fault state in which all the phase currents have the same steady state magnitude. In this case, the phase currents are limited by the high phase inductance to a value close to that of the rated current, allowing the fault to be thermally sustained for a long time. The high torque pulsations that result from the fault decrease exponentially until only a small breaking torque remains. The major concern in this case is related to the high currents that result during the transient operation until the safe steady state is reached that are likely to generate high demagnetizing fields within the machine. In order to fully assure the integrity of the PMSM drive, the detection and protection circuits must act quickly enough so that the remaining healthy transistors are not damaged, the PMs are not demagnetizing and the machine windings are not put at risk. In any case, the adequacy of provoking a symmetrical short-circuit fault instead of directly switching off the inverter when a PMSM machine is supplied by a half-bridge inverter has been demonstrated with the conducted simulations.

6.3 Conclusions

In the present chapter, the post-fault operation of a PMSM drive subject to different faults has been investigated. In addition to reviewing previously proposed modified control strategies, a general approach to compute suitable current references for the post-fault operation under winding open-circuit faults and terminal short-circuit faults has been proposed. The method, described in Appendix F of the present thesis, is aimed at minimizing the stator copper losses while maintaining the main harmonic of the air-gap MMF waveform.

It has been checked via FE simulations and dynamic simulations involving the PMSM drive model developed in the previous chapter that the proposed post-fault control strategies can improve the performance of the considered PMSM drive operating after a fault. The price to be paid in order to keep the drive’s torque capability is an increase in machine losses and the need for the drive to be overrated in terms of available voltage and current withstand capability of the power semiconductor switches. It has been demonstrated that the constraints imposed by the available current and voltage limitations can seriously limit the mean torque deliverable under fault conditions. In this sense, winding open-circuit faults involving two adjacent phases and terminal short-circuit faults between two non-adjacent phases may impose the stringer limitations. In future fault-tolerant PMSM designs, more attention should be paid to the aforementioned limits; specially to the constraint imposed by the available DC bus voltage.

It has been shown that the use of a full H-bridge inverter can improve the performance of a PMSM drive under fault conditions by reducing the stator copper losses, lessening the torque ripple and decreasing the necessary current and voltage overratings. Although the supply from a full H-bridge inverter has been thoroughly treated only in the case of a single phase winding open-circuit fault, the proposed method allows to easily compare the performance of the drive when supplied from a half-bridge or a full H-bridge inverter.
6.3. CONCLUSIONS

Regarding inverter faults in which a power switch is permanently closed, it has been shown that such a fault is so dramatic that the only possible remedial action, when an standard half-bridge inverter is considered, is to render the machine to a symmetrical short-circuit state. The adequacy of this remedial strategy has been checked by comparing the drive’s transient response after a switch short-circuit fault when the inverter is shut down or when the aforementioned strategy is applied. In checking the dynamic response of the designed PMSM drive after such a fault or after other fault scenarios, the 5-phase PMSM drive model developed in the previous chapter has proven to be useful.
Chapter 7

Experimental Results

A number of tests have been conducted on the designed prototype in order to check the soundness of the design, validate the proposed fault-tolerant PMSM design methodology and evaluate the adequacy of the studied post-fault remedial strategies. In the following chapter, the experimental setups used in the different tests are described and the results for each test are given. After a brief discussion of the obtained results, the conclusions for the chapter are stated.

7.1 Experimental setup

Two different test benches have been used in the characterization and testing of the designed 5-phase PMSM. In the first one, the PM machine has been driven by an asynchronous machine in order to perform the standard open-circuit and short-circuit tests. The second test bench, in which the machine has been operated in motoring mode, has been employed to conduct various dynamic tests, both under healthy and fault conditions.

7.1.1 Setup for standard open-circuit and short-circuit tests

A first batch of tests have been conducted for the designed 5-phase PMSM driven by a 5 kW, four pole asynchronous machine. The asynchronous machine has been vector controlled with a closed loop speed control. The drive supplying the asynchronous machine is a 36 kW modular servo controller making use of a standard 3-phase half-bridge inverter operating at a switching frequency of 10 kHz (Baumüller b maXX 4000). Both open-circuit and short-circuit tests have been conducted with the machine phase windings forming a star connection. A picture of the test bench used in the characterization of the prototype PMSM is shown in Figure 7.1.

During the tests, the speed of both machines has been obtained by filtering the rotor position, measured at the 5-phase PMSM via a 1000 pulses/rev. incremental encoder (Omron E6B2-CWZ1X 1000P/R 2M). The phase to neutral voltages have been measured by means of standard voltage probes, whereas the measurement of the phase currents has been accomplished with the aid of 5 Hall effect 30 A_{RMS} current probes (LEM PR30). A 50 N·m torque transducer placed in between both machines has been used in all the measurements. Additionally, the machine winding temperatures have been measured by employing 10 resistance thermometers (Pt100). Five temperature sensors have been placed inside a number of slots in the machine (one per phase winding), whereas the other five have been located in the machine end-windings. All the analog signals have been recorded with a multichannel data acquisition system with a maximum sampling rate of $100 \cdot 10^6$ samples/s (Yokogawa SL-1000).
7.1.2 Setup for dynamic tests

A second set of tests has been conducted for the 5-phase PMSM operating in motoring mode. For these tests, the machine has been coupled to a dynamometric brake composed of an eddy-current and powder brake dynamometers mounted in tandem configuration. The employed dynamometric set (Magtrol 2WB115+2PB115) allows to operate in between 0 and 15000 rpm and absorb loads up to 30 kW. The machine has been supplied by two 3-phase rectifier + IGBT inverter modules (Semikron SEMITEACH - IGBT), from which five of the total six inverter legs have been used. Each inverter stack can be supplied either with AC or DC voltages and deliver phase currents up to 30 A_{RMS}. The inverter drive gate signals have been controlled by a Digital Signal Processor (DSP) unit (dSPACE DS1104), programmable via a serial communication line. In all the dynamic tests, a DC bus voltage of 300 V and a constant switching frequency of 5 kHz have been used. The experimental setup used for the PMSM operating in motoring mode is shown in Figure 7.2.

A current control in the $\alpha - \beta$ reference frame has been implemented in the DSP unit. Standard PI controllers with no anti-windup control and without feedforward compensation have been used in the control of the currents. The reason for this is that the system became unstable when trying to implement
these features. A closed loop speed control has been added to the control scheme in order to test the drive at various speeds.

In the dynamic tests, the phase currents, the rotor position and the drive’s output torque have been recorded and fed to the control system driving the two 3-phase inverters (dSPACE DS1104). Owing to the limited number of analog input channels supported by the employed control unit, it has not been possible to record any other variable simultaneously. All the analog signals have been sampled at a frequency of 5 kHz.

7.2 Machine characterization

7.2.1 Phase resistance and inductance measurements

The winding impedance of each machine phase has been measured by employing a common LCR meter (Hioki 3522-50 LCR HiTESTER). In addition to measuring the DC phase resistance, additional measurements have been conducted for various frequencies. No significant unbalance among the values obtained for each phase has been identified. The mean measured phase resistance and inductance values are collected in Table 7.1.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Phase resistance, $R_{ph}$ [Ω]</th>
<th>Phase inductance, $L_{ii}$ [mH]</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC</td>
<td>0.894</td>
<td>−</td>
</tr>
<tr>
<td>50 Hz</td>
<td>1.015</td>
<td>17.29</td>
</tr>
<tr>
<td>100 Hz</td>
<td>1.237</td>
<td>16.98</td>
</tr>
<tr>
<td>200 Hz</td>
<td>1.576</td>
<td>17.00</td>
</tr>
</tbody>
</table>

Table 7.1: Mean values for phase resistance and inductance measurements at different frequencies

The measurements have been conducted for a winding temperature of about 20 °C. Compared to the resistance and inductance values estimated during the design stage of the machine: $R_{ph} = 0.773$ Ω (DC at 20 °C), $L_{ii} = 18.88$ mH, it can be appreciated that there is a significant drift in the values of these parameters. The higher value of the phase resistance is most likely due to the fact that the prototype machine has larger end-windings than expected and due to the added resistance of the output cables, the terminal connections, etc. that during the design stage have not been considered. The measured phase inductance is about 10% lower than expected.

In order to doublecheck the obtained results, the phase impedance has been measured with a different LCR meter as well (Fluke PM6306). The conducted measurements have lead to very similar resistance values, but lower phase inductance magnitudes. In this case, the mean measured winding inductance value has been of $L_{ii} = 15.75$ mH, with almost no variation at different frequencies (in the range 50 – 100 Hz). The obtained results do not offer much certainty regarding the measurement of the phase inductances and suggest that measuring them by a common LCR meter is not totally appropriate.

7.2.2 Open-circuit test

The first test that has been conducted for the PMSM being driven by the asynchronous machine is the measurement of the back-EMF. In this test, the machine phase windings have been star connected and the remaining phase terminals have been left open-circuited. The test has been performed for a number of speeds, ranging from 0 to 1000 rpm. The obtained induced voltage waveform is shown in Figure 7.3a for the rated speed of 600 rpm. The predicted vs measured back-EMF values for an ambient temperature of 20 °C are displayed in Figure 7.3b. From this test, the PM flux linkage values collected in Table 7.2 have been derived.

In observing the results, it can be concluded that the obtained results are highly satisfactory. Only a slight decrease in the main harmonic and a minor increase in absolute terms in the third harmonic of the PM flux linkage are observed. The slight differences between the previously estimated and measured results can be due to a number of factors, such as different characteristics of the employed PM and iron materials, manufacturing tolerances, etc.
CHAPTER 7. EXPERIMENTAL RESULTS

(a) Measured back-EMF waveform at 600 rpm

(b) Back-EMF vs speed characteristic

Figure 7.3: Open-circuit test results ($T_{PM} = 20 \, ^\circ\text{C}$)

<table>
<thead>
<tr>
<th>PM flux linkage (1st harm.), $\Lambda_{m1}$ [mWb$_{RMS}$]</th>
<th>Measured</th>
<th>Estimated</th>
</tr>
</thead>
<tbody>
<tr>
<td>PM flux linkage (3rd harm.), $\Lambda_{m3}$ [mWb$_{RMS}$]</td>
<td>1.453</td>
<td>0.107</td>
</tr>
<tr>
<td>PM flux linkage (5th harm.), $\Lambda_{m5}$ [mWb$_{RMS}$]</td>
<td>0.345</td>
<td>0.428</td>
</tr>
<tr>
<td>PM flux linkage (7th harm.), $\Lambda_{m7}$ [mWb$_{RMS}$]</td>
<td>0.023</td>
<td>0.001</td>
</tr>
</tbody>
</table>

Table 7.2: PM flux linkage harmonics ($T_{PM} = 20 \, ^\circ\text{C}$)

7.2.3 Short-circuit test

A second test has been conducted on the prototype PM machine by connecting all the phase windings together at their input and output terminals. Therefore, the voltage drop across all the phase windings is forced to be the same and a symmetrical short-circuit is established. The test has been performed for a number of speeds, ranging from 0 to 1500 rpm. The main measured variables have been the phase short-circuit currents and the braking torque. If only the first harmonic of the PM flux linkage is considered and purely sinusoidal currents are assumed, such variables can be analytical estimated from the following equations [293]:

\[
I_{sc} = \frac{\omega_e \Lambda_{m1}}{\sqrt{R_{ph}^2 + (\omega_e L_{ph})^2}} \quad (7.1)
\]

\[
T_{brk} = -\frac{m p R_{ph} \Lambda_{m1} \omega_e}{R_{ph}^2 + (\omega_e L_{ph})^2} \quad (7.2)
\]

where it has been assumed that: $L_{d1} = L_{q1} = L_{ph}$. It must be noted that the breaking torque calculated according to equation (7.2) corresponds solely to the electromagnetic torque resulting from the interaction between the magnetic fields created by the PMs and the stator currents. Friction, iron and magnet losses are not accounted by the previous formulas, but also oppose the motion of the rotor, causing the breaking torque measured at the machine’s shaft to be larger.

The predicted versus measured current and breaking torque values for different speeds are displayed in Figure 7.4. The “predicted” values refer to those calculated during the design stage of the machine and, thus, they make use of the machine parameters presented in Chapter 4. A second set of values has been calculated from the resistance and PM flux linkages measured in previous tests ($R_{ph} = 0.894 \, \Omega$, $\Lambda_{m1} = 106.0 \, \text{mWb}_{RMS}$), together with the phase inductance obtained from the experimentally measured characteristic current ($I_{ch} = 5.251 \, A_{RMS}$), $L_{ph} = \Lambda_{m1}/I_{ch} = 20.19 \, \text{mH}$. In Figure 7.4, these values are labeled as “estimated”. In all the calculations, equal winding and magnet temperatures have been assumed ($T_w = T_{PM} = 20 \, ^\circ\text{C}$) and the effect of the stator resistance increase with frequency has been ignored.
7.3 Tests under healthy operating conditions

In order to verify the adequate operation of the PMSM drive, a number of tests have been conducted for the PM machine operating in motoring mode. A closed loop speed control has been established and the drive’s speed has been varied from 0 to 600 rpm. By varying the reference speed and the load absorbed by the dynamometric brake, the drive has been tested at various torque and speed conditions. The measured current, torque and speed waveforms for the machine operating under rated torque conditions and a speed reference of 175.1 rpm are shown in Figure 7.5. The average measured output torque is $T_{\text{shaft,mean}} = 26.40 \text{ N\cdotm}$, with an approximate peak-to-peak ripple of $\Delta T_{\text{shaft}} = 3.543 \text{ N\cdotm}$. The absorbed phase currents have a mean RMS value of: $I_{\text{mean}} = 4.616 \text{ A}_{\text{RMS}}$. Assuming these correspond solely to current in the $q_1$-axis ($i_{q_1} = \sqrt{2}I_{\text{mean}}$) and, by considering only the first harmonic of the flux linkage due to the PMs ($\lambda_{m_1} = 106.0 \text{ mWb}_{\text{RMS}}$), an electromagnetic torque of 26.91 N-m is obtained. By comparing the measured and estimated torque values, the machine can be considered to be adequately characterized. The little discrepancies are due to a number of factors, including the additional braking torque caused by other loss components besides stator copper losses, magnet temperature increase, calibration of the current and torque transducers, etc.
CHAPTER 7. EXPERIMENTAL RESULTS

Figure 7.5: Experimental results under healthy operating conditions (speed reference of 175.1 rpm)

The high ripple observed in the phase current waveforms is due to noise in the current sensor measurements, on one hand, and due to the inability of the currents to perfectly follow their references, on the other. This issue is exemplified in Figure 7.6, in which the measured phase currents referred to the $\alpha - \beta$ reference frame and the established current references are displayed.

The measured currents show small oscillations and a little delay with respect to the reference currents, indicating that the bandwidth of the employed current controllers is low. The delay between currents is more evident at higher speeds, for which the current waveforms cannot adequately follow their references. The low bandwidth of the current controllers has been an issue that has not been possible to solve. Increasing the PI current controller gains resulted in the system becoming unstable and the current sampling rate could not be increased due to hardware limitations. Owing to this, it has not been possible to adequately test the drive at rated speed conditions and, therefore, the rest of experimental tests have been conducted at a lower speed.

As a consequence of the non-purely sinusoidal phase currents and the higher content of flux linkage harmonics, the measured torque ripple is higher than predicted. Anyhow, the peak-to-peak torque ripple is less than 15% of the average torque, which is not excessive. Furthermore, the closed speed loop and the system’s inertia help filter this ripple, leading to an almost constant drive speed.

7.4 Tests under winding open-circuit faults

A number of experiments have been conducted in order to check the drive’s response to different winding open-circuit faults. The same fault scenarios as in previous chapters have been considered: the open-circuit of a single phase, the loss of two adjacent phases and the open-circuit of two non-adjacent phases. For each fault scenario, the operation when no remedial actions after the fault are taken and when the...
control strategy is modified accordingly has been checked. In all cases, a constant DC bus voltage of 300 V has been employed.

7.4.1 Open-circuit of a single phase

Operation with no remedial action

The open-circuit of a single phase winding has been emulated by placing a circuit breaker between the inverter output and the machine terminals of phase A. First, the system has been driven to a speed of 100 rpm with the dynamomemetric break imposing a constant load of 20 N·m. Then, phase A has been open-circuited by suddenly opening the series connected circuit breaker. The fault has caused the torque waveform to become pulsating and has led to a drop in the average torque and speed. In order to compare the drive’s response under different fault scenarios at similar conditions, the speed reference has been increased until the drive has reached a speed close to 100 rpm. The measured current, torque and speed waveforms at open-circuit steady state conditions are shown in Figure 7.7, whereas the measured and reference $\alpha - \beta$ currents are displayed in Figure 7.8.

![Figure 7.7: Experimental results for single phase open-circuit fault with no remedial action](image)

![Figure 7.8: Reference and measured currents after single phase open-circuit fault with no remedial action, $[i_{\alpha,\beta}]$](image)

In observing the figures, it can be easily appreciated that the fault causes the system not to follow the current references anymore. As a consequence, the phase currents turn non-sinusoidal, the torque becomes pulsating and the speed ripple is increased. Significant third harmonic currents are absorbed by the machine. The measured mean torque is of $T_{\text{shaft,mean}} = 20.24$ N·m, with a peak-to-peak ripple of $\Delta T_{\text{shaft}} = 26.86$ N·m. As in the simulations presented in Chapter 5 (see Figure 5.10), the dominating torque harmonic is the 2nd order one. It must be noted, though, that different current control strategies have been considered in the two cases. In order to compensate for the fault, the healthy phase current magnitudes increase to a mean value of $I_{\text{mean}} = 6.031 \ A_{\text{RMS}}$; being the currents in the phases adjacent...
to the faulted one the ones to increase most. Even if the drive is able to operate under fault conditions, the large torque pulsations and the increase in copper losses demand that appropriate remedial actions are taken.

**Operation by modifying the control strategy**

In order to improve the system’s behavior after the fault, the current references to be followed have been modified as discussed in the previous chapter. In this case, the modified control strategy aimed at minimizing the stator copper losses while maintaining the main harmonic of the armature MMF has been applied; see equation (6.1). It is important to note that no fault detection algorithm has been implemented, but the change in the control strategy has been done manually. The current, torque and speed waveforms obtained by modifying the control strategy with no torque derating \((\chi = 1)\) are shown in Figure 7.9, whereas the measured currents referred to the \(\alpha - \beta\) reference frame are compared to their references in Figure 7.10.

Despite the increased noise in the phase current measurements, it can be easily appreciated that, by modifying the control strategy, the system recovers its ability to adequately follow the current references. This leads to a significant reduction in the torque and speed pulsations and to a decrease in the healthy phase current magnitudes. In this case, the average measured torque is of \(T_{\text{shaft, mean}} = 19.63\ \text{N} \cdot \text{m}\), with a peak-to-peak ripple of \(\Delta T_{\text{shaft}} = 3.531\ \text{N} \cdot \text{m}\). The mean healthy phase current value decreases to \(I_{\text{mean}} = 4.890\ \text{A}_{\text{RMS}}\); meaning that stator copper losses are reduced compared to the state in which no remedial action is taken after the fault. It is, therefore, demonstrated the soundness of the proposed modified control strategy.

It must be noted that, owing to the reduced drive speed, the output torque attainable by the drive is not limited by the available DC bus voltage. As discussed in the previous chapter, it is expected that,
7.4. TESTS UNDER WINDING OPEN-CIRCUIT FAULTS

when the drive operates at higher speeds, only a fraction of the rated torque remains available due to saturation of the current controllers.

7.4.2 Open-circuit of two adjacent phases

Operation with no remedial action

The second emulated fault scenario has been the open-circuit of two adjacent phases. In this case, two circuit breakers have been connected in series with the inverter outputs of phases $C$ and $D$. However, it has not been possible to attain the same torque and speed conditions as in the previous fault scenario. Whenever a load was demanded by the dynamometric brake, the system became unstable and the drive’s speed dropped to zero. The same happened if, operating at no load conditions, the reference speed was increased to try to achieve an average speed of 100 rpm. As a result, steady state conditions measurements have been performed only for the drive operating at 65 rpm with no external load applied. The measured phase current, torque and speed waveforms are shown in Figure 7.11; whereas the reference and measured $\alpha - \beta$ currents for such a state are displayed in Figure 7.12.

![Figure 7.11: Experimental results for two adjacent phases open-circuit fault with no remedial action](image)

![Figure 7.12: Reference and measured currents with two adjacent phases open-circuit fault with no remedial action, $i_{\alpha \beta}$](image)

As the figures indicate, the fault causes the current references to not be followed in any manner. The currents in the healthy phases hardly resemble a sinusoid anymore and large torque pulsations are obtained. Specifically, a torque ripple of $\Delta T_{\text{shaft}} = 18.79 \, \text{N}\cdot\text{m}$ for a mean output torque of $T_{\text{shaft,mean}} = 1.072 \, \text{N}\cdot\text{m}$ has been measured. The small average torque corresponds to the torque needed to overcome friction, iron and magnet losses. The torque, whose dominant harmonics are the 2nd and 4th order ones, becomes negative during a significant part of the electrical period. As a result, the speed oscillates significantly as well. The measured faulted phase currents exhibit a small drift from zero, indicating that small calibration errors are present in the current sensors.
Operation by modifying the control strategy

The change in the control strategy has allowed the drive to recover its pre-fault controlability. In addition to being able to adequately follow the imposed current references, by modifying the control strategy it has been possible to test the drive with a two adjacent phases open-circuit while delivering a significant mean torque. In Figure 7.13 the measured current, torque and speed waveforms are shown for the PMSM drive operating at 80 rpm with a resistive load of 20 N·m. The applied current references, shown in Figure 7.14, are those of equation (6.4).

![Figure 7.13: Experimental results for two adjacent phases open-circuit fault with modified control strategy](image)

Since the system must compensate for the loss of two phases, the current magnitudes in the healthy phases increase substantially with respect to their pre-fault values ($I_{\text{mean}} = 14.95\, \text{A}_{\text{RMS}}$). This leads to significant saturation harmonics to appear in the torque waveform, whose dominant harmonic is, by far, the second order one. Despite the large torque pulsations, the situation is vastly improved with respect to the state when no remedial actions after the fault are placed. The phase current waveforms are much more sinusoidal and the drive is able to deliver a net torque without becoming unstable. In this case, the measured mean output torque and peak-to-peak torque values are of $T_{\text{shaft,mean}} = 19.94\, \text{N·m}$ and $\Delta T_{\text{shaft}} = 15.79\, \text{N·m}$, respectively. These values are fairly similar to those obtained via FE simulations, as discussed in the previous chapter ($T_{\text{mean}} = 19.82\, \text{N·m}$, $\Delta T = 12.79\, \text{N·m}$, see Table 6.6).

### 7.4.3 Open-circuit of two non-adjacent phases

The open-circuit fault of two non-adjacent phases has been emulated by opening the inverter output cables belonging to phases $B$ and $E$. However, no matter the speed and load conditions, every time the corresponding circuit breakers were opened, the system became unstable and the speed dropped to
zero. Therefore, it has not been possible to perform measurements for the drive operating at steady state conditions under a two non-adjacent phases open-circuit fault without modifying the control law.

The only workaround to obtain some data regarding this fault scenario has been to modify the control strategy prior to opening the corresponding circuit breakers. The applied modified current references are those of equation (6.5) with no torque derating ($\chi = 1$). By doing so, the currents in phases $B$ and $E$ have effectively dropped to zero before emulating the fault and the system has retained its controllability. After opening the circuit breakers, the speed and load references have been varied until similar conditions to the ones in the previous tests have been obtained. Figure 7.15 shows an example of the measured current, torque and speed waveforms for such conditions. Since they do not offer much meaningful information, the waveforms of the $\alpha - \beta$ currents and the current references are not shown in here. As in the fault scenario involving the loss of two adjacent phases, the current references are quite adequately followed with just a small delay and a little ripple due to noise in the phase current measurements.

Figure 7.15: Experimental results for two non-adjacent phases open-circuit fault with modified control strategy

Again, the measured phase currents waveforms are quite sinusoidal and the speed ripple low. The system is capable of delivering a significant amount of torque with a small ripple. In this case, the measured mean torque and peak-to-peak torque ripples are of $T_{\text{shaft,mean}} = 19.61$ N·m and $\Delta T_{\text{shaft}} = 5.986$ N·m, respectively, with an average RMS current magnitude of $I_{\text{mean}} = 7.375$ A$_{\text{RMS}}$. Since the peak phase currents needed to deliver such a mean torque are lower than in the case the open-circuited phases are adjacent, the torque ripple arising from saturation of the iron material is lower. In the same way as in the previous fault conditions, the experimentally obtained results are comparable in order of magnitude to those obtained via FE simulations ($T_{\text{mean}} = 24.18$ N·m, $\Delta T = 7.339$ N·m, see Table 6.7).

7.5 Tests under terminal short-circuit faults

The fault-tolerant operation of the 5-phase PMSM drive under terminal short-circuit faults has been investigated next. Specifically, the drive’s response after the short-circuit of a whole phase winding and the short-circuit between the terminals of two phases, either adjacent or non-adjacent, has been examined.

7.5.1 Short-circuit of a single phase

Operation with no remedial action

The sudden short-circuiting of a whole phase winding has been emulated by placing a circuit breaker between both terminals of phase $A$. At some point in time, the circuit breaker has been closed causing the desired fault conditions. As in previous fault scenarios, the sudden short-circuiting of phase $A$ has caused the system to become unstable whenever a resistive torque has been applied by the dynamometer brake. Hence, steady state operation has only been possible at no load conditions. The measured current, torque and speed waveforms are shown in Figure 7.16. Owing to the reduced number of analog inputs allowed
by the employed DSP unit, it has not been possible to record the fault current in the short-circuited phase simultaneously with the rest of electrical variables. Since the calculation of the currents referred to the $\alpha - \beta$ reference frame involves synchronously measuring the currents really flowing through all phase windings, measurements for the $[i_{\alpha \beta}]$ currents have not been obtained.

As it can be appreciated from the previous figure, the terminal short-circuit of a single phase leads to significant noise in the phase current measurements. The torque becomes highly pulsating, leading to a significant speed ripple. For a measured mean torque of $T_{\text{shaft,mean}} = 0.370$ N-m, the torque ripple and mean branch current have been of $\Delta T_{\text{shaft}} = 30.46$ N-m and $I_{\text{mean}} = 7.386$ A$_{RMS}$, respectively. Undoubtedly, is the branch current belonging to the faulted phase the one that increases most ($I_{A,\text{branch}} = 16.07$ A$_{RMS}$).

**Operation by modifying the control strategy**

In order to improve the drive’s response after the fault, the control strategy has been modified as established in equation (6.9) and the fault has been isolated by opening the corresponding inverter leg, in that order. Doing it the other way around caused the system to become unstable. By applying the proposed remedial actions, it has been possible to test the drive while delivering a net output torque. The measured current, torque and speed waveforms for a speed of 95 rpm and a braking torque of 30 N-m, are shown in Figure 7.17.

Despite the high noise in the current measurements, it can be appreciated how the phase currents recover their sinusoidal looking waveforms. As a result, the drive is able to deliver a mean torque of $T_{\text{shaft,mean}} = 29.47$ N-m with a moderate ripple, $\Delta T_{\text{shaft}} = 7.518$ N-m. In this case, the dominant torque harmonics are the second and fourth order ones.
7.5. TESTS UNDER TERMINAL SHORT-CIRCUIT FAULTS

The waveform of the short-circuit current has been recorded by employing the same multichannel data acquisition system used in the characterization of the machine. The data has been recorded, though, for a slightly lower drive speed (84.29 rpm). The measured and predicted short-circuit currents waveforms for such operating conditions are shown in Figure 7.18. The delay between both signals stems from the fact that they have not been recorded synchronously and it is, thus, artificial.

As shown in the previous figure, the measured and predicted short-circuit currents match quite well in magnitude. Higher order harmonics are present in the measured short-circuit current as a consequence of PM flux linkage harmonics and iron saturation. A small magnetic coupling between machine phases may be another reason for the increased harmonic content. Anyhow, the estimation of the short-circuit current waveform is good enough and the suitability of the proposed modified control strategy is demonstrated.

7.5.2 Short-circuit of two adjacent phases

A short-circuit fault between two adjacent phases has been emulated next. In order to protect the inverter and not cause a DC bus shoot-through, circuit breakers have been connected in series with the faulted phase terminals and they have been opened before causing the short-circuit fault. As in previous fault scenarios, owing to the system becoming unstable after the fault, only steady state measurements for the drive operating with no externally applied load have been conducted. The measured current, torque and speed waveforms for an average speed of 53.55 rpm are shown in Figure 7.19.

Like in the previous fault scenarios, the drive losses its ability to properly follow the current references and the resulting phase currents hardly resemble a sinusoid anymore. Significant torque harmonics are introduced, of which the first, the second and the fourth are the most important ones. As a consequence of the high and fast torque pulsations, a significant speed ripple arises. As in the fault scenario involving the open-circuit of two adjacent phases, a small drift from zero of the measured faulted phase branch...
currents can be appreciated. For a measured mean torque of $T_{\text{shaft,mean}} = 1.021 \text{ N·m}$, the torque ripple and mean branch current are of $\Delta T_{\text{shaft}} = 20.03 \text{ N·m}$ and $I_{\text{mean}} = 3.355 \text{ A}_{\text{RMS}}$, respectively.

In this case, it has not been possible to test the drive with the proposed modified control strategy. Whenever the current references were modified, the system became unstable after a few seconds and the drive stopped. This is thought to be due to the low bandwidth of the employed current controllers, although this matter should be examined in greater detail in the future.

7.5.3 Short-circuit of two non-adjacent phases

Operation with no remedial action

The results for the drive operating at no load conditions and under a short-circuit fault between two non-adjacent phases are similar to those when the faulted phases are adjacent. In this case, the terminals of phases $B$ and $E$ have been short-circuited after opening the corresponding input cables by means of two circuit breakers. The drive is not able to deliver a net torque other than the one necessary to oppose the braking torque caused by other loss components besides stator copper losses and operation beyond 54 rpm is not possible. Compared to the previous fault scenario, the current magnitudes in the healthy phases are higher and larger torque and speed ripples are obtained. In particular, a mean output torque of $T_{\text{shaft,mean}} = 1.242 \text{ N·m}$ and a peak-to-peak torque ripple of $\Delta T_{\text{shaft}} = 41.39 \text{ N·m}$ have been measured for a mean healthy phase current magnitude of $I_{\text{mean}} = 3.872 \text{ A}_{\text{RMS}}$. The obtained current, torque and speed waveforms are displayed in Figure 7.20. In this case, the dominating torque harmonic is the second order one.

![Figure 7.20: Experimental results for single phase open-circuit fault with no control modification](image)

Operation by modifying the control strategy

By modifying the current references as established in equation (6.22), a vast improvement in the drive’s controllability has been obtained. It has been possible to test the drive at steady conditions for a speed of 88 rpm and a resistive torque imposed by the dynamometric brake of 20 N·m. The experimentally obtained current, torque and speed waveforms are collected in Figure 7.21.

In this case, since the system has to compensate for the loss of two phases, the healthy phase current magnitudes increase significantly with respect to their pre-fault values ($I_{\text{mean}} = 14.36 \text{ A}_{\text{RMS}}$). This causes a substantial saturation of the iron material, which, in turn, leads to large torque pulsations. Anyhow, despite the considerable torque ripple, the situation is vastly improved with respect to the state when no remedial actions after the fault are placed. The phase current waveforms become more sinusoidal and the drive is able to deliver a net torque without becoming unstable. In this case, the measured mean output torque and peak-to-peak torque values are of $T_{\text{shaft,mean}} = 19.75 \text{ N·m}$ and $\Delta T_{\text{shaft}} = 19.50 \text{ N·m}$, respectively. The comparison between the measured and predicted short-circuit current magnitudes is similar to that of the single phase short-circuit case and, thus, it is not repeated here.
7.6 Tests under transistor open-circuit faults

7.6.1 Opening of a single transistor

The last conducted dynamic test has been the emulation of a single transistor open-circuit fault. In this case, from a certain point onwards, a 0 signal has been permanently commanded to the high side switch of phase \( D \), while the low side switch has been left to continue switching according to the gate signals computed by the control unit. The obtained phase current, torque and speed waveforms are collected in Figure 7.22, whereas the reference and measured currents referred to the \( \alpha - \beta \) reference frame are shown in Figure 7.23.

The experimental results are very close to those regarding the winding open-circuit of a single phase. In this case, despite the low side switch being left to operate normally, the emulated fault conditions have led the faulted phase to not conduct current during any time on the electrical period. As it can be observed in the previous figures, the drive is able to deliver a net output torque, but the current references are not appropriately followed. This leads to an increase in the phase current harmonic content and to high torque pulsations. The measured mean torque and peak-to-peak torque ripples are of \( T_{\text{shaft,mean}} = 20.02 \text{ N}\cdot\text{m} \) and \( \Delta T_{\text{shaft}} = 21.21 \text{ N}\cdot\text{m} \), respectively; being the first and second order torque harmonics the dominant ones.

Owing to time constraints and, since the operation of the drive is very close to that of a single phase open-circuit fault, the operation when the control strategy is modified has not been tested. It has been assumed that both the torque pulsations and the stator copper losses can be reduced by modifying the current references, as demonstrated for the single phase open-circuit fault of phase \( A \). In successive experiments, this fact should be checked.
Two tasks have been accomplished in the present chapter. On one hand, the designed 5-phase PM machine prototype has been characterized via a number of measurements and tests. Specifically, the winding resistance, the phase inductance and the flux linkage due to the PMs have been obtained ($R_{ph} = 0.894 \, \Omega$, $\Lambda_{m1} = 106.0 \, \text{mWb}_{RMS}$ and $L_{ph} = 20.19 \, \text{mH}$). The previous set of parameters has proven to be useful in accurately predicting various electrical variables, including the healthy state electromagnetic torque and the fault currents arising after a short-circuit fault. The differences between the experimentally obtained and the predicted machine parameters are small and serve to validate the proposed fault-tolerant PMSM design methodology. Additionally, the machine has proven its intrinsic ability to limit the short-circuit currents.

On the other hand, the post-fault operation of the designed machine prototype has been experimentally investigated for a number of fault scenarios. For most fault conditions, the post-fault operation when no remedial actions after the fault are placed and when the control strategy is modified in order to compensate for the fault have been examined. The summary for the conducted tests is collected in Table 7.3. In order to allow for a fair comparison between different fault conditions to be made, the torque ripple and copper loss values have been given as a ratio of the measured mean torque.

<table>
<thead>
<tr>
<th>Fault Type</th>
<th>$\frac{\Delta T_{\text{shaft}}}{T_{\text{shaft,mean}}}$</th>
<th>$\frac{\sum I^2_{\text{branch,RMS}}}{T_{\text{shaft,mean}}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Healthy</td>
<td>0.134</td>
<td>4.035</td>
</tr>
<tr>
<td>Open-circuit of a single phase</td>
<td>1.327</td>
<td>7.306</td>
</tr>
<tr>
<td>Open-circuit of two adjacent phases</td>
<td>17.52</td>
<td>23.04</td>
</tr>
<tr>
<td>Open-circuit of two non-adjacent phases</td>
<td>--</td>
<td>0.305</td>
</tr>
<tr>
<td>Short-circuit of a single phase</td>
<td>82.22</td>
<td>992.6</td>
</tr>
<tr>
<td>Short-circuit of two adjacent phases</td>
<td>19.62</td>
<td>33.56</td>
</tr>
<tr>
<td>Short-circuit of two non-adjacent phases</td>
<td>33.33</td>
<td>37.41</td>
</tr>
<tr>
<td>Opening of a single transistor</td>
<td>1.059</td>
<td>5.232</td>
</tr>
</tbody>
</table>

Table 7.3: Summary of dynamic test results

As the results indicate, all faults lead to an increase in the torque pulsations and to higher stator copper losses compared to the healthy state. However, for most fault conditions, a notable reduction in these parasitic effects can be obtained by adopting the proposed modified control strategies. Furthermore, the change in the control law has allowed the drive to operate while delivering a net torque for a number of fault conditions in which the system previously became unstable. The only fault scenario for which the proposed remedial actions have not led to an improvement in the drive’s response has been the terminal short-circuit fault between two adjacent phases. This fault operation mode should be further investigated in the future. For the rest of fault scenarios, the experimental results validate the adequacy
of the proposed modified control strategies.

A limitation of the conducted experimental tests is that the employed current controllers bandwidth was low and the dynamic tests could not be performed for the rated speed conditions. This matter should be addressed in the future and new tests be carried out for the drive operating at rated speed conditions. It is expected that by operating at a higher frequency, the constraint imposed by the available DC bus voltage comes into view.
Chapter 8

Research Findings and Future Work

The main objective of the present thesis has been to develop the necessary tools for the design and analysis of fractional-slot concentrated-winding fault-tolerant permanent magnet synchronous machines. Additionally, owing to the various advantages they offer, the use of multiphase machines, in which the number of phases is higher than three, has been contemplated. The following chapter collects the main findings derived from the conducted research work. It summarizes the tasks undertaken during the development of the thesis, together with the achievements accomplished on each area. In addition to stating the main conclusions, future research lines that have emerged from the conducted work are proposed.

8.1 Findings of the investigation

8.1.1 Regarding the state of the art in fault-tolerant PMSM drives

In Chapter 2, the different methods to grant fault-tolerance to a PMSM drive have been discussed and their benefits and drawbacks examined. The key points arising from such a comparison are stated below:

- Solutions based on redundant machine and inverter structures can be made fault-tolerant to any possible single element fault with no need to modify the control law after the fault, but their cost is large.

- Modified inverter topologies have been proposed to add some post-fault functionality over the standard 3-phase half-bridge inverter, but either their post-fault performance is poor (switch-redundant topology, additional leg connected to neutral point) or the increase in power converter cost substantial (additional redundant leg, direct AC/AC matrix converters).

- Machine designs based on single phase modular and multiple 3-phase configurations offer the best compromise between post-fault capability and overall system cost. In particular, single phase modular designs have proven to be particularly effective as they allow the drive to sustain multiple electrical faults. In order for a multiple-stator design to achieve the same feature, the number of phase windings must be increased to 9, at least.

- The discussed methods are not rigid and the different fault-tolerant approaches can be combined. The best choice is always the one that leads to the desired degree of fault-tolerance at minimum cost.

From the previous analysis, it has been decided to orient the research work towards single phase modular PM machine designs. The requirements and penalties arising from this fault-tolerant method have been reviewed and the characteristics of the winding topologies commonly employed in these designs have been studied. Since they are commonly employed in fault-tolerant designs, the main properties of multiphase AC systems have been examined as well.
8.1.2 Regarding the winding selection for FSCW fault-tolerant PMSMs

Fractional-slot concentrated-winding machines are known for allowing to choose among a great number of combinations with reference to the number of phases, poles and slots. In Chapter 3, the main design principles for selecting the proper winding for a fault-tolerant PMSM have been presented. The basic requirements are that the winding inherently leads to a high phase self-inductance and a low mutual inductance between phases. Reviewing previously proposed criteria for selecting FSCW combinations that are adequate for fault-tolerant designs, it has been found out that the traditional rules are restricted to odd phase number machines or to specific winding configurations. Moreover, no numerical comparison is generally established among the most promising combination candidates regarding the values of the mutual inductances in the machine. In order to fill this gap, the relationships between the self-inductance and the mutual inductance values have been studied for radial-flux PMSMs with either one, two or four layer windings. In the investigations an analytical procedure has been employed. The main findings for the conducted research are stated below:

- For fault-tolerant applications, the most promising FSCW combinations are those for which the number of spokes per phase in the star of slots \( (Q_{mt}) \) is even and for which the number of poles is close to the number of slots \( (2p) \approx Q \).
- Single-layer windings offer better isolation characteristics than double-layer FSCWs, but the effect of iron saturation on the magnetic coupling between phases must be addressed.
- For single-layer surface-mounted permanent magnet machines, the beneficial effects arising from choosing combinations for which \( 2p \approx Q \) may be masked by the large effective air-gap.
- It is possible to design double-layer FSCW machines with a low mutual coupling between phases when the number of phases is odd. In the case of surface-mounted permanent magnets, since the slot leakage inductance carries a significant weight on the total inductance, it is particularly important to select combinations for which \( 2p \approx Q \).
- It has been found out that even phase number double-layer windings and FSCWs having 4-layers are not suitable for fault-tolerant applications, as they lead to significantly larger mutual couplings between phases than other winding configurations.
- Two numerical factors that are solely a function of the winding layout have been introduced for the selection of the optimal phase, pole and slot number combination for FSCW PMSMs. These factors allow the most promising combinations to be chosen regardless of the geometry of the machine.
- In addition to validating the above criteria by an application example and a Finite Element analysis, the most promising machine candidates in terms of fault-tolerance for a given ISA application have been obtained.
- In the selection of the optimal winding layout for a fault-tolerant FSCW machine, the effects on the radial forces, the rotor losses, the cogging torque, etc. must be addressed as well. Obviously, the importance given to each particular aspect will be highly dependent on the particular fault-tolerant application at hand.

8.1.3 Regarding the design of a FSCW multiphase fault-tolerant PMSM

In Chapter 4, a general methodology for the design of FSCW multiphase fault-tolerant PMSMs has been proposed. The design procedure, based on previous work carried out in the research group, has been adapted to account for the particularities of these type of machines. The main points to highlight regarding the work conducted on the design of fault-tolerant machines are:

- The proposed methodology is general and valid for machines of any size and power rating. It allows to design machines with any number of phases and account for different design choices, such as the rotor topology, the waveform on the back-EMF, imposed geometrical constraints, etc. From a small set of basic specifications, all the geometrical and constructional aspects of the machine are obtained.
8.1. Findings of the Investigation

- In the sizing and in the electromagnetic performance evaluation of the machine, an analytical formulation is used. This allows a number of geometries to be quickly sized and calculated, minimizing the amount of time required by the design process. Parameter sensitivity analysis can be very easily conducted, leading to an increased knowledge about the influence of certain variables in the performance characteristics of the machine.

- From the housing of a small power asynchronous machine, a fault-tolerant 5-phase PMSM has been designed. Key aspects in the design of the prototype have been addressed, including the definition of the machine specifications, the selection of suitable electric and magnetic loadings, the optimization via FE analysis, etc. The prototype machine exhibits a substantial torque density increase over the original asynchronous machine.

- Experimental results have demonstrated the intrinsic fault-tolerant capability of the designed machine prototype. The good agreement between the experimentally obtained and the predicted machine parameters demonstrate the validity of the proposed fault-tolerant PMSM design methodology.

8.1.4 Regarding the development of a FSCW multiphase PMSM drive model

In Chapter 5, a five-phase FSCW PMSM drive model suitable for fault analysis has been developed. The key points regarding the development of the model and its use are summarized below:

- The model, while simple and fast to simulate, retains the main machine characteristics and is able to accurately predict the behavior of the PMSM drive as long as the effect of iron saturation is not important.

- The derived model has proven to be useful in predicting the behavior of a 5-phase PMSM drive subject to winding open-circuit, terminal short-circuit and inverter faults. Additionally, it has allowed to check the response of the PMSM drive when different remedial strategies after each fault are introduced.

8.1.5 Regarding the fault-tolerant operation of a PMSM drive

In Chapter 6, the remedial actions aimed at mitigating the parasitic effects arising from the operation in fault conditions for a PMSM have been reviewed. Specifically, the fault-tolerant operation of the drive consisting on the designed PMSM prototype and a five-phase half-bridge inverter has been investigated. The conducted research work can be summarized as follows:

- A unified method to compute the suitable current references for PMSM drives subject to winding open-circuit and terminal short-circuit faults has been derived. The method, aimed at minimizing the stator copper losses while preserving the main harmonic of the air-gap MMF, is valid for both the supply from a half-bridge and a full H-bridge inverter. It can be readily applied in the fault-tolerant operation of other phase number drives as well. As the main harmonic of the MMF is preserved, the increase in stator iron and rotor losses under fault conditions has been checked to be minimal for the designed prototype.

- The coefficients that relate the first and the third harmonic current references to be applied in post-fault operation mode are constant value parameters. Moreover, they are equal in the cases of winding open-circuit faults and terminal short-circuit faults. Therefore, the data storage requirements and the number of calculations to be performed by the control system are kept at a minimum.

- Remedial strategies for the operation with inverter faults have been examined as well. It has been shown that the best way of dealing with a transistor short-circuit fault in the case of a half-bridge inverter is by causing a symmetrical short circuit condition in the machine terminals.

- The validity of the proposed modified control strategies has been validated both experimentally and via FE simulations. The main benefits to gain from the modification of the control strategy are the ability to operate the drive under fault scenarios for which the drive itself, without any remedial
action, cannot operate and a substantial reduction in the torque pulsations arising from the faulted conditions.

- Except in the case of a terminal short-circuit fault between two adjacent phases, experimental results have demonstrated the adequacy of the proposed fault-tolerant operation strategies.

- Winding open-circuit faults involving the loss of two adjacent phases and terminal short-circuit faults between two non-adjacent phases require the highest current and voltage overratings in order to maintain the torque capability after fault conditions.

- Iron saturation may cause a significant reduction in the drive’s torque capability and produce a significant torque ripple under fault conditions. This effect has been checked both via FE simulations and experimental results for the designed PMSM prototype.

8.2 Future work

The present research work has presented a methodology to design and analyze fractional-slot concentrated-winding multiphase fault-tolerant permanent magnet synchronous machines. In order to complete the conducted investigation, the following research lines are suggested:

- It has been demonstrated that, for a fault-tolerant FSCW PMSM drive to maintain its pre-fault torque capability, large provisions must be generally made in terms of available DC bus voltage, power semiconductor transistor current withstand capability and machine loss increases. These aspects should be incorporated into the design process of a fault-tolerant machine. The post-fault torque requirements should be clearly specified in order to adequately choose the drive’s current and voltage levels.

- A proper machine design would require the thermal withstand capability of the machine to be checked in the whole speed range both under healthy and faulted conditions. In order to address the uneven distribution of losses under post-fault operating conditions, a thermal model that accounts for each phase separately should be adopted. A simple method is employed by the commercial software Motor-CAD [234], in which only the worst slot in terms of copper losses is modeled. The extra losses arising from the fact that it is considered that all the stator slots have the same losses are subtracted from the stator back iron. A more comprehensive analysis would require that each stator tooth and slot are independently modeled.

- The implemented PMSM drive model has been derived upon a number of simplifications; most notably, the effect of iron saturation on the machine parameters has been disregarded. It is expected that, by including iron saturation and other complex phenomena such as iron losses, parameter temperature dependency, etc. the precision of the model can be improved and additional insight about the consequences of each fault is gained. In this sense, the accuracy of the developed drive model should be compared with experimentally obtained results better than with FE simulations.

- In the post-fault operation of the machine, only the operation below base speed has been considered. Furthermore, problems with the bandwidth of the implemented current controllers have prevented experimental tests to be conducted on the design prototype operating at rated speed conditions. Further analysis and experimentation should be conducted in order to demonstrate the fault-tolerant capability of the designed PMSM drive in the whole speed range.

- In the case of terminal short-circuit faults between two adjacent phases, it has not been possible to operate the drive with the proposed modified control strategy. The reasons for this should be further investigated in the future.

- In particular, different attempts during the experimental tests have shown that the ability of the system to adequately follow the new set of references after a fault depends, to a certain extent, on when the control law is modified. More research should be done to determine the best moment to introduce the appropriate remedial actions after a fault.

- In order to grant real fault-tolerance to the designed PMSM drive, techniques for the diagnosis of the different faults ought to be implemented. This should not only be limited to the addressed
fault scenarios, but the diagnosis of other kind of drive faults, such as inter-turn short-circuit faults, current and position sensor faults, bearing damage, etc. should be implemented as well. It is expected that, in a near future, the manufactured PMSM serves as a tool for extracting fault signatures and test fault detection and diagnosis techniques.

- In order to increase the degree of fault-tolerance of the designed PMSM drive, remedial actions to allow the machine to operate under fault scenarios not contemplated in the present thesis should be investigated and implemented.
Bibliography


[110] *Analog Servo Drive 30A20AC*, ADVANCED Motion Controls, 3805 Calle Tecate, Camarillo, CA, 93012, October 2012, rev. 2.01.


[203] *Electrical steel catalogue - Non Oriented Fully Processed*, Cogent Power Ltd, PO Box 30, Newport, South Wales NP19 0XT, United Kingdom, 2002.


tolerant control of flux-switching permanent-magnet machine with redundancy,” Industrial Elec-

[288] M. Barcaro and N. Bianchi, “Air-gap flux density distortion and iron losses in anisotropic syn-


[290] G. Dajaku and D. Gerling, “Stator slotting effect on the magnetic field distribution of salient pole
3676–3683, September 2010.

chronous reluctance motor,” Energy Conversion, IEEE Transactions on, vol. 23, no. 2, pp. 466–473,
June 2008.

in synchronous reluctance and PM-assisted synchronous reluctance motors,” Industry Applications,

[293] M. Barcaro, “Design and analysis of interior permanent magnet synchronous machines for electric
vehicles,” Ph.D. dissertation, Department of Electrical Engineering, University of Padova, Padova,
Italy, 2011.

reduce cogging torque in surface-mounted permanent magnet motors,” Magnetics, IEEE Transac-

noise reduction in permanent magnet motors by teeth pairing,” Magnetics, IEEE Transactions on,


torque in surface-mounted permanent-magnet motors,” Magnetics, IEEE Transactions on, vol. 42,

permanent-magnet machines,” Magnetics, IEEE Transactions on, vol. 45, no. 4, pp. 2023–2031,
April 2009.

[299] Z. Zhu, L. Wu, and Z. Xia, “An accurate subdomain model for magnetic field computation in
slotted surface-mounted permanent-magnet machines,” Magnetics, IEEE Transactions on, vol. 46,
no. 4, pp. 1100–1115, April 2010.

[300] P. Kumar and P. Bauer, “Improved analytical model of a permanent-magnet brushless DC motor,”

[301] N. Boules, “Two-dimensional field analysis of cylindrical machines with permanent magnet excita-
1984.

[302] ———, “Prediction of no-load flux density distribution in permanent magnet machines,” Industry

[303] X. Wang, Q. Li, S. Wang, and Q. Li, “Analytical calculation of air-gap magnetic field distribution
and instantaneous characteristics of brushless DC motors,” Energy Conversion, IEEE Transactions

[304] L. Fang, S.-O. Kwon, and J.-P. Hong, “Conformal transformation technique for prediction of the
magnetic field distribution in an IPM motor,” in Electrical Machines and Systems, 2005. ICEMS


Appendix A

Multiphase Systems

Electrical multiphase systems are defined as those whose windings are constituted by one or various phases. In addition to the predominant 3-phase systems, a number of different winding arrangements are possible. Some of these possibilities are listed in Table A.1 [135]. Generally speaking, multiphase systems are divided among non-reduced, reduced and normal systems.

Non-reduced winding systems have separate windings for positive and negative magnetic axes. Since both windings produce parallel fluxes, no real multiphase system is created. In terms of magnetic axes, non-reduced winding systems are reduced to having half the number of original phases, \( m' \). By removing the windings corresponding to either the positive or negative axes, a reduced form of the winding system is obtained. If the reduction results in a radially symmetric distribution of the phases, the system is said to be a normal system.

Reduced \( m \)-phase systems have an electrical phase angle of:

\[
\alpha_m = \frac{\pi}{m} \quad \text{(A.1)}
\]

while normal \( m \)-phase systems have a phase angle of:

\[
\alpha_m = \frac{2\pi}{m} \quad \text{(A.2)}
\]

In general terms, reduced systems require a neutral line to be star connected. Without a neutral conductor, the phase windings cannot operate independently. Just consider the example of a reduced two-phase system. If no neutral connection is provided, the same current flows in the windings of both phases and a single magnetic axis is formed. The only exception are reduced systems in which the phase windings have a radially symmetric arrangement. These windings can be constructed by turning 180° the direction of the appropriate phasors in the original reduced system. The process is illustrated in Table A.1 for a 6-phase system. The result is two 3-phase systems displaced 30 electrical degrees that allow a non-loaded star connection. This construction is often termed a split-phase winding arrangement. For the transformation of a reduced system to a radially symmetrical system to be feasible, it suffices that the number of phases is not a power of 2.

Normal systems allow a star connection without neutral line, unless the number of phases is just one, \( m = 1 \).

In the present thesis, non-reduced systems are considered for machines with an even number of phases (e.g: \( m' = 4, m' = 6 \)) and normal systems for machines with an odd number of phases (e.g: \( m = 3, m = 5, m = 7 \)). In this way, a non-loaded star connection is allowed for any \( m \)-phase system and the phase angle corresponding to a normal system, (A.2), is considered.
<table>
<thead>
<tr>
<th>Number of phases</th>
<th>Non-reduced systems</th>
<th>Reduced systems</th>
<th>Normal systems</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td><img src="image1" alt="Diagram" /></td>
<td><img src="image2" alt="Diagram" /></td>
<td>--</td>
</tr>
<tr>
<td></td>
<td>$m' = 4$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td><img src="image3" alt="Diagram" /></td>
<td><img src="image4" alt="Diagram" /></td>
<td><img src="image5" alt="Diagram" /></td>
</tr>
<tr>
<td></td>
<td>$m' = 6$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td><img src="image6" alt="Diagram" /></td>
<td><img src="image7" alt="Diagram" /></td>
<td>--</td>
</tr>
<tr>
<td></td>
<td>$m' = 8$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td><img src="image8" alt="Diagram" /></td>
<td><img src="image9" alt="Diagram" /></td>
<td><img src="image10" alt="Diagram" /></td>
</tr>
<tr>
<td></td>
<td>$m' = 10$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td><img src="image11" alt="Diagram" /></td>
<td><img src="image12" alt="Diagram" /></td>
<td>--</td>
</tr>
<tr>
<td></td>
<td>$m' = 12$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td><img src="image13" alt="Diagram" /></td>
<td><img src="image14" alt="Diagram" /></td>
<td><img src="image15" alt="Diagram" /></td>
</tr>
<tr>
<td></td>
<td>$m' = 14$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>...</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table A.1: Winding systems for multiphase electrical machines (adapted from [135])
Consider the 8-pole, 5-phase and 6-phase machines examples shown in Figures A.1a and A.1b.

![8-pole machine examples](image)

(a) 5-phase  
(b) 6-phase

Figure A.1: 8-pole machine examples

The electrical angle between the phasors of two adjacent slots is $\alpha_e = p \cdot \frac{2\pi}{Q_s}$; that is, $4\pi/5$ for the 5-phase machine and $2\pi/3$ for the 6-phase machine. Therefore, the electrical displacement between the coils belonging to the $A$ and $B$ phases of the 5-phase machine is: $\alpha_m = -2\alpha_e \equiv 2\pi/5$, whereas for the 6-phase machine: $\alpha_m = 2\alpha_e + \pi \equiv \pi/3$. In both cases, the identity $\alpha_m = \frac{2\pi}{m}$ holds.
Appendix B

Reference Frames and Transformation Convention

Electrical angle convention

In the present thesis, the electrical angle for stator variables, $\theta_e$, is defined so that the first harmonic of the phase $A$ flux linkage due to the permanent magnets has a zero phase angle with respect to the sine:

$$\begin{align*}
\lambda_{A,m}(\theta_e) &= \sqrt{2}\Lambda_m \sin (1(\theta_e - 0\alpha_m)) + ...
\lambda_{B,m}(\theta_e) &= \sqrt{2}\Lambda_m \sin (1(\theta_e - 1\alpha_m)) + ...
\lambda_{C,m}(\theta_e) &= \sqrt{2}\Lambda_m \sin (1(\theta_e - 2\alpha_m)) + ...
\end{align*}$$

(B.1)

Figure B.1: Electrical angle convention: flux linkage due to permanent magnets

Reference frames for a 5-phase PMSM

Figures B.2 and B.3 show the stator, $\alpha - \beta$ and synchronous reference frame convention adopted in the present thesis. For a 5-phase machine, in addition to the synchronous reference frame rotating at fundamental frequency ($d_1 - q_1$), an additional reference frame rotating at three times the fundamental frequency can be defined ($d_3 - q_3$).
**APPENDIX B. REFERENCE FRAMES AND TRANSFORMATION CONVENTION**

**Figure B.2:** Stator reference frame for a 5-phase machine

**Figure B.3:** \(\alpha - \beta\) and synchronous reference frame at fundamental frequency for a 5-phase machine

\(\alpha - \beta\) transformation for a 5-phase PMSM

A number of different Clarke transform definitions exist for a 5-phase machine [158, 165, 431, 432]. The definition used in this thesis considers the \(\alpha_1\)-axis of the fixed stator reference frame aligned with the \(A\)-phase of the 5-phase system and the following matrix transformation between phase quantities:

\[
[T] = \frac{2}{5} \begin{bmatrix}
1 & \cos 1\alpha_m & \cos 2\alpha_m & \cos 3\alpha_m & \cos 4\alpha_m \\
0 & \sin 1\alpha_m & \sin 2\alpha_m & \sin 3\alpha_m & \sin 4\alpha_m \\
1 & \cos 3\alpha_m & \cos 6\alpha_m & \cos 9\alpha_m & \cos 12\alpha_m \\
0 & \sin 3\alpha_m & \sin 6\alpha_m & \sin 9\alpha_m & \sin 12\alpha_m \\
\frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}}
\end{bmatrix} \tag{B.2}
\]

\[
[x_{\alpha\beta}] = [T] [x_{ph}] \tag{B.3}
\]

where \([x_{ph}]\) is the vector of phase quantities (either voltages, currents or fluxes) and \([x_{\alpha\beta}]\) is the vector of those same quantities referred to the \(\alpha - \beta\) reference frame. It holds that:

\[
[T^{-1}] = \frac{5}{2} [T]^T \tag{B.4}
\]
*d – q* transformation for a 5-phase PMSM

To transform any variable from the fixed stator reference frame, \( \alpha – \beta \), to the synchronous reference frame, \( d – q \), the following Park transform is used [158]:

\[
[Q] = \begin{bmatrix}
\sin 1\theta_c & -\cos 1\theta_c & 0 & 0 & 0 \\
\cos 1\theta_c & \sin 1\theta_c & 0 & 0 & 0 \\
0 & 0 & \sin 3\theta_c & -\cos 3\theta_c & 0 \\
0 & 0 & \cos 3\theta_c & \sin 3\theta_c & 0 \\
0 & 0 & 0 & 0 & 1
\end{bmatrix}
\]

(B.5)

\[
[x_{dq}] = [Q] [x_{\alpha\beta}]
\]

(B.6)

The global transformation from the stator 5-phase system to the synchronous reference frame is given by:

\[
[x_{dq}] = [Q] [T] [x_{ph}] = [P] [x_{ph}]
\]

(B.7)

\[
[P] = \frac{2}{5} \begin{bmatrix}
sin (1 (\theta_c - 0\alpha_m)) & sin (1 (\theta_c - 1\alpha_m)) & \ldots & \sin (1 (\theta_c - 4\alpha_m)) \\
cos (1 (\theta_c - 0\alpha_m)) & \cos (1 (\theta_c - 1\alpha_m)) & \ldots & \cos (1 (\theta_c - 4\alpha_m)) \\
-\sin (3 (\theta_c - 0\alpha_m)) & \cos (3 (\theta_c - 1\alpha_m)) & \ldots & \cos (3 (\theta_c - 4\alpha_m)) \\
-\cos (3 (\theta_c - 0\alpha_m)) & -\sin (3 (\theta_c - 1\alpha_m)) & \ldots & \cos (3 (\theta_c - 4\alpha_m)) \\
\frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \ldots & \frac{1}{\sqrt{2}}
\end{bmatrix}
\]

(B.8)

The previous matrices have the following properties:

\[
[Q^{-1}] = [Q]^T
\]

(B.9)

\[
[P^{-1}] = \frac{5}{2} [P]^T
\]

(B.10)

**Application example**

The previous transformations are better understood with an example. Consider a balanced 5-phase system with a phase quantity composed of the first and third harmonics:

\[
\begin{align*}
x_A (\theta_c) &= \sqrt{2} X_1 \sin (1 (\theta_c + \phi_1 - 0\alpha_m)) + \sqrt{2} X_3 \sin (3 (\theta_c + \phi_3 - 0\alpha_m)) \\
x_B (\theta_c) &= \sqrt{2} X_1 \sin (1 (\theta_c + \phi_1 - 1\alpha_m)) + \sqrt{2} X_3 \sin (3 (\theta_c + \phi_3 - 1\alpha_m)) \\
x_C (\theta_c) &= \sqrt{2} X_1 \sin (1 (\theta_c + \phi_1 - 2\alpha_m)) + \sqrt{2} X_3 \sin (3 (\theta_c + \phi_3 - 2\alpha_m)) \\
x_D (\theta_c) &= \sqrt{2} X_1 \sin (1 (\theta_c + \phi_1 - 3\alpha_m)) + \sqrt{2} X_3 \sin (3 (\theta_c + \phi_3 - 3\alpha_m)) \\
x_E (\theta_c) &= \sqrt{2} X_1 \sin (1 (\theta_c + \phi_1 - 4\alpha_m)) + \sqrt{2} X_3 \sin (3 (\theta_c + \phi_3 - 4\alpha_m))
\end{align*}
\]

(B.11)

where \( X_1 \) and \( X_3 \) are the first and third harmonic RMS magnitudes of the considered variable. The quantities referred to the \( \alpha – \beta \) reference frame are:

\[
\begin{align*}
x_{\alpha_1} (\theta_c) &= \sqrt{2} X_1 \sin (1 (\theta_c + \phi_1)) \\
x_{\beta_1} (\theta_c) &= -\sqrt{2} X_1 \cos (1 (\theta_c + \phi_1)) \\
x_{\alpha_3} (\theta_c) &= \sqrt{2} X_3 \sin (3 (\theta_c + \phi_3)) \\
x_{\beta_3} (\theta_c) &= -\sqrt{2} X_3 \cos (3 (\theta_c + \phi_3)) \\
x_0 &= 0
\end{align*}
\]

(B.12)
while referred to the $d-q$ synchronous reference frame:

\[
\begin{align*}
    x_d &= \sqrt{2} X_1 \cos (\phi_1) \\
    x_q &= \sqrt{2} X_1 \sin (\phi_1) \\
    x_d' &= \sqrt{2} X_3 \cos (3\phi_3) \\
    x_q' &= \sqrt{2} X_3 \sin (3\phi_3) \\
    x_0 &= 0
\end{align*}
\] (B.13)

Hence, phase quantities varying at a certain frequency have a constant value when referred to the synchronous reference frame rotating at that same frequency. First harmonic components are reflected in axes $d_1 - q_1$, while third harmonic components are reflected in axes $d_3 - q_3$. Any other harmonic component will be reflected as an electrical angle varying signal in both $d_1 - q_1$ and $d_3 - q_3$ axes.
Appendix C

Analytical Inductance Calculation: Slot-Component

In the following appendix, analytical expressions for the calculation of the slot leakage inductances for any machine with a 1-layer, 2-layer or 4-layer winding are obtained. The formulas are derived from solving the 2-D Poisson problem associated with the slot region, making them more accurate than classically used expressions that assume a leakage flux path parallel to the placement of the conductors in the slot. First, the magnetic 2-D field distribution in the slot due to the armature excitation is obtained. Then, analytical expressions for the magnetic energy in the slot and the slot region are derived. From these expressions, the related components of the self and mutual inductances are deduced.

C.1 Magnetic field: slot region

The solution to the armature magnetic field distribution in the slot region is obtained by considering the Coulomb gauge and by solving the magneto-static Ampère-Maxwell equation in terms of the magnetic vector potential in the slot, $\vec{A}_{\text{slot}}$:

$$\nabla^2 \vec{A}_{\text{slot}} = -\mu_0 \vec{j}_{\text{slot}}$$  \hspace{1cm} (C.1)

where $\mu_0$ is the permeability of free space [H/m] and $\vec{j}_{\text{slot}}$ is the slot current density [A/m$^2$]. In the case of a radial flux machine, the variation of the magnetic field distribution in the axial direction is usually ignored and the problem is reduced to a two-dimensional case:

$$\vec{A}_{\text{slot}} = A_{\text{slot},z} \cdot \vec{u}_z = A_{\text{slot}}(x,y) \cdot \vec{u}_z$$  \hspace{1cm} (C.2)

By dividing the slot into two regions, Up (U) and Down (D):

$$\begin{cases} \nabla^2 A_{\text{slot},U}(x,y) = -\mu_0 j_{\text{slot},U}(x) \\ \nabla^2 A_{\text{slot},D}(x,y) = -\mu_0 j_{\text{slot},D}(x) \end{cases}$$  \hspace{1cm} (C.3)

the current density in the slot can be expressed as:

$$\begin{cases} j_{\text{slot},U}(x) = \frac{2}{bh} (i_{\text{slot},UR} + i_{\text{slot},UL}) + \sum_{n=1,3,5,...}^{\infty} \frac{8}{bh\pi n} (i_{\text{slot},UR} - i_{\text{slot},UL}) \sin \left( \frac{\pi n x}{b} \right) \\ j_{\text{slot},D}(x) = \frac{2}{bh} (i_{\text{slot},DR} + i_{\text{slot},DL}) + \sum_{n=1,3,5,...}^{\infty} \frac{8}{bh\pi n} (i_{\text{slot},DR} - i_{\text{slot},DL}) \sin \left( \frac{\pi n x}{b} \right) \end{cases}$$  \hspace{1cm} (C.4)
APPENDIX C. ANALYTICAL INDUCTANCE CALCULATION: SLOT-COMPONENT

where \( i_{\text{slot,UL}}, i_{\text{slot,UR}}, i_{\text{slot,DL}} \) and \( i_{\text{slot,DR}} \) stand for the total currents in the Up-Left (UL), Up-Right (UR), Down-Left (DL) and Down-Right (DR) sides of the slot, respectively. Equations in (C.4) correspond to the Fourier series expansion of the slot current density in the (U) and (D) regions of the slot, respectively. They are derived upon the assumption that the circulating current in each of the 4 slot subdivisions (UL, UR, DL and DR) is homogeneous.

The solving process is based on [195]. The following simplifications are made in the analysis:

1. Infinite permeability of the stator material.
2. The slot has a rectangular geometry, described by the parameters shown in Figure C.1.
3. End effects in the \( z \) component are ignored.
4. In the slot-opening region, the \( x \) component of the magnetic field intensity is constant. The smaller the slot-opening, the more valid this assumption is.

Applying Ampère’s law to the slot contour yields:

\[
\oint_{C} \vec{H}_{\text{slot}} \, d\vec{l} = H_{\text{slot},x}\left(|x| \leq \frac{b_0}{2}, y = 0\right) b_0 = i_{\text{slot}}
\]  

\[
H_{\text{slot},x}\left(|x| \leq \frac{b_0}{2}, y = 0\right) = \frac{i_{\text{slot}}}{b_0}
\]

\[
H_{\text{slot},x}(x, y = 0) = h(x) = \frac{i_{\text{slot}}}{b} + \sum_{n=2,4,6,\ldots}^{\infty} \frac{4i_{\text{slot}}}{b_0 \pi n} \sin \left(\frac{b_0 \pi n}{2b}\right) \cos \left(\frac{\pi n}{b} x\right)
\]

where \( i_{\text{slot}} \) is the total current in the slot, \( i_{\text{slot}} = i_{\text{slot,UL}} + i_{\text{slot,UR}} + i_{\text{slot,DL}} + i_{\text{slot,DR}} \).

Figure C.1: Simplified slot geometry and nomenclature
The problem’s boundary conditions are:

\[
\begin{align*}
\frac{\partial A_{\text{slot},D}}{\partial x} (x, y = h) &= 0 \\
\frac{\partial A_{\text{slot},D}}{\partial y} (x, y = h) &= 0 \\
\frac{\partial A_{\text{slot},D}}{\partial x} (x = b/2, y) &= 0 \\
\frac{\partial A_{\text{slot},D}}{\partial y} (x = b/2, y) &= 0 \\
\frac{\partial A_{\text{slot},U}}{\partial x} (x, y = 0) &= \mu_0 h (x) \\
\frac{\partial A_{\text{slot},D}}{\partial x} (x = -b/2, y) &= 0 \\
\frac{\partial A_{\text{slot},D}}{\partial y} (x = -b/2, y) &= 0 \\
\frac{\partial A_{\text{slot},U}}{\partial y} (x, y = h/2) &= \frac{\partial A_{\text{slot},D}}{\partial y} (x, y = h/2) \\
\frac{\partial A_{\text{slot},U}}{\partial x} (x, y = h/2) &= \frac{\partial A_{\text{slot},D}}{\partial x} (x, y = h/2) \\
A_{\text{slot},U} (x, y = h/2) &= A_{\text{slot},D} (x, y = h/2)
\end{align*}
\]  
(C.8)

By applying the boundary conditions stated in (C.8) to equation (C.1), the problem becomes determined and the solution, expressed in terms of the magnetic field strength in the slot, is:

\[
\begin{align*}
H_{\text{slot},U,x} (x, y) &= \frac{2}{bh} i_{\text{slot},U} (h - y) - \sum_{n=1,3,5,...}^{\infty} \frac{4}{b^2 h} (\Delta i_{\text{slot},U} - \Delta i_{\text{slot},D}) \left( b \pi n \right)^2 \sin \left( \frac{\pi n}{b} x \right) \left( \sinh \left( \frac{\pi n}{b} y \right) \right) \\
&\quad - \sum_{n=2,4,6,...}^{\infty} \frac{4}{bb_0} i_{\text{slot}} \left( \frac{b}{\pi n} \right) \sin \left( \frac{b_0 \pi n}{2b} \right) \cos \left( \frac{\pi n}{b} x \right) \left( \sinh \left( \frac{\pi n}{b} y \right) \right)
\end{align*}
\]  
(C.9)

\[
\begin{align*}
H_{\text{slot},U,y} (x, y) &= -\sum_{n=1,3,5,...}^{\infty} \frac{8}{b^2 h} \Delta i_{\text{slot},U} \left( b \pi n \right)^2 \cos \left( \frac{\pi n}{b} x \right) + \sum_{n=1,3,5,...}^{\infty} \frac{4}{b^2 h} (\Delta i_{\text{slot},U} - \Delta i_{\text{slot},D}) \left( b \pi n \right)^2 \\
&\quad \cos \left( \frac{\pi n}{b} x \right) \left( \cosh \left( \frac{\pi n}{b} y \right) \right) - \sum_{n=2,4,6,...}^{\infty} \frac{4}{bb_0} i_{\text{slot}} \left( \frac{b}{\pi n} \right) \sin \left( \frac{b_0 \pi n}{2b} \right) \sin \left( \frac{\pi n}{b} x \right) \\
&\quad \sinh \left( \frac{\pi n}{b} y \right) \left( \cosh \left( \frac{\pi n}{b} y \right) \right)
\end{align*}
\]  
(C.10)

\[
\begin{align*}
H_{\text{slot},D,x} (x, y) &= \frac{2}{bh} i_{\text{slot},D} y + \frac{1}{b} i_{\text{slot}} + \sum_{n=1,3,5,...}^{\infty} \frac{4}{b^2 h} (\Delta i_{\text{slot},U} - \Delta i_{\text{slot},D}) \left( b \pi n \right)^2 \sin \left( \frac{\pi n}{b} x \right) \left( \sinh \left( \frac{\pi n}{b} y \right) \right) \\
&\quad - \sum_{n=2,4,6,...}^{\infty} \frac{4}{bb_0} i_{\text{slot}} \left( \frac{b}{\pi n} \right) \sin \left( \frac{b_0 \pi n}{2b} \right) \cos \left( \frac{\pi n}{b} x \right) \left( \cosh \left( \frac{\pi n}{b} y \right) \right)
\end{align*}
\]  
(C.11)

\[
\begin{align*}
H_{\text{slot},D,y} (x, y) &= -\sum_{n=1,3,5,...}^{\infty} \frac{8}{b^2 h} \Delta i_{\text{slot},D} \left( b \pi n \right)^2 \cos \left( \frac{\pi n}{b} x \right) - \sum_{n=1,3,5,...}^{\infty} \frac{4}{b^2 h} (\Delta i_{\text{slot},U} - \Delta i_{\text{slot},D}) \left( b \pi n \right)^2 \\
&\quad \cos \left( \frac{\pi n}{b} x \right) \left( \cosh \left( \frac{\pi n}{b} y \right) \right) - \sum_{n=2,4,6,...}^{\infty} \frac{4}{bb_0} i_{\text{slot}} \left( \frac{b}{\pi n} \right) \sin \left( \frac{b_0 \pi n}{2b} \right) \sin \left( \frac{\pi n}{b} x \right) \left( \cosh \left( \frac{\pi n}{b} y \right) \right)
\end{align*}
\]  
(C.12)
APPENDIX C. ANALYTICAL INDUCTANCE CALCULATION: SLOT-COMPONENT

where:

\[
\begin{align*}
    i_{\text{slot},U} &= i_{\text{slot},UR} + i_{\text{slot},UL} \\
    i_{\text{slot},D} &= i_{\text{slot},DR} + i_{\text{slot},DL} \\
    \Delta i_{\text{slot},U} &= i_{\text{slot},UR} - i_{\text{slot},UL} \\
    \Delta i_{\text{slot},D} &= i_{\text{slot},DR} - i_{\text{slot},DL}
\end{align*}
\]  

(C.13)

C.2 Magnetic field: slot-opening region

In the slot-opening region, the current density is null and finding the magnetic field distribution is equal to solving the associated Laplace equation:

\[
\nabla^2 A_{\text{slot-opening}} = 0
\]  

(C.14)

Under the assumption that the \(x\) component of the magnetic field intensity in the considered region is constant, the following expressions are derived:

\[
H_{\text{slot-opening},x}(x, y) = \frac{i_{\text{slot}}}{b_0}
\]  

(C.15)

\[
H_{\text{slot-opening},y}(x, y) = 0
\]  

(C.16)

C.3 Magnetic energy

C.3.1 Magnetic energy stored in a single slot

Let \(w\) be the magnetic energy density in the slot and slot-opening regions:

\[
w = \frac{1}{2} \vec{B} \cdot \vec{H} = \frac{1}{2} \mu_0 |\vec{H}|^2 = \frac{1}{2} \mu_0 \left( H_x^2 + H_y^2 \right)
\]  

(C.17)

The magnetic energy stored in a slot, including the slot-opening region, as a function of the currents in the same is:

\[
W_{\text{slot}} = \iiint_V w dV = \frac{1}{2} \mu_0 l_{ef} \left( \int_{x=\frac{b}{2}}^{b} \int_{y=-\frac{h}{2}}^{\frac{h}{2}} \left( H_{\text{slot},U,x}^2 + H_{\text{slot},U,y}^2 \right) dx dy + \int_{x=\frac{b}{2}}^{b} \int_{y=0}^{\frac{h}{2}} \left( H_{\text{slot},D,x}^2 + H_{\text{slot},D,y}^2 \right) dx dy \\
+ \int_{x=-\frac{b}{2}}^{\frac{b}{2}} \int_{y=-h_0}^{0} \left( H_{\text{slot-opening},x}^2 + H_{\text{slot-opening},y}^2 \right) dx dy \right)
\]  

(C.18)

where \(l_{ef}\) represents the effective length of the machine. After some calculations, the above expression turns into:

\[
W_{\text{slot}} = \frac{1}{2} \mu_0 l_{ef} (w_1 + w_2 + w_3 + w_4)
\]  

(C.19)

where:
C.3. Magnetic Energy

\[
\begin{align*}
    w_1 &= (i^2_{\text{slot,D}} + 3i_{\text{slot}}i_{\text{slot,U}} + i^2_{\text{slot,U}}) \frac{h}{6b} \\
    w_2 &= (\Delta i^2_{\text{slot,U}} + \Delta i^2_{\text{slot,D}}) \frac{b}{6h} \\
    w_3 &= -(\Delta i_{\text{slot,U}} - \Delta i_{\text{slot,D}})^2 \sum_{n=1,3,5,...}^{\infty} \frac{16b^2}{h^2} \frac{1}{(\pi n)^2} \tanh \left( \frac{h\pi n}{2b} \right) \\
    w_4 &= i_{\text{slot}}^2 \left( \frac{h_0}{b_0} + \sum_{n=2,4,6,...}^{\infty} \frac{8b^2}{b_0^2} \frac{1}{(\pi n)^3} \sin^2 \left( \frac{h\pi n}{2b} \right) \tanh \left( \frac{h\pi n}{2b} \right) \right)
\end{align*}
\]  

(C.20)

C.3.2 Magnetic energy: slot region

Assuming that all the slots have the same geometry, the magnetic energy stored in the slot region of the machine is:

\[
W_{\text{slot-region}} = \sum_{j=1}^{Q} W_{\text{slot,j}}
\]

(C.21)

where \(Q\) is the number of slots in the machine. The currents in a slot \(j\), as a function of the phase currents, can be expressed as:

\[
i_{\text{slot,x,j}} = \frac{N_s m^2}{2Q} [k_{x,j}] [i_{\text{ph}}] \quad x \in \{UL, UR, DL, DR\}
\]

(C.22)

where \(m\) is the phase number and \(N_s\) the number of turns wound in series per phase. Vector \([i_{\text{ph}}]_{(m \times 1)}\) represents the phase currents and vector \([k_{x,j}]_{(1 \times m)}\) is constituted by sign type variables (-1, 0 or 1) that describe which phase layer \(x\) of slot \(j\) belongs to.

\[
[i_{\text{ph}}] = \begin{bmatrix} i_A \\ i_B \\ \vdots \\ i_m \end{bmatrix}
\]

(C.23)

Introducing expressions (C.19),(C.20) and (C.22) in (C.21) and expanding, yields:

\[
W_{\text{slot-region}} = \mu_0 l_e \frac{N^2_s}{p} [i_{\text{ph}}]^T [\lambda_{\text{slot}}] [i_{\text{ph}}]
\]

(C.24)

where \(p\) is the number of pole-pairs and \([\lambda_{\text{slot}}]_{(m \times m)}\) the permeance coefficient matrix associated to the slot region:

\[
[\lambda_{\text{slot}}] = [\beta] \frac{h}{3b} + [\gamma] \frac{b}{12h} - [\delta] \frac{16b^2}{h^2} \sum_{n=1,3,5,...}^{\infty} \frac{\tanh \left( \frac{h\pi n}{2b} \right)}{(\pi n)^5} + [c] \left( \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,...}^{\infty} \frac{\sin^2 \left( \frac{h\pi n}{2b} \right)}{(\pi n)^3} \tanh \left( \frac{h\pi n}{2b} \right) \right)
\]

(C.25)

and:

\[
[\beta] = \frac{m}{32qQ} \sum_{j=1}^{Q} \left( [k_{DR,j}] + [k_{DL,j}] \right)^T \left( [k_{DR,j}] + [k_{DL,j}] \right) + 3 \left( [k_{UR,j}] + [k_{UL,j}] + [k_{DR,j}] + [k_{DL,j}] \right)^T \left( [k_{UR,j}] + [k_{UL,j}] \right)
\]

(C.26)
\[ [\gamma] = \frac{m}{8qQ} \sum_{j=1}^{Q} \left[ (|k_{UR,j}| - |k_{UL,j}|)^T (|k_{UR,j}| - |k_{UL,j}|) + (|k_{DR,j}| - |k_{DL,j}|)^T (|k_{DR,j}| - |k_{DL,j}|) \right] \] (C.27)

\[ [\delta] = \frac{m}{16qQ} \sum_{j=1}^{Q} \left[ (|k_{UR,j}| - |k_{UL,j}| - |k_{DR,j}| + |k_{DL,j}|)^T (|k_{UR,j}| - |k_{UL,j}| - |k_{DR,j}| + |k_{DL,j}|) \right] \] (C.28)

\[ [\epsilon] = \frac{m}{16qQ} \sum_{j=1}^{Q} \left[ (|k_{UR,j}| + |k_{UL,j}| + |k_{DR,j}| + |k_{DL,j}|)^T (|k_{UR,j}| + |k_{UL,j}| + |k_{DR,j}| + |k_{DL,j}|) \right] \] (C.29)

where \( q \) is the number of slots per pole and phase. Variables \( |k_{UL,j}|, |k_{UR,j}|, |k_{DL,j}| \) and \( |k_{DR,j}| \) stand for the sign type vectors introduced in equation (C.22) for the Up-Left (UL), Up-Right (UR), Down-Left (DL) and Down-Right (DR) zones of the slot, respectively.

### C.4 Inductance calculation

The slot components for the self-inductance and the mutual inductance between phases can be obtained by their relation to the magnetic energy. If:

\[
\begin{align*}
[i_{ph,AB}] &= [i_A \ i_B \ldots \ 0]^T \\
[i_{ph,A}] &= [i_A \ \ 0 \ldots \ 0]^T \\
[i_{ph,B}] &= [0 \ i_B \ldots \ 0]^T
\end{align*}
\] (C.30)

then:

\[
\begin{align*}
W_{\text{slot-region,AB}} &= \mu_0 \ell_f N_s^2 [i_{ph,AB}]^T [\lambda_{\text{slot}}][i_{ph,AB}] \\
W_{\text{slot-region,A}} &= \mu_0 \ell_f N_s^2 [i_{ph,A}]^T [\lambda_{\text{slot}}][i_{ph,A}] \\
W_{\text{slot-region,B}} &= \mu_0 \ell_f N_s^2 [i_{ph,B}]^T [\lambda_{\text{slot}}][i_{ph,B}]
\end{align*}
\] (C.31)

and the \( A \)-phase self-inductance, \( L_{\text{slot,AA}} \), is obtained as:

\[
L_{\text{slot,AA}} = \frac{2W_{\text{slot-region,A}}}{i_A^2} = 2\mu_0 \ell_f N_s^2 \lambda_{\text{slot,AA}} = \frac{2\mu_0 \ell_f N_s^2}{p} \lambda_{\text{slot,AA}}
\] (C.32)

while the mutual inductance between phases \( A \) and \( B \), \( L_{\text{slot,AB}} \), in turn, is obtained as:

\[
L_{\text{slot,AB}} = \frac{W_{\text{slot-region,AB}} - W_{\text{slot-region,A}} - W_{\text{slot-region,B}}}{i_{AB}} = 2\mu_0 \ell_f N_s^2 \left( \frac{[\lambda_{\text{slot,AB}}] + [\lambda_{\text{slot,BA}}]}{2} \right)
\]

\[
= -2\mu_0 \ell_f \frac{N_s^2}{p} \lambda_{\text{slot,AB}}
\] (C.33)

where:
\[
\lambda_{\text{slot,AA}} = \beta_{AA} \frac{h}{3b} + \gamma_{AA} \frac{b}{12h} - \delta_{AA} \frac{16b^2}{h^2} \sum_{n=1,3,5,\ldots}^{\infty} \frac{\tanh \left( \frac{h\pi n}{2b} \right)}{(\pi n)^5} + \epsilon_{AA} \left( \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,\ldots}^{\infty} \frac{\sin^2 \left( \frac{h\pi n}{2b} \right)}{(\pi n)^3 \tanh \left( \frac{h\pi n}{2b} \right)} \right) \tag{C.34}
\]

\[
\lambda_{\text{slot,AB}} = \beta_{AB} \frac{h}{3b} + \gamma_{AB} \frac{b}{12h} - \delta_{AB} \frac{16b^2}{h^2} \sum_{n=1,3,5,\ldots}^{\infty} \frac{\tanh \left( \frac{h\pi n}{2b} \right)}{(\pi n)^5} + \epsilon_{AB} \left( \frac{h_0}{b_0} + \frac{8b^2}{b_0^2} \sum_{n=2,4,6,\ldots}^{\infty} \frac{\sin^2 \left( \frac{h\pi n}{2b} \right)}{(\pi n)^3 \tanh \left( \frac{h\pi n}{2b} \right)} \right) \tag{C.35}
\]

Coefficients \(\beta_{ii}, \gamma_{ii}, \delta_{ii}, \epsilon_{ii}\) (C.34), \(\beta_{ij}, \gamma_{ij}, \delta_{ij}\) and \(\epsilon_{ij}\) (C.35) are dimensionless parameters that depend solely on the winding arrangement.
Appendix D

Analytical Inductance Calculation: Air-Gap-Component

The process for obtaining analytical expressions for the air-gap inductances in a radial flux machine is analogous to the one followed in Appendix C. Again, the formulas are derived from solving a 2-D Poisson problem. Therefore, the derived expressions are more accurate than traditionally used formulas based on a 1-D analysis. In the description of the problem, cylindrical coordinates are considered.

D.1 Magnetic field: air-gap region

In the air-gap region of a machine there is no current circulation and obtaining the magnetic field distribution is equivalent to solving the associated Laplace problem:

\[ \nabla^2 \vec{A}_{agap} = 0 \quad (D.1) \]

where \( \vec{A}_{agap} \) is the vector potential in the air-gap. In solving the problem, the following simplifications are considered:

1. Rotor and stator regions have infinite permeability.
2. A smooth, uniform air-gap as the one depicted in Figure D.1 is considered (magnets are ignored).
3. End-effects in the \( z \) component are ignored.
4. In the slot-opening region, the \( \theta \) component of the magnetic field intensity is constant and its value is:

\[ H_{agap,\theta} (r = R_s, \theta) = -\frac{i_{slot}}{R_s \alpha_0} \quad (D.2) \]

The problem’s boundary conditions are:

\[
\begin{cases}
A_{agap} (r, \theta) = A_{agap} (r, \theta + 2\pi) \\
H_{agap,\theta} (r = R_s, \theta) = f (\theta) \\
H_{agap,\theta} (r = R_r, \theta) = 0
\end{cases} \quad (D.3)
\]

where function \( f (\theta) \) is determined by the sum of the currents in each slot:

\[ f (\theta) = \sum_{n=1,2,3\ldots}^{\infty} f_{cn} \cos(n\theta) + f_{sn} \sin(n\theta) \quad (D.4) \]
APPENDIX D. ANALYTICAL INDUCTANCE CALCULATION: AIR-GAP-COMPONENT

\[ f_{cn} = -\frac{N_s m}{2\pi R_s Q} \sum_{j=1}^{Q} \left( [k_{UR,j} + k_{U L,j} + k_{D R,j} + k_{D L,j}] [i_{ph}] \cos \left( n_{\alpha_{slot}} \left( j - \frac{1}{2} \right) \right) \right) \]  \hspace{1cm} (D.5)

\[ f_{sn} = -\frac{N_s m}{2\pi R_s Q} \sum_{j=1}^{Q} \left( [k_{UR,j} + k_{U L,j} + k_{D R,j} + k_{D L,j}] [i_{ph}] \sin \left( n_{\alpha_{slot}} \left( j - \frac{1}{2} \right) \right) \right) \]  \hspace{1cm} (D.6)

where \( \alpha_{slot} \) is the slot angle, \( \alpha_{slot} = \frac{2\pi}{Q} \), and \( \text{sinc}(x) = \frac{\sin(x)}{x} \). The solution to the problem is:

\[ A_{agap}(r, \theta) = -\sum_{n=1,2,3\ldots}^{\infty} \frac{\mu_0 R_s}{n} \left( \frac{r}{R_s/R_r} \right)^n + \left( \frac{R_r}{r} \right)^n \left( f_{cn} \cos(n\theta) + f_{sn} \sin(n\theta) \right) \]  \hspace{1cm} (D.7)

\[ H_{agap,r}(r, \theta) = \sum_{n=1,2,3\ldots}^{\infty} \frac{R_s}{r} \left( \frac{r}{R_s/R_r} \right)^n + \left( \frac{R_r}{r} \right)^n \left( f_{cn} \sin(n\theta) - f_{sn} \cos(n\theta) \right) \]  \hspace{1cm} (D.8)

\[ H_{agap,\theta}(r, \theta) = \sum_{n=1,2,3\ldots}^{\infty} \frac{R_s}{r} \left( \frac{r}{R_s/R_r} \right)^n - \left( \frac{R_r}{r} \right)^n \left( f_{cn} \cos(n\theta) + f_{sn} \sin(n\theta) \right) \]  \hspace{1cm} (D.9)

D.2 Magnetic energy: air-gap region

The magnetic energy stored in the air-gap region is:

\[ W_{agap} = \frac{1}{2} \mu_0 l_{ef} \int_{r=R_r}^{R_s} \int_{\theta=0}^{2\pi} \left( H_{agap,r}^2 + H_{agap,\theta}^2 \right) rdrd\theta \]  \hspace{1cm} (D.10)

\[ = \frac{1}{2} \mu_0 l_{ef} \pi R_s^2 \sum_{n=1,2,3\ldots}^{\infty} \frac{1}{n} \left( f_{cn}^2 + f_{sn}^2 \right) \left( \frac{R_s}{R_r} \right)^n + \left( \frac{R_r}{R_s} \right)^n \]
D.3 Inductance calculation

The air-gap component of the self-inductance of phase A, \( L_{\text{agap,AA}} \), is:

\[
L_{\text{agap,AA}} = \frac{2W_{\text{agap,A}}}{i_A^2} = 2\mu_0 l_{ef} \frac{N_s^2}{p} [\lambda_{\text{agap,AA}}] = 2\mu_0 l_{ef} \frac{N_s^2}{p} \lambda_{\text{agap,AA}}
\]  
(D.11)

whereas the air-gap component of the mutual inductance between phases A and B, \( L_{\text{agap,AB}} \):

\[
L_{\text{agap,AB}} = \frac{W_{\text{agap,AB}} - W_{\text{agap,A}} - W_{\text{agap,B}}}{i_A i_B} = 2\mu_0 l_{ef} \frac{N_s^2}{p} \left( [\lambda_{\text{agap,AB}}] + [\lambda_{\text{agap,BA}}] \right) = -2\mu_0 l_{ef} \frac{N_s^2}{p} \lambda_{\text{agap,AB}}
\]  
(D.12)

where \( \lambda_{\text{agap,AA}} \) and \( \lambda_{\text{agap,AB}} \) are the air-gap permeance coefficients of the self and mutual inductances, respectively:

\[
[\lambda_{\text{agap}}] = \sum_{n=1,2,3...}^{\infty} \frac{1}{\pi n} \text{sinc}^2 \left( \frac{n\alpha_0}{2} \right) \left( [\epsilon] + [\zeta_n] \right) \frac{(R_s/R_r)^n + (R_r/R_s)^n}{(R_s/R_r)^n - (R_r/R_s)^n}
\]  
(D.13)

\[
[\zeta_n] = \frac{m}{16qQ} \sum_{j,l=1}^{Q} \left( [k_{UR,j}] + [k_{UL,j}] + [k_{DR,j}] + [k_{DL,j}] \right)^T ([k_{UR,l}] + [k_{UL,l}] + [k_{DR,l}] + [k_{DL,l}]) \cos (n\alpha_{\text{slot}}(j - l))
\]  
(D.14)

Matrix \([\epsilon]\) is given by equation (C.29). Elements in matrix \([\zeta_n]\) are dimensionless parameters that depend solely on the winding topology and the subscript index, \(n\). These parameters are periodic in the variable \(n\), where the period is equal to the number of slots, \(Q\), or fractions of said number.
APPENDIX D. ANALYTICAL INDUCTANCE CALCULATION: AIR-GAP-COMPONENT
Appendix E

Winding Diagram for the Prototype

Figure E.1: Winding diagram for the prototype \( m = 5, p = 11, Q = 20, n_l = 1 \)
APPENDIX E. WINDING DIAGRAM FOR THE PROTOTYPE
Appendix F

Copper Loss Minimization under Fault Conditions

A general procedure to modify the control strategy of an AC machine subject to both open-circuit and short-circuit faults is presented in the following appendix. The method seeks to select the current references that minimize the stator copper losses while maintaining the main harmonic of the armature MMF. First, the generalities of the method are described. Then, the particular equations depending on whether the fault involves a single phase or two phases, or whether the zero sequence current can be controlled or not, are derived. Finally, specific examples in the case of a 5-phase machine are shown.

F.1 General description of the method

The main purpose of modifying the control strategy in an AC drive subject to faults is that, under fault conditions, the pre-fault current references cannot be adequately followed, leading to parasitic phenomena such as: a decrease in the torque capability, an increased torque ripple, increase losses, etc. By ensuring that the main harmonic of the armature MMF retains its pre-fault waveform, the mean torque capability of the drive can be maintained. Furthermore, if the higher order components of the torque producing currents are negligible, as they usually are in machines designed to induce a sinusoidal back-EMF, the previous condition also ensures a minimum torque ripple. Translating it into equations, maintaining the first harmonic of the armature MMF with respect to the pre-fault state is equal to preserving the instantaneous values of currents $i_{\alpha_1}$ and $i_{\beta_1}$.

In a multiphase AC machine, the additional degrees of freedom stemming from the increased number of phases over a traditional 3-phase machine can be used to ensure the above condition even under different faults. Furthermore, depending on the number of faulted phases, some liberty to select the current references besides $i_{\alpha_1}$ and $i_{\beta_1}$ may remain. This freedom can be then exploited to minimize the machine losses. In the following, the minimization of the stator copper losses is considered.

The instantaneous copper losses in an AC machine can be expressed both as a function of the phase currents and the currents referred to the stationary $\alpha - \beta$ reference frame:

$$P_{Cu}(t) = \sum_{l=1}^{m} R_{ph} |i_{ph l}|^2 = \sum_{l=1}^{m} R_{ph} |i_{\alpha\beta l}|^2$$

(F.1)

where $[i_{ph l}] = [i_A \ i_B \ ... \ i_m]^T$ and $[i_{\alpha\beta l}] = [i_{\alpha_1} \ i_{\beta_1} \ ... \ i_0]^T$. By operating in the $\alpha - \beta$ reference frame, minimizing the stator copper losses while maintaining the main harmonic of the armature MMF is equivalent to finding a solution to the following minimization problem:
\[ \min f = \sum_{l=1}^{m} [i_{\alpha\beta}]_l^2, \quad \text{subject to } \begin{cases} \hat{i}_{\alpha_1} = \chi \cdot i_{\alpha_1, h} \\ \hat{i}_{\beta_1} = \chi \cdot i_{\beta_1, h} \end{cases} \] (F.2)

where \( i_{\alpha_1, h} \) and \( i_{\beta_1, h} \) are the healthy (pre-fault) state values of the the first harmonic current components in the \( \alpha - \beta \) reference frame. Parameter \( \chi \) is an optional derating factor that can be introduced in order to ensure a number of goals with the modified post-fault control strategy. If the objective is that the main harmonic torque component retains its pre-fault value, then the derating factor must be set to unity (\( \chi = 1 \)). On the other hand, if it is sought that the drive maintains its pre-fault losses or current levels, a lower factor must be selected (\( \chi < 1 \)).

Depending whether the zero sequence current in the machine can be controlled, an additional constraint for the minimization problem must be added. For example, in case the machine is supplied by a full H-bridge inverter, each phase can be independently controlled and the zero sequence current remains a degree of freedom that can be used to minimize machine losses. On the contrary, if the machine is star connected and no path for the zero sequence current to flow is provided, a supplementary constraint must be added to reflect this fact. Additional constraints can be imposed as well in order to cancel specific higher order current components. Finally, a number of constraints must be imposed in order to consider the different fault scenarios.

Once the minimization problem and all the imposed constraints have been stated, the easiest way of finding the optimal solution is by employing the method of Lagrange multipliers. By introducing an additional parameter for each constraint, the previous minimization problem is equivalent to finding the stationary points of a function of the form:

\[ F = \sum_{l=1}^{m} [i_{\alpha\beta}]_l^2 + \lambda_{\alpha_1} (i_{\alpha_1} - \chi \cdot i_{\alpha_1, h}) + \lambda_{\beta_1} (i_{\beta_1} - \chi \cdot i_{\beta_1, h}) + \ldots \] (F.3)

where \( \lambda_{\alpha_1}, \lambda_{\beta_1}, \text{etc.} \) are the Lagrange multipliers associated to each of the imposed constraints.

The proposed methodology, based on the work by [414], is general and valid for both induction and synchronous machines with any phase number. The only requirement of the method is that the Clarke transformation for the multiphase system is defined so that a zero sequence term results from the transformation of the phase variables to the variables expressed in the \( \alpha - \beta \) reference frame. The method described in the previous reference, proposed for a 6-phase induction machine subject to a single phase winding open-circuit fault and supplied by a full H-bridge inverter, has been extended to account for both open-circuit and short-circuit faults, different supply conditions and the possibility of having more than one faulted phase at a time. The solution to the previous minimization problem under different conditions is derived in the following sections.

\section*{F.2 Single phase fault}

When a phase is at fault, the current in such a phase cannot be controlled by modifying the machine supply, but it is determined by the particular fault conditions, the speed and the machine parameters. For instance, an open-circuit fault forces the current in the faulted phase to be zero, while if the fault is a short-circuit one, the current in the faulted phase varies sinusoidally with time. In general terms, the current in a faulted phase \( k \) has a time expression \([i_{\alpha\beta}]_k = i_k (t)\) that cannot be altered by modifying the current control references.

Two distinct cases are considered for the fault of a single phase, depending on whether the zero sequence current can be controlled or not. As previously commented, in the first case, the zero sequence current is a degree of freedom that can be varied in order to reduce the stator copper losses, while in the second case it is forced to follow a certain time expression, much in the same fashion the faulted phases do: \([i_{\alpha\beta}]_m = i_0 (t)\).
F.2. SINGLE PHASE FAULT

F.2.1 With control over the zero sequence current

Let \( k \) be the faulted phase for which the controlability of the phase current has been lost. If the faulted phase current has a time expression of the form \([i_{ph}]_k = i_k(t)\), by applying the reverse Clarke transformation (see Appendix B), the previous fault condition can be expressed as:

\[
[i_{ph}]_k = \sum_{l=1}^{m} [T^{-1}]_{kl} [i_{\alpha\beta}]_l = i_k(t)
\] (F.4)

Hence, the function \( F \) to minimize becomes:

\[
F = \sum_{l=1}^{m} [i_{\alpha\beta}]_l^2 + \lambda_\alpha (i_{\alpha_1} - \chi \cdot i_{\alpha_1,h}) + \lambda_{\beta_1} (i_{\beta_1} - \chi \cdot i_{\beta_1,h}) + \lambda_k \left( \sum_{l=1}^{m} [T^{-1}]_{kl} [i_{\alpha\beta}]_l - i_k(t) \right)
\] (F.5)

Instead of directly looking for the stationary points of function \( F \) with the constraints related to the MMF waveform, it is easier to remove those restrictions in a first step and impose the current constraints latter. By deriving the previous function with respect to each one of its components and equating to zero:

\[
\frac{\partial F}{\partial [i_{\alpha\beta}]_l} = 2 [i_{\alpha\beta}]_l + \lambda_k [T^{-1}]_{kl} = 0, \quad l = 1... m
\] (F.6)

Imposing now that \( i_{\alpha_1} \) and \( i_{\beta_1} \) maintain their pre-fault values (with the optional derating):

\[
\begin{align*}
[i_{\alpha\beta}]_1 &= \chi \cdot i_{\alpha_1,h} \\
[i_{\alpha\beta}]_2 &= \chi \cdot i_{\beta_1,h} \\
[i_{\alpha\beta}]_l &= -1/2\lambda_k [T^{-1}]_{kl}, \quad l = 3... m
\end{align*}
\] (F.7)

and substituting into the fault constraint (F.4):

\[
[i_{ph}]_k = [T^{-1}]_{k1} \chi \cdot i_{\alpha_1,h} + [T^{-1}]_{k2} \chi \cdot i_{\beta_1,h} - \frac{1}{2} \lambda_k \sum_{l=3}^{m} [T^{-1}]_{kl} [T^{-1}]_{kl} = i_k(t)
\] (F.8)

Therefore:

\[
\lambda_k = 2 \frac{[T^{-1}]_{k1} \chi \cdot i_{\alpha_1,h} + [T^{-1}]_{k2} \chi \cdot i_{\beta_1,h} - i_k(t)}{\sum_{l=3}^{m} [T^{-1}]_{kl} [T^{-1}]_{kl}}
\] (F.9)

and:

\[
[i_{\alpha\beta}]_l = - [T^{-1}]_{kl} \frac{[T^{-1}]_{k1} \chi \cdot i_{\alpha_1,h} + [T^{-1}]_{k2} \chi \cdot i_{\beta_1,h} - i_k(t)}{\sum_{l=3}^{m} [T^{-1}]_{kl} [T^{-1}]_{kl}} , \quad l = 3... m
\] (F.10)

Although equation (F.10) may seem complicated at first glance, all the terms besides the currents are constant value parameters, so the previous equation reduces to a very simple expression once particular conditions are considered. For example, in the case of a 5-phase machine, if the faulted phase corresponds to phase A \((k = 1)\), the current references become:
the derived control strategy is specially well suited for the current control in the and stored in look-up tables in order to be applied in the post-fault operation mode of the drive. Hence, as:

\[ k \]

Let now \( k \) be the phase indexes for which the controlability of the currents has been lost: \( i_{α, h} \) and \( i_{β, h} \). Thus, in case of a fault, the current references expressed in the \( α – β \) reference frame become a linear combination of the first harmonic current components in the pre-fault state: \( i_{α, h} \) and \( i_{β, h} \), plus a term depending on the current in the faulted phase. Different coefficients for the currents in the \( α – β \) axes are obtained depending on the faulted phase index, \( k \). These linear coefficients can be easily pre-calculated and stored in look-up tables in order to be applied in the post-fault operation mode of the drive. Hence, the derived control strategy is specially well suited for the current control in the \( α – β \) reference frame [414].

\[ F.11 \]

\[ F.12 \]

Thus, in case of a fault, the current references expressed in the \( α – β \) reference frame become a linear combination of the first harmonic current components in the pre-fault state: \( i_{α, h} \) and \( i_{β, h} \). Different coefficients for the currents in the \( α – β \) axes are obtained depending on the faulted phase index, \( k \). These linear coefficients can be easily pre-calculated and stored in look-up tables in order to be applied in the post-fault operation mode of the drive. Hence, the derived control strategy is specially well suited for the current control in the \( α – β \) reference frame [414].

\[ F.13 \]

\[ F.14 \]

F.2.2 Without control over the zero sequence current

In the case the zero sequence current cannot be controlled, a further constraint must be added to the minimization problem in order to reflect the fact that \( i_0 \) is forced to follow a certain time expression: \( [i_{α, β}]_m = i_0 (t) \). By introducing the previous restriction into equation (F.7) and following the same procedure, the next expression is derived for \( l = 3... m − 1 \):

\[ \frac{[T^{-1}]_{kl}}{\sum_{l=3}^{m-1} [T^{-1}]_{kl}} [T^{-1}]_{kl} \left( \begin{array}{c} i_{α, h} \\ i_{β, h} \\ i_{α, h} \\ i_{β, h} \\ i_0 \end{array} \right) = \left( \begin{array}{c} i_0 (t) \\ i_0 (t) \\ i_0 (t) \end{array} \right) \]

For example, in the case of a 5-phase machine with a faulted phase \( k = 1 \) and no control over the zero sequence current:

\[ F.14 \]

F.3 Two phase fault

A fault scenario involving the loss of control over two phase currents can happen either with the open-circuit of two phases or the short-circuit between the machine terminals of two phases, for instance. The analysis for such a fault condition is analogous to the previously conducted one, though the number of constraints to be considered and the number of terms in the corresponding equations increase.

F.3.1 With control over the zero sequence current

Let now \( k \) and \( j \) be the phase indexes for which the controlability of the currents has been lost: \( [i_{ph}]_k = i_k (t) \), \( [i_{ph}]_j = i_j (t) \). By applying the reverse Clarke transformation, the fault conditions can be stated as:

\[ F.14 \]
F.3. TWO PHASE FAULT

Hence, the function to be minimized, without the constraints corresponding to the main MMF harmonic, is:

\[
F = \sum_{l=1}^{m} [i_{\alpha\beta}]_l^2 + \lambda_k \left( \sum_{l=1}^{m} [T^{-1}]_{kl} [i_{\alpha\beta}]_l - i_k (t) \right) + \lambda_j \left( \sum_{l=1}^{m} [T^{-1}]_{jl} [i_{\alpha\beta}]_l - i_j (t) \right) \tag{F.15}
\]

Deriving function \( F \) with respect to each current component and equating to zero:

\[
\frac{\partial F}{\partial [i_{\alpha\beta}]} = 2 [i_{\alpha\beta}]_l + \lambda_k [T^{-1}]_{kl} + \lambda_j [T^{-1}]_{jl} = 0, \quad l = 1 \ldots m \tag{F.16}
\]

By imposing now that \( i_{\alpha_1} \) and \( i_{\beta_1} \) are a function of their pre-fault values:

\[
\begin{align*}
[i_{\alpha\beta}]_1 &= \chi \cdot i_{\alpha_1, h} \quad \quad [i_{\alpha\beta}]_2 = \chi \cdot i_{\beta_1, h} \\
[i_{\alpha\beta}]_3 &= -1/2 \lambda_k [T^{-1}]_{kl} - 1/2 \lambda_j [T^{-1}]_{jl}, \quad l = 3 \ldots m 
\end{align*} \tag{F.17}
\]

and substituting into the original constraints (F.14):

\[
\begin{align*}
[T^{-1}]_{kl} \chi \cdot i_{\alpha_1, h} + [T^{-1}]_{k2} \chi \cdot i_{\beta_1, h} - \frac{1}{2} \lambda_k \sum_{l=3}^{m} [T^{-1}]_{kl} [T^{-1}]_{kl} - \frac{1}{2} \lambda_j \sum_{l=3}^{m} [T^{-1}]_{jl} [T^{-1}]_{jl} - i_k (t) &= 0 \\
[T^{-1}]_{jl} \chi \cdot i_{\alpha_1, h} + [T^{-1}]_{j2} \chi \cdot i_{\beta_1, h} - \frac{1}{2} \lambda_k \sum_{l=3}^{m} [T^{-1}]_{jl} [T^{-1}]_{jl} - \frac{1}{2} \lambda_j \sum_{l=3}^{m} [T^{-1}]_{jl} [T^{-1}]_{jl} - i_j (t) &= 0 
\end{align*} \tag{F.18}
\]

The previous equation establishes a linear system for parameters \( \lambda_k \) and \( \lambda_j \) that can be readily solved:

\[
\begin{bmatrix}
a_{kk} & a_{kj} \\
a_{jk} & a_{jj}
\end{bmatrix}
\begin{bmatrix}
\lambda_k \\
\lambda_j
\end{bmatrix}
= 
\begin{bmatrix}
b_k \\
b_j
\end{bmatrix} \tag{F.19}
\]

where:

\[
\begin{align*}
a_{kk} &= \frac{1}{2} \sum_{l=3}^{m} [T^{-1}]_{kl} [T^{-1}]_{kl} \\
a_{kj} &= \frac{1}{2} \sum_{l=3}^{m} [T^{-1}]_{kl} [T^{-1}]_{jl} \\
a_{jk} &= \frac{1}{2} \sum_{l=3}^{m} [T^{-1}]_{jl} [T^{-1}]_{kl} \\
a_{jj} &= \frac{1}{2} \sum_{l=3}^{m} [T^{-1}]_{jl} [T^{-1}]_{jl}
\end{align*} \tag{F.20}
\]

\[
\begin{align*}
b_k &= [T^{-1}]_{kl} \chi \cdot i_{\alpha_1, h} + [T^{-1}]_{k2} \chi \cdot i_{\beta_1, h} - i_k (t) \\
b_j &= [T^{-1}]_{jl} \chi \cdot i_{\alpha_1, h} + [T^{-1}]_{j2} \chi \cdot i_{\beta_1, h} - i_j (t)
\end{align*} \tag{F.21}
\]

Once the values of the Lagrange multipliers have been obtained, they can be substituted in equation (F.17) in order to compute the suitable current references. For instance, in the case of a 5-phase machine in which the faulted phases are phases C and D \((k = 3, j = 4)\), the current references become:
while if the faulted phases are $B$ and $E$ $(k = 2, j = 5)$:

\[
\begin{align*}
  i_{a_1} &= \chi \cdot i_{a_{1, h}} \\
  i_{a_3} &= 0.4198 \cdot \chi \cdot i_{a_{1, h}} + 0.2595 \cdot (i_C (t) + i_D (t)) \\
  i_{\beta_3} &= 0.6180 \cdot i_{\beta_{1, h}} + 0.5257 \cdot (i_C (t) - i_D (t)) \\
  i_0 &= 0.9607 \cdot i_{a_{1, h}} + 0.5937 \cdot (i_C (t) + i_D (t))
\end{align*}
\]

### F.3.2 Without control over the zero sequence current

The final case considered is the fault of two phases without control over the zero sequence current. By imposing that $[i_{a_3}]_m = i_0 (t)$ in equation (F.17), an analogous linear system to that in equation (F.19) is derived for the Lagrange multipliers with the following coefficients:

\[
\begin{align*}
  a_{kk} &= \frac{1}{2} \sum_{l=3}^{m-1} [T^{-1}]_{kl} [T^{-1}]_{kl} \\
  a_{kj} &= \frac{1}{2} \sum_{l=3}^{m-1} [T^{-1}]_{kl} [T^{-1}]_{jl} \\
  a_{jk} &= \frac{1}{2} \sum_{l=3}^{m-1} [T^{-1}]_{jl} [T^{-1}]_{kl} \\
  a_{jj} &= \frac{1}{2} \sum_{l=3}^{m} -1 [T^{-1}]_{jl} [T^{-1}]_{jl}
\end{align*}
\]

\[
\begin{align*}
  b_k &= [T^{-1}]_{kl} \chi \cdot i_{a_{1, h}} + [T^{-1}]_{k2} \chi \cdot i_{\beta_{1, h}} + [T^{-1}]_{km} \chi \cdot i_0 (t) - i_k (t) \\
  b_j &= [T^{-1}]_{jl} \chi \cdot i_{a_{1, h}} + [T^{-1}]_{j2} \chi \cdot i_{\beta_{1, h}} + [T^{-1}]_{jm} \chi \cdot i_0 (t) - i_j (t)
\end{align*}
\]

Solving the aforementioned linear system and substituting in (F.17), the current references are obtained. For instance, for a 5-phase machine with the phases $C$ and $D$ faulted $(k = 3, j = 4)$:

\[
\begin{align*}
  i_{a_1} &= \chi \cdot i_{a_{1, h}} \\
  i_{a_3} &= 2.6180 \cdot \chi \cdot i_{a_{1, h}} + 0.1432 \cdot i_0 (t) + 1.6180 \cdot (i_C (t) + i_D (t)) \\
  i_{\beta_3} &= -0.6180 \cdot \chi \cdot i_{\beta_{1, h}} + 0.2937 \cdot i_0 (t) + 0.5257 \cdot (i_C (t) - i_D (t))
\end{align*}
\]

while if the faulted phases are $B$ and $E$ $(k = 2, j = 5)$:

\[
\begin{align*}
  i_{a_1} &= \chi \cdot i_{a_{1, h}} \\
  i_{a_3} &= 0.3820 \cdot \chi \cdot i_{a_{1, h}} - 0.0603 \cdot i_0 (t) - 0.6180 \cdot (i_B (t) + i_E (t)) \\
  i_{\beta_3} &= 1.6180 \cdot \chi \cdot i_{\beta_{1, h}} - 0.5642 \cdot i_0 (t) - 0.8507 \cdot (i_B (t) - i_E (t))
\end{align*}
\]
Appendix G

Current Control in the $\alpha - \beta$ Reference Frame

The machine currents in a PMSM can be controlled referred to the $\alpha - \beta$ reference frame much in the same fashion as they are traditionally controlled in the synchronously rotating $d-q$ reference frame. The main difference between both control methods lies in the definition of the transfer function between the machine voltages and currents. From the machine constitutive equations:

\[
\begin{align*}
\lambda_{d_1} &= \lambda_{m_1} + L_{d_1} i_{d_1} \\
\lambda_{q_1} &= L_{q_1} i_{q_1} \\
\lambda_{d_3} &= \lambda_{m_3} + L_{d_3} i_{d_3} \\
\lambda_{q_3} &= L_{q_3} i_{q_3} \\
\lambda_0 &= L_0 i_0
\end{align*}
\]  

(G.1)

after some tedious math involving the transformation matrices defined in Appendix B, it can be demonstrated that the flux-linkages expressed in the $\alpha - \beta$ reference frame are:

\[
\begin{align*}
\lambda_{\alpha_1} &= \lambda_{m_1} \sin (\theta_e) + L_{\alpha_1} i_{\alpha_1} + L_{\alpha_1 \beta_1} i_{\beta_1} \\
\lambda_{\beta_1} &= -\lambda_{m_1} \cos (\theta_e) + L_{\beta_1} i_{\beta_1} + L_{\beta_1 \alpha_1} i_{\alpha_1} \\
\lambda_{\alpha_3} &= \lambda_{m_3} \sin (3\theta_e) + L_{\alpha_3} i_{\alpha_3} + L_{\alpha_3 \beta_3} i_{\beta_3} \\
\lambda_{\beta_3} &= -\lambda_{m_3} \cos (3\theta_e) + L_{\beta_3} i_{\beta_3} + L_{\beta_3 \alpha_3} i_{\alpha_3} \\
\lambda_0 &= L_0 i_0
\end{align*}
\]  

(G.2)

where:

\[
\begin{align*}
L_{\alpha_1} &= L_{d_1} \sin^2 (\theta_e) + L_{q_1} \cos^2 (\theta_e) \\
L_{\beta_1} &= L_{d_1} \cos^2 (\theta_e) + L_{q_1} \sin^2 (\theta_e) \\
L_{\alpha_3} &= L_{d_3} \sin^2 (3\theta_e) + L_{q_3} \cos^2 (3\theta_e) \\
L_{\beta_3} &= L_{d_3} \cos^2 (3\theta_e) + L_{q_3} \sin^2 (3\theta_e) \\
L_{\alpha_1 \beta_1} &= L_{\beta_1 \alpha_1} = -\frac{1}{2} (L_{d_1} - L_{q_1}) \sin (2\theta_e) \\
L_{\alpha_3 \beta_3} &= L_{\beta_3 \alpha_3} = -\frac{1}{2} (L_{d_3} - L_{q_3}) \sin (6\theta_e)
\end{align*}
\]  

(G.3)

and $\lambda_{m_1}$, $\lambda_{m_3}$, $L_{d_1}$, etc. are assumed to be constant value parameters that do not depend on the electrical angle for the stator variables $\theta_e$. For a surface-mounted PMSM under no saturation conditions, it is approximately: $L_{d_1} = L_{q_1} = L_{d_3} = L_{q_3} = L_0 = L_{ph}$ and the $\alpha - \beta$ inductance expressions reduce to:
\[ \begin{cases} L_{\alpha_1} = L_{\beta_1} = L_{\alpha_3} = L_{\beta_3} = L_{ph} \\ L_{\alpha_1\beta_1} = L_{\beta_1\alpha_1} = L_{\alpha_3\beta_3} = L_{\beta_3\alpha_3} = 0 \end{cases} \] (G.4)

Hence, the machine voltage equations referred to the \( \alpha - \beta \) reference frame become:

\[
[u_{\alpha\beta}] = R_{ph} [i_{\alpha\beta}] + \frac{d}{dt} [\lambda_{\alpha\beta}] = R_{ph} [i_{\alpha\beta}] + L_{ph} \frac{d[i_{\alpha\beta}]}{dt} + [e_{\alpha\beta}] \] (G.5)

where \([e_{\alpha\beta}] = [e_{\alpha_1} \quad e_{\beta_1} \quad e_{\alpha_3} \quad e_{\beta_3} \quad e_0]^T:\)

\[
\begin{align*}
e_{\alpha_1} &= \omega \epsilon_1 \lambda_{m_1} \cos (\theta_e) \\
e_{\beta_1} &= \omega \epsilon_1 \lambda_{m_1} \sin (\theta_e) \\
e_{\alpha_3} &= 3\omega \epsilon_1 \lambda_{m_1} \cos (3\theta_e) \\
e_{\beta_3} &= 3\omega \epsilon_1 \lambda_{m_1} \sin (3\theta_e) \\
e_0 &= 0
\end{align*}
\] (G.6)

Under the previous assumption and from the current control viewpoint, all the \( \alpha - \beta \) current components have the same transfer function:

\[
G(s) = \frac{I_{\alpha\beta}(s)}{U_{\alpha\beta}(s)} = \frac{1}{R_{ph} + L_{ph}s} \] (G.7)

and thus, they can be controlled by using the same control structure, as shown in Figure G.1. In the control structure displayed in the aforementioned figure, standard PI controllers are used and the induced voltage vector \([e_{\alpha\beta}]\) is introduced as a feed-forward compensation term in order to improve the system’s response. Once the \( \alpha - \beta \) voltage references generated by the current controllers are calculated, they can be transformed to the fixed stator reference frame in order to serve as inputs of the PWM voltage supply strategy: \([u_{phN}^*] = [T^{-1}] [u_{\alpha\beta}^*].\)

Figure G.1: Structure of the \( \alpha - \beta \) current control loop