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Advanced Power Loss Modeling and Model-Based Control of Three-Phase Induction Motor Drive Systems

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Advanced Power Loss Modeling and Model-Based Control of Three-Phase Induction Motor Drive Systems

Yiqi Liu, Ph.D.
University of Connecticut, 2017

Three-phase induction motor (IM) drive systems are the most important workhorses of many industries worldwide. This dissertation addresses improved modeling of three-phase IM drives and model-based control algorithms for the purpose of designing better IM drive systems. Enhancements of efficiency, availability, as well as performance of IMs, such as maximum torque-per-ampere capability, power density, and torque rating, are of major interest.

An advanced power loss model of three-phase IM drives is proposed and comprehensively validated at different speed, load torque, flux and input voltage conditions. This model includes a core-loss model of three-phase IMs, a model of machine mechanical and stray losses, and a model of power electronic losses in inverters. The drive loss model shows more than 90% accuracy and is used to design system-level loss minimization control of a motor drive system, which is integrated with the conventional volts-per-hertz control and indirect field-oriented control as case studies. The designed loss minimization control leads to more than 13% loss reduction than using rated flux for the testing motor drive under certain conditions. The proposed core-loss model is also used to design an improved model-based maximum torque-per-ampere control of IMs by considering core losses. Significant increase of torque-per-ampere capability could be possible for high-speed IMs. A simple model-based time-domain fault diagnosis method of four major IM faults is provided; it is nonintrusive, fast, and has excellent fault sensitivity and robustness to noise and harmonics. A fault-tolerant control scheme for sensor failures in closed-loop IM drives is also studied, where a multi-controller drive is proposed and uses different controllers with minimum hand-off transients when switching between controllers. A finite element analysis model of medium-voltage IMs is explored, where electromagnetic and thermal analyses are co-
simulated. The torque rating and power density of the simulated machine could be increased by 14% with proper change of stator winding insulation material.

The outcome of this dissertation is an advanced three-phase IM drive that is enhanced using model-based loss minimization control, fault detection and diagnosis of machine faults, fault-tolerant control under sensor failures, and performance-enhancement suggestions.
Advanced Power Loss Modeling and Model-Based Control of Three-Phase Induction Motor Drive Systems

Yiqi Liu

B.S., Central South University, 2012
M.S., University of Connecticut, 2016

A Dissertation
Submitted in Partial Fulfillment of the Requirements for the Degree of
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2017
APPROVAL PAGE

Doctor of Philosophy Dissertation

Advanced Power Loss Modeling and Model-Based Control of Three-Phase Induction Motor Drive Systems

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University of Connecticut
2017
To My Mother and Father

Shaoying Gao and Aizhong Liu
ACKNOWLEDGMENTS

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CHAPTER 1
INTRODUCTION

1.1 Overview

With the development of semiconductor devices, power electronic converters, electric machines, and machine control, inverter-fed electric motor drive systems have become the workhorse of the modern industry. They are widely used in various industrial, residential and commercial applications, such as in drilling, pumping, propulsion, heating, ventilation and air-conditioning systems, manufacturing processes and others [1]–[4]. Moreover, due to the increasing interests in vehicular electrification and the use of clean energy in micro-grid systems, such as using doubly-fed induction machines in wind power generation, utility of electric motor or generator drive systems in these applications has been significantly growing [5]–[8].

An electric motor drive system typically consists of four parts: source, power electronic drive, machine and load. As shown in Fig. 1, the source can be DC or AC in the single-phase, three-phase or higher-phase form. The drive is the “brain” of the system, which has a low-power control stage and a high-power energy conversion stage. This work specifically focuses on three-phase induction motor (IM) drive systems, which are the most commonly used. Compared to other electric motors, three-phase IMs are simple, robust and easy to control. They do not rely on rare-earth materials and thus are low-cost compared to permanent magnet synchronous machines (PMSMs). They also have a smoother torque response and lower torque ripple than switched reluctance machines (SRMs).

Due to the popularity and wide use of three-phase IM drive systems, improving the system’s efficiency, reliability, and performances, e.g. power density and torque rating, are of tremendous interest to researchers worldwide. It is reported that electric motor drive systems
account for more than 40% of the global electricity consumption [9], [10]. Three-phase IM drive systems are one of the largest single energy consumers. Therefore, increasing their efficiency is important for reducing global energy consumption and operating costs, as well as for reducing greenhouse gas emission and for environmental protection.

![Block diagram of electric motor drive systems](image)

**Fig. 1.** Block diagram of electric motor drive systems

IMs are designed of highest efficiency at rated operating condition. However, their efficiency could drop significantly at other operating conditions at rated flux. Typical relationships between motor efficiency and load conditions for IMs of different ratings are shown in Fig. 2 [11]. It is seen that considerable room for efficiency enhancement is available at low loads. Moreover, larger IMs tend to have higher efficiency and thus less room for efficiency enhancement in percentage. However, the same percentage increase of efficiency of large IMs can be converted to larger absolute values of energy savings. Model-based loss minimization control is convenient to integrate with the conventional control algorithms designed for speed and/or torque control. Properly modeling machine mechanical and stray losses, which are commonly ignored in the power loss model of electric motors, and nonlinear core loss are important to increase the accuracy of the model-based loss minimization control. Since the optimal operating point of an overall motor drive system, i.e. motor and drive combined, can be different from that of an individual motor or a drive [12], creating a system-level power loss model of IM drives is important to guarantee efficiency enhancement of the overall system. This model would need to integrate the machine copper loss, core loss, mechanical loss and stray loss.
as well as conduction and switching power electronic losses in the drive. The model has to be in a properly unified form for the design of loss minimization control.

![Graph showing efficiency of IMs with respect to load](image)

**Fig. 2.** Typical efficiency of IMs with respect to load [11]

Fault detection and diagnosis as well as fault-tolerant control of IM drives are important to avoid catastrophic failures, shutdown, associated repair and operational costs, etc., especially for safe-critical applications, such as electric vehicles, elevators and escalators. There are four major types of faults in IMs, which are eccentricity fault, bearing fault, broken rotor bar fault and stator short winding fault. Power electronic circuits in drives could also fail, such as open- or short-circuit faults of power switches. Moreover, the sensors in closed-loop IM drives could fail as well. Developing simple, low-cost, nonintrusive, unambiguous fault detection and diagnosis methods with adequate accuracy and sensitivity is the progressing direction. Moreover, fault-tolerant control after sensor failures is important to avoid immediate machine shutdown or loss of control, and to keep the operation continuity. Lower-performance controllers, even open-loop controllers, can be used as backups to deal with sensor failures, but special care needs to be taken regarding to the hand-off transient when switching between controllers.

Increasing power density, power and torque ratings are of great interest for large propulsion induction machines. One promising approach is to use new insulation material for
stator windings, which has higher thermal conductivity. Therefore, more current can be pushed into the machine without exceeding the original temperature limit of using the conventional micaceous insulation. Finite-element-analysis modeling is suitable for the co-simulation of electromagnetic and thermal analyses.

1.2 Problem Statement

There is a continuous pursuit of better three-phase IM drive systems that have higher efficiency, reliability and power density for current and future applications.

1.3 Research Statement

This dissertation provides an advanced power loss model of three-phase IM drive systems. This model includes a general dynamic core-loss model of three-phase IMs, a model of machine mechanical and stray losses, and a model of power electronic losses in inverters. Detailed illustration and derivation of the model is given as well as comprehensive simulation and experimental verifications. Model-based loss minimization control of an IM drive system is developed based on the proposed power loss model, which is integrated with the conventional volts-per-hertz (V/f) control and indirect field-oriented control (IFOC) as case studies. The proposed core-loss model can be used as a general basis for various control design in addition to loss minimization control. An improved model-based maximum torque-per-ampere (MTPA) control is proposed as another example of using the core-loss model, which could have higher torque-per-ampere capability than the conventional MTPA control by considering core losses in the control design. Fault detection and diagnosis methods of the four major types of faults in IMs are comprehensively reviewed. A new model-based fault detection and diagnosis method is provided, which is simple, nonintrusive, robust and can detect all four major types of IM faults. Fault-tolerant control of closed-loop IM drives with sensor failures is studied, where a new
synchronous-frame multi-controller drive and control are provided to solve this problem with minimum hand-off transients when switching between controllers. Finite-element-analysis modeling of medium-voltage IMs (MVIMs) is explored for large propulsion applications, where re-rating the machine to higher power density, power, and torque is possible by using a new nano-structured insulation material for stator winding.

The dissertation proceeds as follows: Chapter 2 gives the literature review on various models of three-phase IMs, loss minimization control, fault detection and diagnosis methods of IMs and fault-tolerant control of IMs with sensor failures. Chapter 3 thoroughly describes the proposed power loss model of three-phase IMs, which includes the core-loss model for copper and core losses, and the model for mechanical and stray losses. The model of power electronic losses in inverters is introduced in Chapter 4 and is integrated with the machine loss model. Mode-based loss minimization control algorithms are designed for individual IMs and overall IM drive systems, and integrated with volts-per-hertz (V/f) control and indirect field-oriented control (IFOC) in Chapter 4. An improved model-based maximum torque-per-ampere (MTPA) control by considering core losses is also discussed in Chapter 4. Chapter 5 explains the proposed model-based time-domain fault detection and diagnosis method of the four major types of IM faults. It also illustrates the proposed synchronous-frame multi-controller drive and control for IMs under sensor failures. Chapter 6 discusses using silver rotor bar in IMs to improve machine efficiency. It also describes the multi-physics finite-element-analysis modeling and simulation of medium-voltage IMs as well as the re-rating process. Conclusions and contributions of this dissertation are given in Chapter 7.
CHAPTER 2
LITERATURE REVIEW

2.1 Modeling of Losses in Three-Phase IMs

Machines are generally designed for full-load conditions where the copper loss is dominant. Core loss gradually takes the dominance as the load decreases. Therefore, incorporating core loss is important in the modeling of machines that frequently operate at relatively low-load conditions. But even under high-load conditions, considering core loss can lead to better estimates of a machine’s total loss and efficiency, and render more accurate model-based analysis and control design. Various IM models have been proposed in the literature for different applications. Two popular models used in the literature are: 1) The per-phase equivalent circuit [13], and 2) The dynamic three-phase model [14]. The first model is simple, but it cannot work in dynamic conditions neither perform qd0-frame transform, which is the basis of advanced vector control algorithms. The second model does not have the same two issues as the first one, but it cannot estimate core loss or iron loss. These two models will be shown in Chapter 3.

There are several other analytical IM models in the literature. In [15], an arbitrary qd0-frame model has been proposed with the core loss being expressed directly as parallel resistors in magnetizing branches of d-q equivalent circuits. But the model is proposed for steady-state vector controller design and the model accuracy is not provided. In [16], [17], simplified d-q axis equivalent circuits are proposed for IM loss minimization control, where core loss resistor is immediately after/before the q-axis stator resistor, respectively. This model is shown in Fig. 4 and notations of variables are referred to [17]. These simplified equivalent circuits ignore stator leakage inductance and are only valid in the rotor reference frame. A modified version of this model is proposed in [18] which includes leakage inductances. But the model still only works for
rotor reference frame, and the core loss resistance is assumed to be independent of frequency. Moreover, elaborated IM models that use winding functions are applied in [19] and [20]. But these models are generally too complicated for controller design. Empirical models are also proposed in the literature, such as the classical Steinmetz’s equation and its modified versions for core loss estimation [21], [22]. Similarly, IM total loss is modelled as a complex polynomial function of slip in [23]. These “black box” models are lack of internal interpretation, and are heavily rely on the accuracy and completeness of training data. Finite element analysis (FEA) [24], [25] and artificial intelligence (AI) [26] are also used in the model-based IM analysis. Compared to other types of models, analytical models are the most suitable for design of flux observers and machine controllers. For example, in the loss-minimisation control design, the optimal control variable can be solved analytically, numerically or iteratively based on the analytical relationship between the machine loss and the control variable(s) [27]–[29].

Fig. 3. Core-loss model used in [15]
Compared to sinusoidal voltage source, PWM excitation induces additional machine copper and core losses, or PWM harmonic losses, due to the higher harmonics as well as possible negative thermal effects and change in the machine’s operating point [30]. It is indicated in [31] that the PWM harmonic copper loss is typically larger than the PWM harmonic core loss at the low-frequency end of the harmonic spectrum, but it decreases fast and can be neglected at high harmonic frequencies. On the other hand, the PWM harmonic core loss is mainly composed of eddy-current loss in the high frequency range of the harmonic spectrum as shown in [32]. The hysteresis loss component of the PWM core loss is approximately inversely proportional to switching frequency [33]. It is shown in [34], [35] that the amplitude modulation index plays an important role in PWM harmonic core loss, whereas the impacts of modulation function waveform and load condition are insignificant. If skin effect is considered, the PWM harmonic core loss will decrease slightly with harmonic frequency at high frequencies, and the decrease of the PWM harmonic copper loss with frequency will be significantly slower. The PWM harmonic loss can be decided by curve fitting of experimental harmonic loss-factor curve [31], or by FEA model [35].

Mechanical loss is a friction type of loss, which can be modelled as a function of speed or using a look-up table to store the values for different speeds. Stray loss is the most complex and least studied type of major losses in IMs. It includes all types of machine losses except
conventionally defined copper, core and mechanical losses. Although it occupies only a few percentages of machine output power, accurate estimation of stray loss is critical in precise machine efficiency estimation, control design and efficiency enhancement. Complete causes of stray loss are still unknown and are actively being studied by many researchers [36]–[41]. But it has been recognized recently that the space-harmonic induced additional losses in rotor cage, stator core and rotor core (including pulsating loss, rotor surface loss and mainly the inter-bar current loss) are the main components of stray loss [38], [39]. FEA can model stray loss by considering space harmonics, but it is very time-consuming and cannot be used for controller design. It is common to decide stray loss experimentally following standards such as [42], where the stray loss is estimated as a function of torque squared. Only a few of circuit-based stray-loss models are available in the literature, where the majority of the stray loss is lumped into an additional resistor in the circuit [40], [41].

Based on the literature review, there is no dynamic core-loss model of three-phase IMs, which can perform $qd0$-frame analysis and design in any reference frame while considering core loss. Mechanical and stray losses are commonly ignored in the power loss model, which decreases the model accuracy for loss minimization control. These problems are tackled in Chapter 3.

2.2 Loss Minimization Control of Electric Motor Drives

Due to the significant economic, energy and environment effects, loss minimization control (LMC) of electric motor drives is being greatly studied. Most of the methods just target electric motors instead of motor drive systems [12], since electric motors usually consume the majority of the energy. However, optimizing the efficiency of the overall motor drive is the progressing direction, since 1) The energy consumption in drives sometimes could be not
negligible, e.g. drives having high switching frequency and/or motors running at very low-power conditions in a dynamic operation cycle; 2) Further efficiency enhancement over the system is required.

There are different ways to classify the LMC methods of electric motors. Depending on the load variability, the LMC can be defined as dynamic or steady-state types. The dynamic loads usually are used in transportation, such as urban electric vehicles [43], electric ships [44], and traction motors, such as elevators and escalators [45]. The steady-state loads usually are used in pumps, fans, automatic manufacturing systems, etc. Depending on whether the LMC can minimize loss in real-time, the LMC can be defined as offline and online types. Offline methods are simple. They set the optimal values of some variables, e.g. flux, in the controller before running the machine. These optimal values are usually stored in the factory-setting of the machine [46]. Online LMC are more complicated. They require sensing feedback to adjust the command iterative during the operation of the machine, and thus have self-adjust capability to load or excitation changes [47]. Depending on whether the LMC relies on a model of electric machines, the LMC can also be defined as model-based, physics-based and hybrid types [48].

Model-based LMC processes the feedback through machine models to calculate the optimal command values. The accuracy and completeness of the machine model is the key for the model-based LMC. Different control variables are used in different models to minimize losses which are functions of different model parameters. Some examples of the control variables are slip frequency [23], [49], magnetizing flux [22], rotor flux and d-axis rotor flux component [50], [51].

Maximum torque-per-ampere (MTPA) control is a straightforward model-based control method to improve torque-per-ampere capability as well as to increase machine efficiency.
Given speed and torque, certain slip and current pattern can lead to decreased sum of machine copper and core losses, which increases machine efficiency. When including machine drives in efficiency assessment, MTPA control is further appreciated since the losses in drives are positively related to motor currents [52]. The conventional MTPA control was designed based on the conventional dynamic copper-loss model of three-phase IMs ignoring core loss [53]. Later implementation and modification of the conventional MTPA control in vector controllers [54] and scalar controllers [46] inherit this property without considering core loss. Only one paper that includes core loss in MTPA analysis is found [55], but no detailed derivation or controller design is provided.

Physics-based LMC requires no model of machines and can control machines as a black box. They achieve the optimal command values iteratively through monitoring the relationship between the command and loss. For example, when using Perturbation & Observation method for loss minimization control, if increase the control variable can decrease loss in the present iteration, then increase the control variable for the next iteration as well. Otherwise, decrease the control variable in the next iteration [56]–[58]. More advanced decision-making algorithm than the logic-type Perturbation & Observation can also be used, such as fuzzy logic [59], [60].

The hybrid LMC has features of the model-based and physics-based LMC at the same time, thus the term “hybrid”. Hybrid LMC has various forms. For example, they can use model-based LMC to roughly decide the optimal command values and then use physics-based LMC to refine the optimal commands [61]. Moreover, hybrid LMC can use machine’s electromechanical properties to deal with model-based parameters, thus minimize losses [62], [63].

Different LMC have different advantages and disadvantages. The selection of LMC varies for various applications. Generally, convergence rate, steady-state error, and parameter
dependence are the three major considerations to decide LMC. There are tradeoffs among the three considerations. For dynamic loads, convergence rate is more important than steady-state error. Off-line LMC only stores optimal commands for several operating points, e.g. rated point and the most frequent operating points. Thus, they can be inaccurate for other operating points. Even for the saved operating points, the optimal commands could be inaccurate due to the change of machine conditions, e.g. slight demagnetizing of the core, and environmental conditions, e.g. temperature and electromagnetic interference. Online methods are more accurate, but they require more powerful processors, and their reliability is decreased due to the possible failure of feedback sensors.

Model-based LMC methods have high convergence rate and are convenient for controller design. Their accuracy and application conditions depend on the accuracy of the model, which usually has a tradeoff with complexity. It is not an easy work to build an accurate and simple machine model that can be applied for a wide range of ratings and operating conditions. Model-based LMC also greatly depend on the accuracy of the model parameters. Some model parameters are changed with operating conditions. Thus, proper parameter adaptation is needed before using the model and the model-based LMC. Due to the possible imperfectness of the model and model parameters, model-based LMC can lead to sub-optimal operating points instead of the actual optimal operating point. The difference between the sub-optimal and optimal operating points depends on the accuracy of the model. Physics-based LMC can be applied to a wide range of machines and operating conditions without needing of machine parameters. Since they keep chasing the optimal operating points, the steady-state error could be small. However, they usually have small convergence rate, since they need to take the feedback signal to improve the control command iteratively. If the convergence rate is increased, the
bouncing around the optimal operating points is also increased and the system could become unstable. Hybrid LMC take the advantages of the previous two types of LMC. They generally have faster convergence rate than physics-based LMC as well as smaller steady-state error and less parameter-dependence than those of model-based LMC. The only possible disadvantage of hybrid LMC is complexity which could be overcome by powerful digital controllers.

*Based on the literature review, unified system-level power loss model of IM drive systems is rarely shown in the literature [64]. Online analytical model-based loss minimization control of the overall system is not found either. These problems are addressed in Chapter 4. Chapter 4 also introduces an improved MTPA control that considers core loss to increase the accuracy.*

### 2.3 Fault Detection and Diagnosis of IMs And Fault-tolerant Control of IMs with Sensor Failures

#### 2.3.1 Fault Detection and Diagnosis of IMs

#### 2.3.1.1 Major Faults in IMs

IMs (IMs) are popularly integrated in equipment and used in many manufacturing processes, industrial applications and facilities. It is important to maintain the health of IMs to keep many industries running well. However, various faults frequently happen in IMs due to tough working conditions, regular wear and tear, enduring and/or overrated loads, unexpected events and many others. Thus, fault detection and diagnosis (FDD) is important to avoid catastrophic failures, shutdown, associated repair and operational costs, and unsafe operation of IMs. The main components in IMs are the stator core and laminations, rotor core and laminations, stator windings, rotor windings or bars, insulating material, shaft, bearings, and housing. Stator and rotor cores and laminations are usually rigid and reliable where their faults could occur during the manufacturing process. Machine housing is also secure if appropriately
grounded and protected even in harsh environmental conditions. As for IM itself, there are four major types of faults that are actively studied in literature: Air-gap eccentricity fault (EF), bearing fault (BF), broken rotor bar fault (BRBF) and stator short winding fault (SSWF).

Air-gap eccentricity caused by bearing or shaft inconsistency is a mechanical type of faults, where the spacing between stator and rotor is not uniform. Three types of eccentricities are usually encountered: static EF, dynamic EF and mixed EF of the previous two, as shown in Fig. 5 [65]. The air gap in the figure is exaggerated for illustrative purpose. For static EF, the air-gap length at each point along circumference is constant, but different positions have different air-gap length. It is usually caused by the displacement of rotor physical center and stator center. In contrast, air-gap length changes periodically at each position along the circumference in dynamic EF. The oval rotor shape and the departure of rotating center from the rotor physical center caused by worn bearings are common reasons for this type of eccentricities. In practice, the two types of eccentricities exist simultaneously forming the third type. Misalignment between shafts is one of the main reasons that can lead to air-gap eccentricity fault in practice [66], [67].

![Fig. 5. Different air-gap eccentricity faults](image_url)
Bearings are used to support rotors and to decrease rotational friction. They usually consist of four parts: inner race, outer race, rolling element and a cage that restricts the relative movement between different rolling elements. Rolling elements have various shapes depending on the applications. The most common ball type rolling elements are used here for illustration purpose as shown in Fig. 6. Bearings can fail even with proper use of motor due to fatigue and wear. Insufficient lubrication, high load, enduring operation, high ambient temperature, etc. can accelerate BF. A BF originates from distributed types, such as raceway roughness and waviness, and then develops to local types, such as cracks, pits and spalls [68]. Based on the location of the local fault, BF can be subdivided into four types: inner-race, outer-race, rolling-element, and cage BFs [69].

Fig. 6. Detailed structure of a bearing with ball-type rolling element

Another type of mechanical faults in a squirrel-cage IM is BRBF [70]. BRBF is mainly caused by intense thermal stress generated from large induced rotor current as well as other electrical, mechanical and environmental stresses. Once one rotor bar is broken, the adjacent rotor bars will have to take over the extra stresses from the broken rotor bar. This fact accelerates subsequent failures in adjacent rotor bars. BRBF can be modeled as a complete rotor circuitry with a “virtual zero current” flowing. This virtual zero current is considered as a superposition of
a “healthy current” as if the rotor bar is not broken and a virtual “faulty current” which has the same magnitude but an opposite sign of the healthy current. Consequently, the IM with a BRBF can be simply modeled as a complete and healthy machine with an extra “faulty current” in the rotor circuitry. The faulty current will generate an oppositely rotating MMF than healthy currents, which can be used for FDD.

The last actively studied fault type is SSWF which contains 1) interturn or turn-to-turn type; 2) coil-to-coil type; 3) phase-to-phase type; and 4) phase-to-ground type. These types are shown in Fig. 7. Here, the interturn SSWF is the most incipient one and the phase-to-ground SSWF is the most serious one. Unlike the previous three types of faults that are generally classified as mechanical faults, the SSWF is an electrical fault that accounts for a majority of electrical failure in IMs [71]. Stator open winding fault is another frequently discussed electrical fault, but not here, since this fault commonly occurs in power converters or drives rather than in IM itself. Interturn insulation breakdown causes SSWF and several factors can contribute to it including thermal, mechanical, and electrical stresses, etc. Details can be found in [72].

![Diagram](image.png)

**Fig. 7.** Different types of stator short winding faults: (a) interturn type; (b) coil-to-coil type; (c) phase-to-phase type; (d) phase-to-ground type
Faults in IMs generate abnormal features in different domains, which are used as fault indicators. These fault-indicative features can be extracted from voltage, current, magnetic, mechanical (vibration), chemical, acoustic, etc., signals using different sensors. The fault indicators in frequency domain are the most popular and well-understood ones since they can be feasibly detected by Fast Fourier Transform (FFT). Among all the feedback signals, stator current(s) of IMs is(are) the mostly used, since current sensors are relatively inexpensive and easy to use, and they are already installed in many motor drive systems for control purpose. Applying FFT on stator current feedback leads to the famous FDD method, Motor Current Signature Analysis (MCSA). The fault characteristic frequency components (FCFCs) of the four major IM faults in stator current spectrum are summarized in TABLE I.

2.3.1.2 Fault Detection and Diagnosis Methods for IMs

Numerous FDD methods for IMs have been reviewed and classified into four categories: time-domain, frequency-domain, time-frequency-domain, and artificial-intelligence-based (AI-based) methods. Since many varieties exist in AI-based methods which involve all the previous three domains, the AI-based methods are treated as a separate category.

For time-domain FDD methods, they are implemented through checking abnormal changes of interested machine features along with time. These methods usually have advantages of simple calculation and implementation, but they generally suffer from relatively low fault sensitivity. Thus, they could encounter difficulties when measuring fault-indicative components of incipient faults or in noisy environments. Two main focuses of recent time-domain FDD methods are finding fault-sensitive time-domain features and increasing fault detectability. Some of the recently published time-domain FDD methods are shown in [73]–[78].
### TABLE I. FCFCs of the Four Major IM Faults in Stator Current Frequency Spectrum

<table>
<thead>
<tr>
<th>Fault</th>
<th>Sub-Type</th>
<th>Fault Characteristic Frequency Components</th>
<th>Ref.</th>
</tr>
</thead>
<tbody>
<tr>
<td>EF</td>
<td>DE/ ME (low frequency range)</td>
<td>[f_{cf1} = f_c \left[1 \pm \frac{2k_i(1-s)}{P}\right]]</td>
<td>[69]</td>
</tr>
<tr>
<td></td>
<td>Principal slot harmonics (PSH)</td>
<td>[f_{cf2} = f_c \left[\frac{2(k_2N_s \pm N_d)(1-s) \pm N_w}{P}\right]]</td>
<td>[70]</td>
</tr>
<tr>
<td>BF</td>
<td>Inner-race BF</td>
<td>[f_{ir} = f_c + \frac{k_iN_b}{2} f_m \left[1 + \frac{D_b}{D_e} \cos \delta\right]]</td>
<td></td>
</tr>
<tr>
<td>BF</td>
<td>Outer-race BF</td>
<td>[f_{or} = f_c + \frac{k_iN_b}{2} f_m \left[1 - \frac{D_b}{D_e} \cos \delta\right]]</td>
<td></td>
</tr>
<tr>
<td>BF</td>
<td>Ball-type rolling-element BF</td>
<td>[f_b = f_c + \frac{k_bD_e}{D_b} f_m \left[1 - \left(\frac{D_b}{D_e} \cos \delta\right)^2\right]]</td>
<td>[79]</td>
</tr>
<tr>
<td>BF</td>
<td>Cage BF</td>
<td>[f_c = f_c + \frac{k_c}{2} f_m \left[1 - \frac{D_b}{D_e} \cos \delta\right]]</td>
<td></td>
</tr>
</tbody>
</table>
| BF                  | Inner-race/ Outer-race BF (simplified equation, for bearings of 6 to 12 ball-type rolling element) | \[
\begin{bmatrix}
  f_{ir} \\
  f_{or}
\end{bmatrix}
= f_c + \begin{bmatrix}
  0.6k_iN_b f_m \\
  0.4k_iN_b f_m
\end{bmatrix}
\] | [80]  |
| BRBF                |                               | \[f_{brbf} = f_c \left[1 \pm 2k_p \delta\right]\]                                                                            | [81]  |
| SSWF                |                               | \[f_{sswf} = f_c \left[\frac{2k_{10}(1-s)}{P} \pm k_{11}\right]\]                                                             | [81]  |

Machine faults generate additional frequency components in various spectra due to resultant periodical vibration of mechanical forces and air-gap spacing, which can be used for FDD. The simplest way to explore spectral properties of signals is to use FFT. However, spectral leakage, low resolution and long measurement period are some major drawbacks of FFT, which impede its usage in non-stationary conditions and for detecting FCFCs near fundamental frequency. Moreover, ambiguous FCFCs can be imitated from low-frequency load oscillation,
unbalanced voltage source, certain machine structures and faults other than the one to be detected. Therefore, many recent efforts have been put on unambiguous detection of FCFCs. Some of the recently published time-frequency-domain FDD methods are shown in [82]–[87].

Advanced time-frequency-domain signal processing techniques are superior than pure frequency-domain methods in dealing with non-stationary signals, which provide more accurate inspections of machine’s dynamic features via continuous spectral analysis using small moving time window. Tradeoffs of these detailed inspections are more complex computation and implementation. Some of the recently published time-frequency-domain FDD methods are shown in [88]–[91].

AI is an effective approach to model complex nonlinear systems by using certain structures and rules based on the understanding of system’s behavior. It places less emphasis on the physical structure and intermediate results of the system, but tries to emulate the input/output relationship directly. Popular AI-based methods are neural networks, fuzzy logic, genetic algorithm, particle swarm optimization, etc. Some of the recently published AI-based FDD methods are shown in [92]–[95].

2.3.1.3 Discussion of the Research Trend, Gap and New Ideas

Based on the review, some of the finds are summarized as follows: 1) Multi-fold/multi-sensor FDD has been more often discussed recently, which can increase FDD accuracy and thus is important for safety-critical applications; 2) Study of FDD for closed-loop inverter-fed motors is growing, where methods considering inverter switching harmonics and withstanding the regulation effect of speed/current loops are developing; 3) Efforts have been made for unambiguous FDD with respect to unbalanced source and oscillating load as well as for slip-dependent FDD at low-load conditions; 4) New sensors are explored, such as infrared
thermography; and 5) Most methods are nonintrusive. Current sensor is still the most popular one, then are voltage sensor and vibration sensor. Torque sensor and intrusive flux sensor are also used in some FDD methods.

Although significant development has been made over the last several years, a few of issues still remain unsolved. First, the distributed type of bearing faults, such as roughness and waviness, are barely studied. Instead of generating a spectral peak, this type of faults has broadband effects on the frequency spectrum. Thus, no FCFC can be easily found. Moreover, most researches and the FCFCs shown in TABLE I for BF are targeting ball-type bearings, modification of FCFCs for other types of bearings may be needed. Similarly, except for rotor asymmetry caused by BRBF, FDD of rotor end-ring fault is expected to see in the future. Second, although several papers have dealt with two simultaneous faults, most papers address single fault situations. The effects of simultaneous or cascaded faults with respect to different combinations of relative fault severity on each individual fault are desired to explore. Third, more research considering speed/current regulation effects in closed-loop control systems are valuable to study, since enforced regulation could significantly change machine’s behavior and suppress conventional FCFCs in feedback. Thus, modification of existing FDD methods and proposal of new fault indicators may be needed. Moreover, effects of inverter-induced harmonics on FCFCs are expected to discuss, since these harmonic-related interferences could seriously deteriorate FDD of incipient faults and of slip-dependent faults at low-load conditions. Intelligently utilizing inverter’s capability for FDD of IMs could be an interesting topic. Fourth, as the reviewed methods are mostly targeting a single motor, FDD of multi-motor or multi-drive systems, which are common in steeling processing and traction applications, is interesting to look at. Fifth, it is observed that structural and thermal effects on IMs are barely included in
FDD in the literature. However, increase of temperature during motor operation could change various fault indicators, such as model-based FCFCs (due to the change of model parameters with respect to temperature). Thus, it will be appreciated to study these factors on some of existing FDD methods. Last but not the least, an integrated methodology or system that can accurately identify all four major IM faults is not found, and quantified fault severity indicators are not standardized. However, such a comprehensive FDD system is very important to provide overall condition monitoring of IMs, especially for critical loads and safety-critical applications. It is believed that developing easy, low-cost, nonintrusive, unambiguous FDD methods with adequate accuracy and sensitivity as well as applicability for stationary and non-stationary conditions, and closed-loop inverter-fed conditions is the progressing direction.

A simple time-domain FDD method that can promptly detect all the four major types of IM faults without the need of additional sensors is introduced in Chapter 5.

2.3.2 Fault-tolerant Control of IMs with Sensor Failures

With growing popularity of variable frequency drives (VFDs), closed-loop controlled IM drives are no longer limited to high-end systems, but are widely applied in various applications [43], [96]–[98]. Sensor failure is one of the major factors that decrease the availability of closed-loop controlled drives. Various FDD methods have therefore been proposed in literature specifically concerning feedback sensors, such as speed/position encoders, current sensors, and voltage sensors [99]–[102]. On the other hand, fault-tolerant control has attracted increasing attention in the last two decades, for its availability and conservative design, in applications where keeping the continuity of operation is the paramount requirement [103], [104]. For safety-critical applications, such as electric vehicles (EV), even maintaining minimum machine operation after sensor failure is much better and safer than immediate machine shutdown. This
also the case for other applications that allow for degraded performance, such as cooling or heating pumps and fans.

As for the fault-tolerant control of IMs with sensor failures, the limited published methods can be divided into two types. The first is a resilient type of fault-tolerant control, where estimators or observers are used as remedial techniques to provide the information that is originally provided by the failed sensor. Therefore, the rest of the original control algorithm remains intact and the same control algorithm is used. A simple current estimator is used in [105] and fuzzy-based encoder and current observers are proposed in [106]. Moreover, various sensorless control of IMs can be used for fault-tolerant control of encoder failure, which also belongs to this type [107], [108]. However, sensorless control methods are sensitive to machine parameters and have poor performance at very low speeds.

The second is a reconfigurable type of fault-tolerant control, where a degraded controller that does not require the failed sensor is applied to replace the original controller. Multi-controller drives are built for fault-tolerant control and degraded controllers are used as backups of high-performance controllers. In [109], a four-controller drive is proposed to increase the drive’s reliability. The large hand-off transient between different controllers is handled by forcing the synchronization of the rotor flux’s phase angle, which is calculated from different controllers, at hand-off instances. The same multi-controller drive is also studied in [110]–[112] but with different methods to mitigate the large hand-off transients when switching controllers. In [110], the phase of the rotor flux in different controllers are monitored. Then, the switching of controllers is authorized only when the difference between the rotor flux’s phases is close to zero, or synchronization is naturally achieved. In [111], a fuzzy-based voltage command is used at the hand-off transient to compensate for the difference between the rotor flux’s phases. Only
simulation results are provided in [110] and [111], and it is indicated in [112] that the previous two transient smoothing methods are difficult to implement in experiments. Another multi-controller drive that uses a vector controller and a simple digital controller as a backup is proposed in [113] to deal with current sensor failure. To mitigate hand-off transients, it properly selects one of the eight space vectors of an inverter as the initial switching command of the digital controller. However, the digital controller cannot be switched to the vector controller smoothly in this drive during sensor recovery condition.

Based on the literature review, a reconfigurable type fault-tolerant control, which has smooth hand-off transients when switching between controllers, is not existing in the literature. To solve this problem, an advanced synchronous-frame IM drive and control are illustrated in Chapter 5.
CHAPTER 3
ADVANCED POWER LOSS MODELING OF THREE-PHASE INDUCTION MOTORS

3.1 A General Analytical Three-Phase IM Core-Loss Model in the Arbitrary Reference Frame

A general analytical core-loss model of three-phase IMs is proposed. This model applies virtual core-loss resistance in conventional dynamic IM model. Thus, it can perform $qd0$-frame transform while considering machine core loss. The $qd0$-frame transform is the basis of many closed-loop vector control of electric machines including IMs. Moreover, since the proposed model is a general model and is an improvement on a well-known model that is used for the design of various controllers and observers, the proposed model can be used to improve the accuracy and performance of those controllers and observers, and to design new model-based controllers and observers by considering core loss. Some of the examples that applies the proposed model to improve machine’s efficiency and torque-per-ampere capability are shown in Chapter 4.

The proposed model can be transformed into different reference frames as desired. Due to its analytical form and dynamic feature, the proposed model is suitable for controller and observer design. Parameters of the proposed model can be obtained conveniently from the machine characterization tests based on the IEEE Standard 112 [42]. Note that even though the proposed model uses parameters extracted from the steady-state characterization tests, the model itself (structure) is dynamic, which can deal with changing load conditions. The values of the model parameters need to adapt for different frequencies and flux levels.

The proposed model is inspired by the conventional dynamic model and the per-phase equivalent circuit model of IMs, which are shown in Fig. 8 and Fig. 9, respectively. In Fig. 8,
each phase branch consists of a resistance \((R_s/R_r')\), a leakage inductance \((L_{ls}/L_{lr}')\) and a magnetizing inductance \((L_{ms}/L_{mr}')\). Flux on each stator or rotor circuit is split into leakage and magnetizing parts, and only the latter part enters the magnetic coupling field. \(L_m\) is the mutual inductance which is equal to \(1.5L_{ms}\). However, this model does not consider core losses. On the other hand, the per-phase equivalent circuit model is much simpler. \(r_{c\_ph}\) is the per-phase equivalent core-loss resistance and can be used for core loss estimation. However, this model only works for sinusoidal-fed stationary condition, and thus is not applicable for design of advanced vector controllers involving dynamic responses and high-frequency components.

Fig. 8. The classical dynamic three-phase IM model ignoring core loss

The proposed model is shown in Fig. 10. It has a similar structure as Fig. 8, which is also a dynamic three-phase model with resistance, leakage inductance and magnetizing inductance in
each phase branch. However, the stator magnetizing branches are modified by three virtual
resistors, $R_c$, which are highlighted in red in Fig. 10. $R_c$ is in parallel with $L_{ms}$ in a similar manner
as in the per-phase equivalent circuit. However, the later derivation will show that $R_c$ and $r_{c,ph}$ are
not the same. Due to the injection of $R_c$, the stator phase currents are split into two parts: one for
flux linkage generation via $L_{ms}$ and the other one for core loss dissipation via $R_c$. Note that the
rest of the stator circuit and the entire rotor circuit in Fig. 10 are the same as Fig. 9, which makes
the $qd0$-frame manipulation of the proposed model convenient by referring to the similar process
as in the conventional dynamic three-phase model.

![Diagram](image)

Fig. 10. The proposed IM model considering core loss

### 3.1.1 Derivation of the $qd0$-frame Forms of the Proposed Model and Loss Expressions

Taking phase $a$ as an example, the phase voltage ($v_{as}$), current ($i_{as}$) and flux ($\lambda_{as}$) can be
calculated based on Fig. 10,

\[
v_{as} = R_c i_{as} + p \lambda_{as},
\]  

\[
i_{as} = \dot{i}_{as} + \frac{L_{ms}}{R_c} p \dot{i}_{as},
\]  

\[
\lambda_{as} = \lambda_{as} + \frac{L_{ms}}{R_c} \dot{i}_{as}.
\]
\[
\lambda_{as} = \left( L_{as} \dot{i}_{as} + L_{ms} \dot{i}_{ms} \right) - 0.5L_{ms} \dot{i}_{bs} - 0.5L_{ms} \dot{i}_{cs}
\]
\[
+ L_{ms} \cos \theta \dot{i}_{as} + L_{ms} \cos \left( \theta + \frac{2\pi}{3} \right) \dot{i}_{bs} + L_{ms} \cos \left( \theta - \frac{2\pi}{3} \right) \dot{i}_{cr},
\]
\[\text{where } p \text{ is the derivative operator, } \theta \text{ is the rotor electrical angle. The other variables in (3.1) to (3.3) are illustrated in Fig. 10. Substituting } i_{as} \text{ in (3.3) using (3.2) leads to a flux expression in terms of only magnetizing currents (} i_{abcs} \text{ and } i_{abcr}'. \text{ Note that the stator magnetizing currents (the currents flowing through } L_{ms} \text{) are changed from } i_{abcs} \text{ in Fig. 8 to } \hat{i}_{abcs} \text{ in Fig. 10. Applying the same analysis to phase } b \text{ and phase } c \text{ as well as to phases on the rotor side, it leads to the voltage, current and flux relationships of the three-phase system in matrix forms,}
\]
\[
\begin{pmatrix}
\dot{v}_{abcs} \\
\dot{v}_{abcr}'
\end{pmatrix}
= \begin{pmatrix}
R_s & 0 \\
0 & R_r'
\end{pmatrix}
\begin{pmatrix}
i_{abcs} \\
i_{abcr}'
\end{pmatrix}
+ p
\begin{pmatrix}
\dot{\lambda}_{abcs} \\
\dot{\lambda}_{abcr}'
\end{pmatrix},
\]
\[\text{(3.4)}\]
\[
\begin{pmatrix}
i_{abcs} \\
i_{abcr}'
\end{pmatrix}
= \begin{pmatrix}
\hat{i}_{abcs} \\
\hat{i}_{abcr}'
\end{pmatrix}
+ \frac{L_{ms}}{R_c}
\begin{pmatrix}
p \hat{i}_{abcs}
\end{pmatrix},
\]
\[\text{(3.5)}\]
\[
\begin{pmatrix}
\lambda_{abcs} \\
\lambda_{abcr}'
\end{pmatrix}
= \begin{pmatrix}
L_{ss} & L_{sr}' \\
L_{rs}' & L_{rr}'
\end{pmatrix}
\begin{pmatrix}
\hat{i}_{abcs} \\
\hat{i}_{abcr}'
\end{pmatrix}
+ \frac{L_{ls} L_{ms}}{R_c}
\begin{pmatrix}
p \hat{i}_{abcs}
\end{pmatrix},
\]
\[\text{(3.6)}\]
Here, bold font represents matrix variables. \( F_{abcs} = [F_{ax} F_{bx} F_{cx}]^T \), where \( F \) can represent voltage, current or flux while the subscript \( x \) can be \( s \) or \( r \) to represent stator or rotor components, respectively. The superscript \( T \) means transpose of a matrix. The matrixes \( R_s, R_r', L_{ss}, L_{sr}', L_{rs}' \) and \( L_{rr}' \) are
\[
R_s = \begin{bmatrix}
R_s & 0 & 0 \\
0 & R_s & 0 \\
0 & 0 & R_s
\end{bmatrix},
\]
\[\text{(3.7)}\]
Transforming both sides of (3.4)–(3.6) into an arbitrary $qd_0$-frame of frequency $\omega$ using the transformation matrix $K$, where $\hat{F}_{qd_0} = F_{abc}$ and $\hat{i}_{qd_0} = K \cdot \hat{i}_{abc}, F_{qd_0x} = [F_{qs} F_{ds} F_{os}]^T$, $\hat{i}_{qd_0x} = [\hat{i}_{qs} \hat{i}_{ds} \hat{i}_{os}]^T$, and 

$$K = \frac{2}{3} \begin{bmatrix}
\cos \theta \cos \left( \theta - \frac{2\pi}{3} \right) \\
\sin \theta \sin \left( \theta - \frac{2\pi}{3} \right) \\
\frac{1}{2} \end{bmatrix} \begin{bmatrix}
\cos \theta \cos \left( \theta + \frac{2\pi}{3} \right) \\
\sin \theta \sin \left( \theta + \frac{2\pi}{3} \right) \\
\frac{1}{2} \end{bmatrix}, \quad \omega = \frac{d\theta}{dt}.$$
Then, the voltage, current and flux in the $qd0$-frame are obtained, which are referred as \textit{machine structural equations}

\begin{align*}
v_{qs} &= R \dot{i}_{qs} + \omega \lambda_{ds} + p \lambda_{qs}, \quad (3.13) \\
v_{ds} &= R \dot{i}_{ds} - \omega \lambda_{qs} + p \lambda_{ds}, \quad (3.14) \\
v_{qr} &= R \dot{i}_{qr} + (\omega - \omega_r) \lambda_{dr} + p \lambda_{qr}, \quad (3.15) \\
v_{dr} &= R \dot{i}_{dr} - (\omega - \omega_r) \lambda_{qr} + p \lambda_{dr}, \quad (3.16) \\
i_{qs} &= \dot{i}_{qs} + \frac{L_{ms}}{R_c} \omega \dot{i}_{ds} + \frac{L_{ms}}{R_c} p \dot{i}_{qs}, \quad (3.17) \\
i_{ds} &= \dot{i}_{ds} - \frac{L_{ms}}{R_c} \omega \dot{i}_{qs} + \frac{L_{ms}}{R_c} p \dot{i}_{ds}, \quad (3.18) \\
\lambda_{qs} &= L_n i_{qs} + L_m (\dot{i}_{qs} + i_{qr}), \quad (3.19) \\
\lambda_{ds} &= L_n i_{ds} + L_m (\dot{i}_{ds} + i_{dr}), \quad (3.20) \\
\lambda_{qr} &= L_r i_{qr} + L_m (\dot{i}_{qs} + i_{qr}), \quad (3.21) \\
\lambda_{dr} &= L_r i_{dr} + L_m (\dot{i}_{ds} + i_{dr}), \quad (3.22) \\
v_{0s} &= R \dot{i}_{0s} + p \lambda_{0s}, \quad (3.23) \\
v_{0r} &= R \dot{i}_{0r} + p \lambda_{0r}, \quad (3.24) \\
i_{0s} &= \dot{i}_{0s} + \frac{L_{ms}}{R_c} p \dot{i}_{0s}, \quad (3.25) \\
\lambda_{0s} &= L_n i_{0s}, \quad (3.26) \\
\lambda_{0r} &= L_n i_{0r}, \quad (3.27) \\
\omega_r &= \frac{P}{2} \omega_{rm}. \quad (3.28)
\end{align*}
The induced electromagnetic torque, $T_e$, calculated in the qd0-frame is

$$T_e = -\frac{pW_m}{\omega_{rm}} = \frac{3P}{4}\left(\lambda_{qr}'i_{dr}' - \lambda_{dr}'i_{qr}'\right). \tag{3.29}$$

These equations suggest the qd0-frame version of the proposed core-loss model as shown in Fig. 11. The impedance branches, $Z_q$ and $Z_d$, are created only to satisfy the Kirchhoff’s current law (KCL) at nodes X and Y in Fig. 11. This qd0-frame structure with lumped impedances was purposely provided following the custom that passive components tend to be used in IM models to deal with core loss. However, a modified qd0-frame version of the proposed core-loss model is shown in Fig. 12, which is based on (3.17) and (3.18), and assume

$$i_{c_{-q1}} = \frac{L_{ms}}{R_c} \hat{\lambda}_{ds}, \tag{3.30}$$

$$i_{c_{-q2}} = \frac{L_{ms}}{R_c} p\left(\hat{i}_{qs}\right), \tag{3.31}$$

$$i_{c_{-d1}} = \frac{L_{ms}}{R_c} \hat{\lambda}_{qs}, \tag{3.32}$$

$$i_{c_{-d2}} = \frac{L_{ms}}{R_c} p\left(\hat{i}_{ds}\right), \tag{3.33}$$

where $i_{c_{-q1}}, i_{c_{-q2}}, i_{c_{-d1}}, i_{c_{-d2}}$ are variables used for illustration convenience.

It is explicitly indicated in Fig. 12 that the impedance branches are essentially consisted of four current-controlled current sources. This feature makes the proposed core-loss model very different from the models published in the literature, which use passive core-loss resistors in qd0-frame [15]–[18]. Equations (3.30)–(3.33) show that the four current sources are functions of $R_c$, while later characterization tests will show that $R_c$ is a function of operating frequency and flux. Therefore, it indicates that $Z_q$ and $Z_d$ or the current sources are changing with frequency.
and flux, and their values may need to adjust for different conditions before the core-loss model is properly used. Note that, in Fig. 12, the core loss is the power loss consumed in the four current/(flux)-controlled voltage sources and the four current-controlled current sources, rather than losses only in the magnetizing branches.

Fig. 11. qd0-frame version 1 of the proposed core-loss model

The torque and speed are related through

\[ T_e - T_L = J \cdot p(\omega_m), \]  

(3.34)

where \( T_L \) is the load torque, \( J \) is the machine inertia, \( T_e \) and \( T_L \) are equal at the steady state.

Assuming a balanced machine and thus ignoring the \( \theta \)-axis circuit, copper loss can be calculated by the Joule losses on the stator and rotor resistors

\[ P_{Cu} = \frac{3}{2} \left[ R_s (i_{qs}^2 + i_{ds}^2) + R_r (i_{qr}^2 + i_{dr}^2) \right], \]  

(3.35)
Fig. 12. qd0-frame version 2 of the proposed core-loss model

where $P_{Cu}$ represents the copper power loss. The 3/2 coefficient is used to compensate the 2/3 factor in $K$. By considering the energy flowing in the machine in instantaneous forms,

$$pW_e + pW_m = pW_{loss} + pW_{st},$$ (3.36)

$$pW_{loss} = P_{Cu} + P_{core},$$ (3.37)

$$pW_{st} = pW_{ss} + pW_{sm},$$ (3.38)

where $W_e, W_m, W_{loss}, W_{st}$ are the electrical input energy, mechanical input energy, dissipated or lost energy, and stored energy of the electro-mechanical field, respectively. $W_{ss}$ and $W_{sm}$ are the energy stored in the leakage flux and magnetizing flux, respectively. $P_{core}$ is the core loss. $p$ is again the derivative operator that converts energy to power. Based on (3.36)–(3.38) and Fig. 11,

$$P_{core} + pW_{sm} - pW_m = pW_e - P_{Cu} - pW_{ss}$$

$$= \frac{3}{2} \left( u_{qs}i_{qs} + u_{qr}i_{qr} + u_{ds}i_{ds} + u_{dr}i_{dr} \right),$$ (3.39)

32
where, \(u_{qs}, u_{ds}, u_{qr}', u_{dr}'\) are shown in Fig. 11 (in blue). Expressing \(u_{qs}, u_{ds}, u_{qr}', u_{dr}'\) as functions of currents and fluxes, and then re-arranging the resultant terms,

\[
P_{\text{core}} + pW_{\text{sm}} - pW_m = \frac{3}{2}\left[ i_{qs} \cdot \omega \lambda_{ds} + i_{qr}' \omega \lambda_{dr}' + \left( i_{qs} - \hat{i}_{qs} \right) \left( i_{qs} + i_{qr}' \right) - i_{dr}' \cdot \omega \lambda_{qr}' - i_{dr} \cdot \omega \lambda_{qr} - \left( i_{dr} - \hat{i}_{dr} \right) \left( i_{dr} + i_{dr}' \right) \right] L_m \cdot p \left( \hat{i}_{qs} + i_{qr}' \right) \]

\[
+ \frac{3}{2}\left[ \left( \hat{i}_{qs} + i_{qr}' \right) L_m \cdot p \left( \hat{i}_{qs} + i_{qr}' \right) + \left( \hat{i}_{dr} + i_{dr}' \right) L_m \cdot p \left( \hat{i}_{dr} + i_{dr}' \right) \right] - \frac{3}{2} \omega_e \left( \lambda_{dr}' - \lambda_{qr}' \right) \left( i_{qr}' - \lambda_{qr}' \right)
\]

(3.40)

Therefore, \(P_{\text{core}}, pW_{\text{sm}}\) and \(pW_m\) have the expressions as the three terms shown on the right side of (3.40) sequentially. Specifically, the expression of core loss is

\[
P_{\text{core}} = \frac{3}{2}\left[ i_{qs} \cdot \omega \lambda_{ds} + i_{qr}' \omega \lambda_{dr}' + \left( i_{qs} - \hat{i}_{qs} \right) \left( i_{qs} + i_{qr}' \right) - i_{dr}' \cdot \omega \lambda_{qr}' - i_{dr} \cdot \omega \lambda_{qr} - \left( i_{dr} - \hat{i}_{dr} \right) \left( i_{dr} + i_{dr}' \right) \right] L_m \cdot p \left( \hat{i}_{qs} + i_{qr}' \right) \]

(3.41)

Equation (3.41) is the general expression of \(P_{\text{core}}\) that works for any arbitrary \(qd0\)-frame. In the synchronous \(qd0\)-frame where \(i_{qd0}, i_{qd0}'\) and \(\hat{i}_{ab}\) are constant at steady state, the derivative terms in (3.41) are zero and \(P_{\text{core}}\) can be simplified to

\[
P_{\text{core,syn}} = \frac{3}{2}\left( i_{qs} \cdot \omega \lambda_{ds} + i_{qr}' \omega \lambda_{dr}' - i_{qs} \cdot \omega \lambda_{qr} - i_{dr}' \cdot \omega \lambda_{qr} \right)
\]

\[
= \frac{3}{2} \omega_e \frac{L_m L_m}{R_e} \left( \hat{i}_{ds}^2 + \hat{i}_{dr}'^2 + \hat{i}_{qr}'^2 + \hat{i}_{qs}^2 \right)
\]

(3.42)

where \(P_{\text{core,syn}}\) and \(\omega_e\) are the synchronous \(qd0\)-frame core loss and frequency, respectively. Moreover,

\[
T_e = -\frac{pW_m}{\omega_m} = \frac{3}{4} \left( \lambda_{qr}' - \lambda_{dr}' \right) \left( i_{dr}' - \lambda_{qr}' \right)
\]

(3.43)
3.1.2 Parameters of the Proposed Core-Loss Model

As shown in Fig. 10, there are six independent parameters in the proposed core-loss model: \( R_s, L_{ls}, R_r', L_{lr}, R_c \) and \( L_m \) (\( L_{sr}' = L_{rs}' = L_m = 1.5 L_{ms} \)). The determination of these parameters can follow the IEEE Standard 112 [42]. Although the machine characterization tests, namely the DC test, locked-rotor test and no-load test, are designed to extract the parameters of the per-phase equivalent circuit model, the derivation below will show that the parameters in the proposed model are the same as the parameters in the per-phase equivalent circuit, except \( r_{c,ph} \) which is equal to \( 1.5 R_c \) instead of \( R_c \).

Starting with the phasor forms of (3.13) and (3.15) by replacing \( p \) by \( j(\omega_e - \omega) \), then using the \( qd0 \)-frame phasor property \( \tilde{F}_{qs} = j\tilde{F}_{qs} \), where the tilde sign means the quantity in phasor form. The phasor forms of (3.13) and (3.15) are,

\[
\begin{align*}
\tilde{V}_{qs} &= R_s \tilde{i}_{qs} + j\omega_e \tilde{\lambda}_{qs}, \\
\tilde{V}_{qr}' &= R_r' \tilde{i}_{qr}' + j(\omega_e - \omega) \tilde{\lambda}_{qr}'.
\end{align*}
\]

Applying the phasor forms of (3.19) and (3.21) to (3.44) and (3.45), and using another phasor property, \( \tilde{F}_{qs} = \tilde{F}_{as} \), to current terms by selecting the initial phases of \( i_{as}, i_{ar}' \), \( \hat{i}_{as} \) to be zero,

\[
\begin{align*}
\tilde{V}_{as} &= (R_s + j\omega_e L_{ls}) \tilde{i}_{as} + j\omega_e L_m \left( \tilde{i}_{as} + \tilde{i}_{ar}' \right), \\
\tilde{V}_{ar}' &= \left( R_r' + j\omega_e L_{lr} \right) \tilde{i}_{ar}' + j\omega_e L_m \left( \tilde{i}_{as} + \tilde{i}_{ar}' \right).
\end{align*}
\]

where \( s \) is the machine slip. Equations (3.46) and (3.47) represent the steady-state per-phase version of the proposed model that is shown in Fig. 13. The resistance branch, \( r_{c,ph} \), is created to satisfy the KCL law at node A in Fig. 13. The same resistance symbol, \( r_{c,ph} \), is used in Fig. 13 as
in Fig. 9, since it is found that Fig. 13 and Fig. 9 are essentially the same, considering $V_{ar}$ is zero in a squirrel-cage IM. Based on Fig. 13, $r_{c,ph}$ can be calculated by

$$r_{c,ph} = \frac{j\omega L_m}{\frac{\tilde{i}_s}{\tilde{i}_s - \tilde{i}_{ar}}}. \quad (3.48)$$

Using the phasor form of (3.17) and applying the previous phasor properties to change the $q$-axis and $d$-axis phasors to the $a$-axis, equation (3.48) is changed to

$$r_{c,ph} = \frac{3}{2} R_c \frac{\tilde{i}_s + \tilde{i}_{ar}}{\tilde{i}_s}. \quad (3.49)$$

In the no-load machine characterization test, the induced rotor current is negligible. Thus, (3.49) can be simplified to

$$R_c \approx \frac{2}{3} r_{c,ph} \text{ (no load)}. \quad (3.50)$$

Therefore, the value of $R_c$ in the proposed core-loss model can be obtained based on the $r_{c,ph}$ value from the no-load characterization test. Note that, similar to $r_{c,ph}$, $R_c$ changes with frequency and flux levels.

3.1.3 Simulation Verification of the Proposed Core-Loss Model

To verify the proposed model, especially the analytical expression of $P_{core}$, a Simulink model is built based on the core-loss model. A 4-pole 1.5 HP IM is used in the simulation. The consistency of the model is checked through

$$P_{in \ (elec)} - P_{out \ (mech)} = P_{Cu} + P_{core} + pW_n + P_{mech}, \quad (3.51)$$

where $P_{mech}$ is the mechanical loss. The instantaneous stored energy can be averaged out if applying an average window longer than the fundamental period of the power source on both
sides of (3.51) ($P_{W_s}$=0). The stray loss is ignored in the simulation verification since the focus is on the core loss expression. $P_{in}$ and $P_{out}$ are calculated by

$$P_{in} = \frac{3}{2} \left( v_{qs}i_{qs} + v_{ds}i_{ds} \right),$$

$$P_{out} = T_e \cdot \omega_{rm} - P_{mech}. \quad (3.53)$$

The simulation verification is carried out at both line-fed and inverter-fed conditions. A linearly increasing $T_L$ followed by several step-down $T_L$ is applied. In the line-fed condition, the line-to-line input voltage is 200 V. In the inverter-fed condition, modulation index ($MI$) is set to 1 and $V_{dc}$ is 308 V. ($P_{Cu} + P_{core}$) and ($P_{in} - P_{out} - P_{mech}$) are compared in the simulation. The result of the line-fed condition is shown in Fig. 14. It is found that ($P_{Cu} + P_{core}$) matches ($P_{in} - P_{out} - P_{mech}$) excellently. The results of the inverter-fed conditions are shown in Fig. 15 and Fig. 16. It is seen that the core-loss model works excellently at different speeds and torques, and it can be equally applied to different reference frames.

![Diagram](image1.png)

**Fig. 13.** The steady-state per-phase version of the proposed core-loss model

![Graph](image2.png)

**Fig. 14.** Comparison of ($P_{in} - P_{out} - P_{mech}$) and ($P_{Cu} + P_{core}$) in the line-fed condition
Fig. 15. Comparison of \( P_{\text{in}}-P_{\text{out}}-P_{\text{mech}} \) and \( P_{\text{Cu}}+P_{\text{core}} \) in the inverter-fed condition at 1735 RPM in the synchronous frame (a) and stationary frame (b).

Fig. 16. Comparison of \( P_{\text{in}}-P_{\text{out}}-P_{\text{mech}} \) and \( P_{\text{Cu}}+P_{\text{core}} \) in the inverter-fed condition at 600 RPM in the synchronous frame (a) and stationary frame (b).

3.1.4 Experimental Validation of the Proposed Core-Loss Model

Further validation of the proposed core-loss model, which compares the simulated and the experimentally measured copper and core losses, is performed. The model validity is examined at no-load and loaded conditions with different frequencies and flux levels. In the no-load validation, both sinusoidal-fed and inverter-fed conditions are considered. In the loaded verification, only pure sinusoidal excitation is used, since the definition of PWM core loss and PWM stray loss are unclear in literature. Thus, they are difficult to separate from the real measurements. The comprehensive experimental validation shows the correctness of the proposed core-loss model in various excitation, frequency, flux and load conditions, which is the prerequisite of later model-based control design using the proposed core-loss model.
3.1.4.1 Experimental Validation at No-Load Conditions

Three IMs (1.5HP, 3HP and 10HP) are tested in no-load conditions to show the scalability of the proposed model. Their basic information are provided in TABLE II. In the no-load validation, machine stray loss and rotor copper loss are properly ignored. The mechanical loss is determined experimentally following IEEE Standard 112 [42]. Basically, several no-load tests are performed under a certain speed with different voltage excitations decreased from rated value. Then, the power \((P_{\text{core}}+P_{\text{mech}})\) versus voltage squared are linearly curve-fitted. The intersection of the curve at y-axis gives \(P_{\text{mech}}\) for that speed considering \(P_{\text{core}}\) is zero at zero voltage. An example of determining \(P_{\text{mech}}\) at rated speed and flux for the 1.5HP IM is shown in Fig. 17. Therefore, \(P_{\text{Cu}}\) and \(P_{\text{core}}\) are determined experimentally through

\[
P_{\text{Cu}} = P_{\text{s,Cu}} = 3I_s^2 R_s,
\]

\[
P_{\text{core}} = P_{\text{in}} - P_{\text{Cu}} - P_{\text{mech}},
\]

where \(I_s\) is the RMS value of the stator current. \(P_{\text{s,Cu}}\) is the stator copper loss. \(I_s\) and \(P_{\text{in}}\) are the measured values. On the other hand, the previous Simulink model is excited at the same conditions as in the experiments to get the simulated machine copper and core losses, where the experimentally decided mechanical loss is added in the simulation to provide the additional load torque.

**TABLE II. INFORMATION OF THE TESTED MACHINES IN NO-LOAD VALIDATION**

<table>
<thead>
<tr>
<th>Machine Type</th>
<th>Dayton 6VPE6</th>
<th>Dayton 6VPE8</th>
<th>Dayton 2MXV4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>1.5HP</td>
<td>3HP</td>
<td>10HP</td>
</tr>
<tr>
<td>Voltage</td>
<td>230 V</td>
<td>230 V</td>
<td>230 V</td>
</tr>
<tr>
<td>Current</td>
<td>4 A</td>
<td>8.1 A</td>
<td>25.8 A</td>
</tr>
<tr>
<td>Pole Number</td>
<td>4</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Speed</td>
<td>1735 RPM</td>
<td>1735 RPM</td>
<td>1755 RPM</td>
</tr>
<tr>
<td>Torque</td>
<td>6.1 N·m</td>
<td>12.3 N·m</td>
<td>41 N·m</td>
</tr>
</tbody>
</table>
Fig. 17. Determination of mechanical loss for the 1.5HP IM at rated speed and flux

First, the model is examined in the line-fed condition which applies sinusoidal excitation having rated frequency and a little weakened flux (the input voltage is 208V instead of 230V). The experimental setup is shown in Fig. 18. The IMs are energized by grid via a VARIable AC Source (VARIAC). A Kollmorgen servomotor AKM65L is used to lock the motors in the lock-rotor test. Yokogawa WT1800 power analyzer is used to measure the machine input voltage, current and power. The experimental and simulated machine losses in the line-fed condition are compared in Fig. 19. Results show excellent loss estimation using the proposed core-loss model.

![Fig. 18. Line-fed test and characterization setup: (a) Block diagram; (b) Photo](image-url)
Then, the proposed model is verified at a decreased speed with rated and weakened fluxes. The Pacific 320AMX AC power supply is used to test the previous 1.5 HP motor, which can give sinusoidal output with independent settings on voltage magnitude and frequency. In this test, the speed of the machine is fixed at 1200 RPM and V/f ratio is decreased from rated value until the speed drops significantly. The comparison of simulated and experimental copper and core losses are shown in Fig. 20, where the proposed model shows excellent loss estimation.

Fig. 19. Comparison of simulated and experimental machine losses in the line-fed situation: (a) Copper loss; (b) Core loss

Fig. 20. Comparison of the simulated and experimental machine losses in the flux-weakening region: (a) Copper loss; (b) Core loss
In the inverter-fed situation, each of the three IMs in TABLE II are tested under seven different speeds while the V/f ratio is kept constant. The experimental setup for the inverter-fed tests is shown in Fig. 21. dSPACE DS1104 is used to provide real-time PWM switching signals from a V/f controller built in MATLAB/Simulink. To alleviate the impacts of harmonics, low pass filters (LPFs) are used at the input of the machines. The simulated and experimental machine losses in the inverter-fed conditions are compared in Fig. 22, and the estimation errors are shown in Fig. 23. Note that the results of the 3HP machine at 800 RPM and 600 RPM, and the 10HP machine at 600 RPM are not obtained experimentally due to the stall of machine. It is observed that the proposed model can estimate the machine total loss with higher than 93% accuracy for all the tested conditions on the three machines. The estimation errors for many conditions are less than 2%. Moreover, even without LPFs, the proposed model can still provide better than 80% estimation accuracy of machine losses under the present level of harmonics (10kHz PWM switching frequency).

Fig. 21. Experiment setup for inverter-fed tests: (a) Block diagram; (b) Photo
Fig. 22. Comparison of the simulated and experimental machine losses at inverter-fed conditions:
(a) Copper loss; (b) Core loss

Fig. 23. Power loss estimation error of the proposed model at inverter-fed conditions
3.1.4.2 Experimental Validation at Loaded Conditions

The proposed core-loss model is further validated with load at different frequencies and flux levels, where the simulated and experimental copper and core losses are compared again. The 1.5HP motor is excited using the Pacific AC source. In the load validation, $P_{\text{mech}}$ and $P_{s,Cu}$ are measured as in the no-load validation. Moreover, based on the IEEE Standard 112, $P_{\text{core}}$ does not change with load. Therefore, $P_{\text{core}}$ in load validation will be the same as that calculated from the no-load validation at the same excitation. On the other hand, the stray loss, $P_{\text{stray}}$, and rotor copper loss, $P_{r,Cu}$, cannot be ignored in the load validation. $P_{r,Cu}$ is calculated from

$$P_{r,Cu} = s \cdot (P_{\text{in}} - P_{s,Cu} - P_{\text{core}}).$$

(3.56)

Then, $P_{\text{stray}}$ is calculated from

$$P_{\text{stray}} = (P_{\text{in}} - P_{\text{out}}) - (P_{s,Cu} + P_{\text{core}}^{x} + P_{\text{mech}}^{x} + P_{r,Cu}).$$

(3.57)

Here, the superscript $x$ means the variable is obtained from no-load tests. It is required that the determination ($R^{2}$) of the linear regression between $P_{\text{stray}}$ and $T_{L}^{2}$ needs to be larger than 0.9. An example of $P_{\text{stray}}$ at different $T_{L}$ is shown in Fig. 24. On the simulation side, the previous Simulink model is applied while an additional torque is added to emulate stray loss effects. It is noted that, similar to the no-load validation, the load validation here are intended to show the correctness of the proposed core-loss model itself. There is no intension to discuss whether the applied core-loss model is more accurate than the per-phase equivalent circuit model in loss estimation under steady-state sinusoidal-fed condition. Compared to the per-phase model, the superiority of the core-loss model is its capability for dynamic $qd0$-frame analysis and design.

The overview of the design of experiments in load tests is shown in Fig. 25 along with the applied V/f ratio, frequency and torque conditions. The manually set V/f ratio is used to roughly decide the flux level. The first type of experiments is the load validation of the core-loss
model at rated flux. In this experiment, the experimental and simulated machine copper and core losses are compared, respectively, at three different frequencies, and six load torques for each frequency. The second type of experiments is the validation of the core-loss model at non-zero load and weakened fluxes. This is especially important for loss minimization control using the proposed core-loss model, since properly adjusting flux level is key for loss minimization control. Therefore, the validity of the core-loss model at different flux levels needs to be ascertained.

![Graph showing stray loss vs. torque squared](image)

**Fig. 24. An example of determining stray loss**

<table>
<thead>
<tr>
<th>Experiment 1</th>
<th>Load Validation Under Rated Flux</th>
</tr>
</thead>
<tbody>
<tr>
<td>V/f: 2.21(rated value)</td>
<td>Frequency(Hz): 20, 40, 60</td>
</tr>
<tr>
<td>Torque(N·m): 5.9(100%) ... 1.5(25%)</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Experiment 2</th>
<th>Load Validation Under Weakened Flux</th>
</tr>
</thead>
<tbody>
<tr>
<td>V/f: 1.91(weakened value)</td>
<td>Frequency(Hz): 40</td>
</tr>
<tr>
<td>Torque(N·m): 5.9(100%) ... 1.5(25%)</td>
<td></td>
</tr>
</tbody>
</table>

**Fig. 25 Overview of the load validation tests**

The results for the first type of experiments are shown in Fig. 26–Fig. 28. In each frequency, $T_L$ decreases from about rated value to about 25% of rated $T_L$. It is clear from the figures that the core-loss model can accurately simulate machine copper and core losses at different frequencies and loads. The estimation error is *relatively* large at the highest $T_L$, since the
real machine could be partially saturated considering additional mechanical and stray losses. Thus, the machine parameters are no longer very accurate for the power loss estimation. Moreover, the voltage drop at the stator resistors increases with $T_L$, which will slightly lower the true flux although the V/f ratio is kept constant. This fact will also slightly change the model parameters from the ones obtained in the characterization tests, and thus contribute to the estimation error. Nevertheless, the estimation error of the power losses is within a few watts, and practical nonidealities, such as slight unbalanced phases, sensor and measurement inaccuracy, temperature effects, electromagnetic interferences, etc., could easily cause it.

On the other hand, it is seen that $P_{\text{core}}$ increases with frequency. In contrast, $P_{\text{Cu}}$ is not seriously affected by frequency, but it is significantly changed with $T_L$. Moreover, it is found that $P_{\text{Cu}}$ at 20Hz is higher than that at 40Hz and 60Hz when load torque is 5.87 N\(\cdot\)m. This is because the input voltage at 20Hz is much smaller than the other two cases (constant V/f ratio). Thus, a similar voltage drop on stator resistors will lead to larger flux drop at 20Hz. Then, insufficient flux will require more current to support the high torque, which generates more copper loss. The results also ascertain that $P_{\text{core}}$ gradually dominates the machine total loss in the low-load conditions. Thus, considering core loss in machine analysis and control design is especially important in the low-load and/or high-frequency conditions.

The results for the second type of experiments are shown in Fig. 29. It is seen that the estimated losses again match the experimental counterparts very well. Moreover, comparing Fig. 27–Fig. 29, it is seen that part of $P_{\text{core}}$ is relocated to $P_{\text{Cu}}$ when the flux decreases. This is the reason of possible efficiency enhancement from proper flux weakening.
Fig. 26. Comparison of experimental and simulated power losses at 20Hz and rated V/f ratio: (a) Copper loss; (b) Core loss

Fig. 27. Comparison of experimental and simulated power losses at 40Hz and rated V/f ratio: (a) Copper loss; (b) Core loss

Fig. 28. Comparison of experimental and simulated power losses at 60Hz and rated V/f ratio: (a) Copper loss; (b) Core loss
Fig. 29. Comparison of experimental and simulated power losses at 40Hz and weakened flux (V/f=1.91): (a) Copper loss; (b) Core loss.

3.1.5 Parameter Sensitivity and Adaptation of the Proposed Core-Loss Model

3.1.5.1 Parameter Sensitivity of the Proposed Core-Loss Model

For model-based analysis and design, the accuracy of the model parameters have important effects on the results. To study the effects of the parameter values on the simulated machine losses, a series of model parameter sensitivity tests are performed on the 1.5HP IM at three different speed and torque conditions as shown in TABLE III. In each simulation run, one of the six independent model parameters, $R_s$, $L_{ls}$, $R'_r$, $L_{lr}'$, $R_c$, $L_m$ or $L_{ms}$, is changed by -20%, -10%, 10%, 20% from their nominal values. The results for the change of losses compared to using the nominal parameters are shown in Fig. 30, and the legend of Fig. 30 is explained in TABLE III due to limited space in the figure.

<table>
<thead>
<tr>
<th>Condition</th>
<th>Speed</th>
<th>Torque</th>
<th>Line color + Marker</th>
</tr>
</thead>
<tbody>
<tr>
<td>CD1 (Condition 1)</td>
<td>1735 RPM</td>
<td>6 N·m</td>
<td>Purple + Dot</td>
</tr>
<tr>
<td>CD2 (Condition 2)</td>
<td>1735 RPM</td>
<td>1 N·m</td>
<td>Green + Cross</td>
</tr>
<tr>
<td>CD3 (Condition 3)</td>
<td>600 RPM</td>
<td>1 N·m</td>
<td>Orange + Square</td>
</tr>
</tbody>
</table>

It is found that, first, $R_c$ mainly affects $P_{core}$; $R_s$ and $L_m(L_{ms})$ mainly affect $P_{Cu}$; $L_{ls}$ and $L_{lr}'$ have minor effects on $P_{core}$ and little effect on $P_{core}$; $R'_r$ has minor effect on $P_{Cu}$ and almost no
effect on $P_{\text{core}}$. These effects increase as the parameter error increases. Second, the increase of $T_L$ mainly increases $P_{\text{Cu}}$. Thus, $\Delta P_{\text{Cu}}$ in percentage is less sensitive to parameters’ variation in high-torque condition due to the increased value of $P_{\text{Cu}}$. Similarly, the increase of speed increases $P_{\text{core}}$. Thus, $\Delta P_{\text{core}}$ in percentage is less sensitive in the high-speed condition. Third, due to the reasons in the previous point, the estimation deviation can exceed 20% in the sensitive low-torque low-speed condition, while the estimation deviation is less than 10% in the relatively insensitive high-torque high-speed condition for the same changing degree of the model parameters. Fourth, the effects of the model parameters on the power loss estimation are monotonous and almost linear before saturation. Thus, if the increase of a certain parameter increases the power loss estimation, then the decrease of the same parameter will decrease the same type of the power loss estimation in almost the same degree.
3.1.5.2 Parameter Adaptation of the Proposed Core-Loss Model

As shown in the previous section that the value of model parameters will affect the proposed model’s accuracy and the estimated machine losses. Therefore, it is important to accurately characterize the model parameters, or adapt the model parameters with the change of frequency and flux level. The adaptation of the model parameters with different frequencies and flux levels are shown next, where the Pacific AC source is used to adjust the input sinusoidal excitation as needed.

First, different frequencies are applied in the characterization tests while rated V/f ratio is used. The input voltage increases along with the frequency until rated value. The extracted machine parameters are shown in Fig. 31. It is seen that $R_s$ is constant from the DC test. $L_{ds}$ and $L_{dr}'$ are also almost constant with respect to frequency. Moreover, $R_r'$ slightly increases with frequency in general, since the practical stray loss, which is assumed to be zero in the no-load test, is included in the $R_r'$ determination. $L_m$ slightly decreases with frequency. It is important to note that $R_c$ significantly increases with frequency. Since $R_c$ considerably affects the estimated
It is expected that $P_{\text{core}}$ is a strong function of frequency. When the V/f ratio is constant, the $P_{\text{core}}$ at different frequencies are shown in Fig. 32. As expected, $P_{\text{core}}$ increases with frequency.

![Fig. 31. Parameters of the proposed core-loss model at different frequencies](image1)

![Fig. 32. Core loss at different frequencies](image2)

Then, the operating frequency is fixed at 40Hz. The input voltage is decreased from rated value with the decrease of the V/f ratio until the input current starts to increase. The extracted machine parameters at different V/f ratios are shown in Fig. 33. It is seen that $R_s$ is still constant while $L_{ls}$, $L_{lr}'$ and $R_{r}'$ are also almost independent of the V/f ratio. $L_m$ and $R_c$ tend to have a parabolic shape. On the other hand, $P_{\text{core}}$ at different flux levels are shown in Fig. 34. It is seen that $P_{\text{core}}$ increases with the V/f ratio, or machine input voltage when the operating frequency is fixed. Note that higher-order harmonics in the PWM excitation will also modify the model parameters beyond the values that are determined by the fundamental excitation component. Therefore, the model parameters obtained from the sinusoidal-fed characterization tests could be
inaccurate for the corresponding inverter-fed conditions depending on the harmonics’ levels. Better accuracy of the core-loss model is expected with the inverter-fed characterization tests, which unfortunately are not available in any literature at the moment.

Fig. 33. Parameters of the proposed core-loss model at different flux levels

![Core Loss](image1)

Fig. 34. Core loss at different flux levels

### 3.2 Modeling of Mechanical and Stray Losses

In addition to copper and core losses, mechanical and stray losses are also important for estimating machine total loss and efficiency, and for further designing loss minimization control or efficiency enhancement control. A series of comprehensive experimental tests have been performed to better understand mechanical and stray losses as well as their change with respect to machine load torque, speed and flux.
3.2.1. Description of the Experimental Tests

As shown in the previous section as well as in Fig. 17 and Fig. 24, $P_{\text{mech}}$ and $P_{\text{stray}}$ are measured experimentally following IEEE Standard 112. To examine the change of $P_{\text{mech}}$ and $P_{\text{stray}}$ with respect to speed and flux, $P_{\text{mech}}$ and $P_{\text{stray}}$ are measured at 1) different speeds and rated flux; and 2) different weakened flux and rated speed. $P_{\text{mech}}$ and $P_{\text{stray}}$ are first tested under the line-fed condition, where the Pacific AC power supply is used to set the speed and voltage (flux). A general-purpose 60Hz 1.5HP IM is used as the subject whose rated speed is 1800RPM. The V/f ratio set by the AC power supply is used to determine the machine flux. For certain speed and flux level, a series of $T_L$ (25%, 37.5%, 50%, 62.5%, 75%, 87.5%, 100%, 112.5%, 125% of rated value) are used in determining $P_{\text{stray}}$ until the torque cannot be supported by the IM due to the weakened speed or flux. In each case, the linearity between $P_{\text{stray}}$ and $T_L^2$ needs to satisfy the IEEE caliber.

Then, $P_{\text{mech}}$ and $P_{\text{stray}}$ are measured in inverter-fed situation using the setup shown in Fig. 21. However, the IEEE Standard 112 only instructs the sinusoidal-fed testing of $P_{\text{mech}}$ and $P_{\text{stray}}$, and no other standards or protocols are available for the inverter-fed test. Therefore, the line-fed testing procedure are referred in our inverter-fed tests of $P_{\text{mech}}$ and $P_{\text{stray}}$. Specifically, $P_{\text{mech}}$ and $P_{\text{core}}$ obtained in the no-load tests are used in the load conditions assuming $P_{\text{mech}}$ and $P_{\text{core}}$ (both fundamental and harmonics parts) are not changing with $T_L$. An open-loop V/f controller is built in Simulink and loaded to the dSPACE platform, which decides the fundamental voltage’s frequency and magnitudes through

$$\omega_e = \frac{P \pi \cdot \text{Spd}^*}{60}, \quad (3.58)$$

$$V_{l_{\text{rms}}} = \sqrt{\frac{3}{2}} \cdot \frac{MI}{2} \cdot V_{\text{dc}} \cdot \frac{V_{f^*} \cdot \text{Spd}^*}{V_{f_{\text{rated}}} \cdot \text{Spd}_{\text{rated}}}. \quad (3.59)$$
Here, $Spd^*$ is the speed command of the V/f controller and the asterisk means commands in the controllers, $Vf^*$ is the V/f ratio command, $MI$ is the modulation index and $V_{dc}$ is the DC link voltage of the inverter. $V_{ll\text{,rms}}$ is the RMS value of the line-to-line fundamental voltage. In the inverter-fed tests, the fundamental voltages are set to be the same as those in the corresponding line-fed tests.

3.2.2. Experimental Results and Discussion

The change of $P_{\text{mech}}$ with respect to speed in the line-fed and inverter-fed tests are shown in Fig. 35. It is seen that $P_{\text{mech}}$ is almost a linear function of speed. Moreover, the excitation harmonics barely show impacts on $P_{\text{mech}}$. On the other hand, it is found that $P_{\text{mech}}$ does not change with flux when decreasing $Vf^*$ at a certain speed.

The change of $P_{\text{stray}}$ with respect to speed in the line-fed and inverter-fed tests are shown in Fig. 36 and Fig. 37, respectively. It is observed that $P_{\text{stray}}$ increases linearly with $T_L^2$ in both line-fed and inverter-fed conditions. In Fig. 36 or Fig. 37, the slope of the lines for different speeds are similar, but the slope of the lines are different between the line-fed condition and the inverter-fed condition. Moreover, both figures show that $P_{\text{stray}}$ increases with speed for a certain $T_L$. However, the quantitative relationship between $P_{\text{stray}}$ and speed is not found. Comparing Fig. 36 and Fig. 37, $P_{\text{stray}}$ shows larger value in the line-fed condition than that of the inverter-fed counterpart, but no explicit impact of harmonics on $P_{\text{stray}}$ is found.

The change of $P_{\text{stray}}$ with respected to $Vf^*$, or stator flux, in the line-fed tests are shown in Fig. 38. It is seen that $P_{\text{stray}}$ shows very little differences at different flux levels except for rated flux. The relatively large departure of $P_{\text{stray}}$ at rated flux could be accidentally caused by the experimental imperfectness. On the other hand, as shown in (3.59), the actual fundamental voltage and flux of an inverter-fed IM are determined by $Vf^*$, $MI$ and $V_{dc}$ for a certain speed.
condition. At rated condition, \( Vf^* \) is 2.213 V·s, \( MI \) and \( V_{dc} \) are set to 0.9 and 417.3 V, respectively. When decreasing the flux, one of these three variables is decreased and the other two variables are kept at rated values. The change of \( P_{stray} \) at different flux, which is achieved through properly decreasing \( Vf^* \), \( MI \) or \( V_{dc} \), are shown in Fig. 39–Fig. 41, respectively. It is seen that \( P_{stray} \) shows divergent forms, as the slope of the curves increases with the decrease of flux (via the decrease of \( Vf^* \), \( MI \) or \( V_{dc} \)) in these figures. However, it is important to note that \( P_{stray} \) is barely affected by flux for \( T_L \) that is less than 50% of rated value. It will be shown in Chapter 4 that this is the major area of interest for loss minimization control of IM drives.

![Fig. 35. Mechanical loss at different speeds in line-fed and inverter-fed conditions](image1)

![Fig. 36. Stray loss at different speeds (fundamental frequencies) in the line-fed condition](image2)
Fig. 37. Stray loss at different speeds (fundamental frequencies) in the inverter-fed condition

Fig. 38. Stray loss at different V/f ratios (flux levels) in the line-fed condition

Fig. 39. Stray loss at different V/f ratios (flux levels) in the inverter-fed condition
3.3.3. Modeling of Mechanical Loss and Stray Loss

Based on the previous analysis,

- $P_{mech}$ and $P_{stray}$ are changing with speed;
- $P_{mech}$ is not affected by flux;
- $P_{stray}$ is also not affected by flux in the region that will be used for loss minimization control in Chapter 4, e.g. $T_L < 0.5 \ T_{L_{\text{rated}}}$.
\begin{itemize}
  \item $P_{\text{mech}}$ is not changing with $T_L$ (technically, $T_L$ will change speed and thus change $P_{\text{mech}}$ in an open-loop control without speed regulation, but this fact is included in $P_{\text{stray}}$);
  \item $P_{\text{stray}}$ is a linear function of $T_L^2$ and the slope of the linear function is different for different speeds.
\end{itemize}

Therefore, $P_{\text{mech}}$ can be modelled as a linear function of speed. Based on Fig. 35,

\begin{equation}
P_{\text{mech}} = 0.0557 \cdot \text{Spd} - 18.426. \tag{3.60}
\end{equation}

$P_{\text{mech}}$ can also be modelled using a look-up table, which has different constant values for different speeds. The look-up table that is corresponding to Fig. 35 is shown in TABLE IV.

On the other hand, $P_{\text{stray}}$ can be modelled as different linear functions with respect to $T_L^2$ in the form

\begin{equation}
P_{\text{stray}} = k_{\text{stray}1} \cdot T_L^2 + k_{\text{stray}2}, \tag{3.61}
\end{equation}

where the coefficients, $k_{\text{stray}1}$ and $k_{\text{stray}2}$, are the extracted from curve fitting of the experimental data. $k_{\text{stray}1}$ and $k_{\text{stray}2}$ of the 1.5 HP IM are shown in TABLE V. It is noted again that although $k_{\text{stray}1}$ and $k_{\text{stray}2}$ change with flux in the inverter-fed condition as shown in Fig. 39–Fig. 41, they change insignificantly in the area that of interest for loss minimization control.

\begin{table}[h]
\centering
\caption{The Look-up Table of Mechanical Loss for the 1.5 HP Motor}
\begin{tabular}{|c|c|c|c|c|c|c|c|}
\hline
\textbf{Speed (RPM)} & 1800 & 1600 & 1400 & 1200 & 1000 & 800 & 600 \\
\hline
\textbf{$P_{\text{mech}}$ (W)} & 85.05 & 70.75 & 57.19 & 43.66 & 35.62 & 26.17 & 18.10 \\
\hline
\end{tabular}
\end{table}

\begin{table}[h]
\centering
\caption{The Look-up Table of the Coefficients of Stray Loss For the 1.5 HP Motor}
\begin{tabular}{|c|c|c|c|c|c|c|c|}
\hline
\textbf{Speed (RPM)} & 1800 & 1600 & 1400 & 1200 & 1000 & 800 & 600 \\
\hline
\textbf{$k_{\text{stray}1}$} & 0.8364 & 0.8147 & 0.7078 & 0.6469 & 0.5928 & 0.5789 & 0.4276 \\
\hline
\textbf{$k_{\text{stray}2}$} & 12.204 & 6.1492 & 4.3671 & 2.5582 & 0.2837 & -0.9288 & -0.6529 \\
\hline
\end{tabular}
\end{table}
3.3 Summary

This chapter presents the details of the proposed dynamic core-loss model of three-phase IMs. This model can be transformed into any arbitrary reference frame for various control design and analysis while taking core losses in consideration. The closed-form derivation of copper and core losses is given. The core-loss model is elaboratively validated under various load, speed and flux conditions in simulation and experiments. The model parameters can be extracted simply following IEEE Standard 112 [42]. Proper parameter adaptation is needed to guarantee the accuracy of the model. The core-loss model shows excellent accuracy (> 93%) for a wide range of operating conditions and machines of different ratings. Note that small IMs have low values of losses and are susceptible to practical imperfectness than large IMs; a few-watts discrepancy in absolute value may be transferred to a large discrepancy in percentage.

Comprehensive tests on mechanical and stray losses are also performed in this chapter, where the effects of load torque, speed and flux on mechanical and stray losses are examined. It is found that, for the load area of interest to LMC, mechanical and stray losses can be treated as constant with respect to flux; Mechanical loss is independent of load torque, whereas stray loss can be modelled as a linear function of load torque squared; Both losses are affected by speed and the effects can be modelled using look-up tables.
CHAPTER 4
MODEL-BASED EFFICIENCY AND PERFORMANCE ENHANCEMENT OF THREE-PHASE INDUCTION MOTORS AND IM DRIVE SYSTEMS

The previous chapter has thoroughly introduced and validated the proposed core-loss model of three-phase IMs, and discussed the parameter adaptation of the model with respect to different speed and flux levels. One major application of the core-loss model is to perform improved model-based loss minimization control (LMC) of IMs or IM drive systems by considering core loss. Since one advantage of the proposed core-loss model is its flexibility to integrate with various controllers in various reference frames, model-based LMC is designed for two popular controllers in this chapter to prove this point, which are the open-loop V/f control and the closed-loop indirect field-oriented control (IFOC). Basically, the LMC technique here is to determine the optimal flux level that satisfies the speed and torque requirements while achieving increased efficiency or decreased loss. This optimal flux is calculated using the power loss model with known speed and torque requirements. Specifically, the LMC decides the V/f ratio command ($V_f^*$) and the synchronous-frame $d$-axis rotor flux command ($\lambda_{dr}^* \phi_e$) in V/f control and IFOC, respectively. The $V_f^*$ and $\lambda_{dr}^* \phi_e$ are set to rated values in the conventional cases when no flux-adjusting technique is engaged.

4.1 Model-based LMC of IMs Using the Proposed Core-loss Model

4.1.1 LMC Designed for the Closed-loop Indirect Field-oriented Control

4.1.1.1 Mathematical Derivation of the Optimal Flux for IFOC

Field-oriented control or vector control is known for improving machine’s dynamic performance by decoupling the flux and torque control loops. In the rotor field-oriented control of IMs, $\dot{\lambda}_{dr}^* \phi_e$ is aligned with the reference frame and thus $\lambda_{dq}^* \phi_e$ is zero. Moreover, $\dot{i}_{dr}^* \phi_e$ is also
required to be zero to guarantee decoupling of the flux and torque loops. For the proposed core-loss model, these two conditions can be satisfied if 1) $\hat{i}_{ds}$ is constant and 2) $\omega_c$ is equal to

$$\omega_c = \omega_r + \frac{R_r \cdot \hat{i}_{qs}}{L_{rr}}.$$  \hfill (4.1)

The proof is provided next.

Use the designated $\omega_c$ in (3.15) and (3.16),

$$v_{qr}' = 0 = R_r \cdot i_{dr}' + \frac{R_r \cdot \hat{i}_{qs}}{L_{rr}} \lambda_{dr} + p\lambda_{qr}'$$

$$= R_r \left( \frac{\lambda_{dr}'}{-L_m \hat{i}_{ds}'} \right) + \frac{R_r \cdot \hat{i}_{qs}}{L_{rr}} \left( L_{rr} \cdot i_{dr}' + L_m \hat{i}_{ds}' \right) + p\lambda_{qr}'$$  \hfill (4.2)

$$v_{dr}' = 0 = R_r \cdot i_{dr}' - R_r \cdot \frac{\hat{i}_{qs}}{L_{rr}} \lambda_{dr} + p\lambda_{dr}'$$

$$= R_r \cdot i_{dr}' - R_r \cdot \frac{\hat{i}_{qs}}{L_{rr}} \lambda_{dr} + L_{rr} \cdot p\lambda_{dr}' + L_m \hat{i}_{dq}'$$  \hfill (4.3)

Simplify (4.2) and (4.3),

$$p\lambda_{qr}' = -\frac{R_r \cdot \lambda_{qr}'}{L_{rr}} - R_r \cdot i_{dr}' \cdot \frac{\hat{i}_{qs}}{i_{ds}}$$  \hfill (4.4)

$$p\lambda_{dr}' = -\frac{R_r \cdot i_{dr}'}{L_{rr}} + \frac{R_r \cdot \hat{i}_{qs}}{L_{rr}} \lambda_{qr} - \frac{L_m}{L_{rr}} \hat{i}_{ds}.$$  \hfill (4.5)

On the other hand, if $\hat{i}_{ds}$ is constant, then $\hat{i}_{qs}$ is also constant

$$\hat{i}_{qs} = \frac{T_L}{3P \cdot \frac{L_m^2}{2 \cdot L_{rr}} \hat{i}_{ds}}.$$  \hfill (4.6)

Then, equations (4.4) and (4.5) imply that $\lambda_{qr}'$ and $i_{dr}'$ are asymptotically going to zero. Using these two conditions in (3.19), (3.20) and (3.29), the $\hat{i}_{qs}$ and $\hat{i}_{ds}$ commands are
\[ i_{ds}^* = \frac{\hat{\lambda}_{dr}^e}{L_m}, \tag{4.7} \]

\[ i_{qs}^* = \frac{T_L}{3 P L_m^2 \frac{L_m}{2 L_{rr}} \hat{i}_{ds}^*}, \tag{4.8} \]

which depend on the flux command and load torque.

In the conventional IFOC, \( \lambda_{dr}^e \) is set to rated value. Thus, no change of flux regulation is needed and the torque response (regulation) is fast at the maximum flux. However, this premium dynamic performance could come with the satisfaction of machine efficiency due to redundant flux in low-load conditions. The optimal flux, or \( \lambda_{dr}^e \), is derived next using the proposed core-loss model, where the value of \( \lambda_{dr}^e \) changes with speed and \( T_L \).

Since \( \lambda_{qr}^e = 0 \) and \( i_{dr}^e = 0 \), \( i_{qr}^e \) can be calculated based on (3.21) and (4.8),

\[ i_{qr}^e = -\frac{L_m}{L_{rr}} \frac{T_L}{2 L_{rr} \hat{\lambda}_{dr}^e}. \tag{4.9} \]

Therefore, \( i_{qs}^e \) and \( i_{ds}^e \) references in steady state are

\[ i_{qs}^e = \hat{i}_{qs}^e + \frac{L_{ms}}{R_c} \omega_e \hat{i}_{ds}^e = \frac{T_L}{3 P L_m^2 \frac{L_m}{2 L_{rr}} \hat{\lambda}_{dr}^e} + \frac{L_{ms}}{R_c} \omega_e \hat{\lambda}_{dr}^e, \tag{4.10} \]

\[ i_{ds}^e = \hat{i}_{ds}^e - \frac{L_{ms}}{R_c} \omega_e \hat{i}_{qs}^e = \frac{\hat{\lambda}_{dr}^e}{L_m} - \frac{L_{ms}}{R_c} \omega_e \frac{T_L}{3 P L_m^2 \frac{L_m}{2 L_{rr}} \hat{\lambda}_{dr}^e}. \tag{4.11} \]

With assumption of decent current regulation, the actual \( i_{qs}^e \) and \( i_{ds}^e \) are equal to the corresponding current commands. Then using (4.7)–(4.11) in (3.35) and (3.42) to find the machine total loss,
\[ P_{\text{mach loss}}(\lambda_{dr}^\ast) = P_{Cu} + P_{\text{core}} + P_{\text{mech}} + P_{\text{stray}} \]
\[ = \frac{2T_L^2}{3} \left( \frac{P L_m}{2 L_{rr}} \right)^2 \lambda_{dr}^\ast \left[ R_s \left( 1 + \left( \frac{L_m \omega_e}{R_e} \right)^2 \right) + \frac{R_e L_m^2}{L_{rr}^2} \omega_e^2 L_{ms} L_{ms} \left( 1 - \frac{L_m}{L_{rr}} \right) \right]. \]  \quad (4.12)

\[ + \frac{3 \lambda_{dr}^\ast}{2 L_m^2} \left[ R_s \left( 1 + \left( \frac{L_m \omega_e}{R_e} \right)^2 \right) + \frac{\omega_e^2 L_{ms} L_{ms}}{R_e} \right] \]
\[ + P_{\text{mech}} + P_{\text{stray}} \]

Note that \( P_{\text{mech}} \) and \( P_{\text{stray}} \) do not change with flux in the LMC application as discussed in Chapter 3. The optimal \( \lambda_{dr}^\ast \) can be calculated by solving

\[ \frac{dP_{\text{mach loss}}(\lambda_{dr}^\ast)}{d\lambda_{dr}^\ast} \bigg|_{\lambda_{dr}^\ast = \text{optimal}} = 0, \]  \quad (4.13)

Applying (4.12) to (4.13) and solving (4.13) for \( \lambda_{dr \_opt}^\ast \),

\[ \lambda_{dr \_opt}^\ast = \sqrt[4]{\frac{16 P L_m^2}{9 P^2} L_{rr}^2 + \frac{L_m^2 R_s \left( R_s - \omega_e^2 L_{rr} L_{ms} \right)}{R_e^2 R_s + R_e \omega_e^2 L_{ms}^2 + \omega_e^2 R_e L_{ms} L_{ms}}} \cdot T_L^2. \]  \quad (4.14)

This is the final closed-form expression of \( \lambda_{dr \_opt}^\ast \) for IFOC. Considering decent speed regulation, the \( \omega_e \) in (4.14) can be estimated as

\[ \omega_e^{\_\text{IFOC}} \approx \frac{\text{Spd}^\ast \cdot \pi}{60P}. \]  \quad (4.15)

Therefore, given speed and load torque requirements as well as the machine parameters, the optimal flux can be determined locally using (4.14) in the IFOC. It is noted that the \( T_L \) in (4.14) should be the sum of the torque added from the dynamometer and the additional torque due to mechanical and stray losses.
4.1.1.2 Simulation Verification of the Proposed LMC for the IFOC

The proposed core-loss model with IFOC is built in MATLAB/Simulink, where rated and optimal $\lambda_{dr}^{e*}$ are applied. A general-purpose 1.5 HP IM is used in the simulation, which is operated at 1800 RPM with 4 $T_L$ (0.3, 0.9, 1.8, and 3 N·m) from the dynamometer. Additional torques are added from the mechanical and stray losses. The machine parameters, which are extracted from the characterization tests at rated flux, are shown in TABLE VI for the 1800RPM condition. The optimal $\lambda_{dr}^{e*}$, which is calculated using (4.14) for the four loads, and rated $\lambda_{dr}^{e*}$ are shown in TABLE VII. It is seen that deeper flux weakening, or smaller $\lambda_{dr, opt}^{e*}$, is used for lower torques as expected.

**TABLE VI. The Parameters of the 1.5HP IM**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>$R_s$ (Ω)</th>
<th>$R_r'$ (Ω)</th>
<th>$L_{d_s}$ (H)</th>
<th>$L_{d_r}'$ (H)</th>
<th>$R_c$ (Ω)</th>
<th>$L_m$ (H)</th>
<th>$P$</th>
<th>$\omega_e$ (rad/s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>1.4437</td>
<td>1.3258</td>
<td>0.0058</td>
<td>0.0086</td>
<td>532</td>
<td>0.1837</td>
<td>4</td>
<td>120$\pi$</td>
</tr>
</tbody>
</table>

**TABLE VII. Rated and the Optimal $\lambda_{dr}^{e*}$ in the IFOC Test**

<table>
<thead>
<tr>
<th>Condition</th>
<th>Rated @ 1800RPM</th>
<th>Optimal @ 1800RPM +0.3 N·m</th>
<th>Optimal @ 1800RPM +0.9 N·m</th>
<th>Optimal @ 1800RPM +1.8 N·m</th>
<th>Optimal @ 1800RPM +3 N·m</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\lambda_{dr}^{e*}$ (Wb)</td>
<td>0.4631</td>
<td>0.181</td>
<td>0.2389</td>
<td>0.3064</td>
<td>0.3789</td>
</tr>
</tbody>
</table>

The simulation results of torque and speed responses are shown in Fig. 42 and Fig. 43, respectively. It is observed that both speed and torque responses can follow the corresponding references excellently. The modification of the flux regulation using different optimal $\lambda_{dr}^{e*}$ barely deteriorates the dynamic responses. On the other hand, the simulation results of the machine efficiency and total loss are shown in Fig. 44 and Fig. 45, respectively. It is seen that the optimal $\lambda_{dr}^{e*}$ can lead to higher machine efficiency or lower machine total loss at the low-load conditions. The improvement of efficiency is more significant at lower torques, since the
redundant degree of rated $\lambda_{dr}^{e*}$, which can be interpreted as the difference between rated and optimal $\lambda_{dr}^{e*}$ in TABLE VII is more severe at lower torques. For the highest torque in the test, the efficiency enhancement is small and further increase of $T_L$ will lead to no room for LMC. Note that the spikes of efficiency or the total loss at the torque changing transients are due to the manual change of $\lambda_{dr, opt}^{e*}$, since it is intended to show several load conditions in one figure to save the space. However, note that the focus here is to compare the efficiency or the total loss in steady state (when curves are flat).

Fig. 42. Torque response using optimal and rated $\lambda_{dr}^{e*}$ in IFOC

Fig. 43. Speed response using optimal and rated $\lambda_{dr}^{e*}$ in IFOC
4.1.1.3 Comparison of the Proposed Core-Loss Model and the Conventional Copper-Loss Model

The proposed core-loss model is evolved from the conventional dynamic three-phase IM model that only considers copper loss. To show the importance of core loss in determining the optimal flux, especially at low-load conditions, the 1.5 HP IM is simulated at 1800 RPM and $T_L=1 \text{ N}\cdot\text{m}$ with $\lambda_{dr}^{e*}$ decreases from rated value to 0.16 Wb. $P_{Cu}$, $P_{core}$ and the sum of them are shown in Fig. 46 for different $\lambda_{dr}^{e*}$. It is seen that at this load condition, the magnitude of $P_{core}$ is comparative to $P_{Cu}$. Thus, $P_{core}$ can effectively impact $P_{mach\_loss}$ and efficiency as well as the determination of the optimal $\lambda_{dr}^{e*}$. If only $P_{Cu}$ is considered, the optimal $\lambda_{dr}^{e*}$ will be $\lambda_{dr1}^{e*}$.
shown in Fig. 46. However, if the proposed core-loss model is used, $\lambda_{dr}^{e*}$ will be $\lambda_{dr2}^{e*}$, which can give additional loss reduction of $\Delta P_{loss_mach}$ over $\lambda_{dr1}^{e*}$, as shown in the figure.

This shows the contribution and the superiority of the proposed model over the conventional copper-loss model. On the other hand, if the machine is operated at a high-load condition, the core-loss model and the copper-loss model will lead to the same optimal $\lambda_{dr}^{e*}$, since $P_{Cu}$ is much larger than $P_{core}$ and thus $P_{core}$ has no effective impact. An example is shown in Fig. 47, where $P_{Cu}$, $P_{core}$ and the sum of them for the machine running at 1800 RPM and 4.5 N·m load are shown.
4.1.2 Loss Minimization Control for the Open-loop V/f Control

4.1.2.1 Mathematical Derivation of the Optimal Flux for V/f Control

In the open-loop V/f control, synchronous $qd0$-frame is used and $v_{ds}^*$ is set to zero. Moreover, $v_{qs}^*$ is set by $Vf^*$ at certain speed condition assuming $MI$ and $V_{dc}$ are fixed at rated values. In the V/f control, the $v_{qs}^*$ is

$$v_{qs}^* = Vf^* \cdot \frac{MI_{\text{rated}} \cdot V_{dc_{\text{rated}}} \cdot \text{Spi}}{2 \cdot Vf_{\text{rated}} \cdot \text{Spd_{\text{rated}}}}. \tag{4.16}$$

On the other hand, since the synchronous-frame voltage, current and flux are constant in steady state, their derivatives are zero. Applying this property and replacing the flux and stator current terms in (3.13)–(3.16) and (3.29) using (3.17)–(3.22), the synchronous-frame voltages and the induced torque in steady state can be expressed using $\hat{i}_{qs}$, $\hat{i}_{ds}$, $i_{qr}'$ and $i_{dr}'$,

$$v_{qs} = \left( R_e - \frac{\alpha_e^2 L_m L_{ms}}{R_e} \right) \hat{i}_{qs} + \left( \frac{R_e L_{ms} \alpha_e}{R_e} + \omega_e L_{ss} + \omega_e L_m \right) \hat{i}_{ds} + \omega_e L_m i_{dr}', \tag{4.17}$$

$$v_{ds} = \left( R_s - \frac{\alpha_s^2 L_m L_{ms}}{R_e} \right) \hat{i}_{ds} - \left( \frac{R_e L_{ms} \alpha_e}{R_e} + \omega_e L_{ss} + \omega_e L_m \right) \hat{i}_{qs} - \omega_e L_m i_{qr}', \tag{4.18}$$

$$v_{q}' = 0 = R_e i_{q}', (\omega_e - \omega_s) L_m i_{dr}' + (\omega_e - \omega_s) L_m \hat{i}_{ds}, \tag{4.19}$$

$$v_{d}' = 0 = R_e i_{d}', (\omega_e - \omega_s) L_m i_{qr}' - (\omega_e - \omega_s) L_m \hat{i}_{qs}. \tag{4.20}$$

$$T_e = \frac{3PL_m}{4} \left( \hat{i}_{qdr}' - \hat{i}_{qsr}' \right). \tag{4.21}$$

Here, $v_{qs}$ and $v_{ds}$ will eventually be equal to the voltage commands, and $T_e$ will be the same as the known $T_L$ in steady state. Therefore, replacing $v_{qs}$, $v_{ds}$ and $T_e$ in (4.17), (4.18) and (4.21) by (4.16), zero and $T_L$ respectively, the resultant five equations construct a six-variable linear set of
equations. The variables are \( V_f^\ast, i_{q^r}, i_{d_L}, i_{q^r'}, i_{d_r'}, \) and \( \omega_r, \omega_e \) in the equations can be calculated by the speed command,

\[
\omega_r = \frac{Spd^\ast \cdot \pi}{60P}.
\]

(4.22)

Technically, solving the set of linear equations, \( i_{q^r}, i_{d_L}, i_{q^r'}, i_{d_r'}, \) and \( \omega_r \) can be expressed as functions of \( V_f^\ast \) for certain \( Spd^\ast \) and \( T_L \). Then, \( P_{\text{mach,loss}} \) can also be expressed as a function of \( V_f^\ast \). Following the same procedure of obtaining \( \lambda_{dr,\text{opt}}^\ast \), the optimal \( V_f^\ast \) can be calculated by solving

\[
\left. \frac{dP_{\text{mach,loss}}(V_f^\ast)}{dVf^\ast} \right|_{Vf^\ast=\text{optimal}} = 0.
\]

(4.23)

However, the expression of \( P_{\text{mach,loss}}(V_f^\ast) \) is too complicated to show here. Moreover, equation (4.23) cannot be solved, at least not instantaneously, in MATLAB using the SOLVE function. Therefore, an alternative approach is applied to find the optimal \( V_f^\ast \).

Considering theoretically \( P_{\text{mach,loss}} \) is a concave function (or efficiency is a convex function) of flux as shown in Fig. 46, and several-digit accuracy of \( V_f^\ast \) is sufficient, a numerical sweep is used to find out the optimal \( V_f^\ast \) (\( V_{f,\text{opt}}^\ast \)). The implementation of the numerical sweep is through MATLAB code, which is shown in Appendix A. Basically, a For loop is applied to \( V_f^\ast \) which decreases from rated value (2.213V·s) with a step of 0.05V·s. \( P_{\text{mach,loss}}(V_f^\ast) \) or machine efficiency is calculated in each iteration for different \( V_f^\ast \). After the For loop concludes, the \( V_f^\ast \) that gives the minimum \( P_{\text{mach,loss}} \) or the maximum efficiency is the estimated \( V_{f,\text{opt}}^\ast \). Note that the accuracy of \( V_{f,\text{opt}}^\ast \) can be improved if decreasing \( V_f^\ast \) step size, but the current step size is small enough since \( P_{\text{mach,loss}} \) or efficiency is relatively flat around \( V_{f,\text{opt}}^\ast \). Moreover, if the step size is
very small, the For loop will take longer to conclude and will deteriorate real-time LMC in experiment.

4.1.2.2 Simulation Verification of the Proposed LMC for V/f Control

A similar simulation verification to that for the IFOC case is applied for the LMC with V/f control. The previous Simulink model of the 1.5 HP IM is used, but the control is changed from the IFOC to the V/f control. A numerical sweep is applied to determine $Vf_{opt}^{*}$ at different operating conditions. The IM is tested at 1800 RPM and four load torques: 0.3, 0.9, 1.8 and 3 N·m. Rated and optimal $Vf^{*}$ are shown in TABLE VIII. Note that the RMS value of rated phase voltage and rated frequency in Hz are used to calculate $Vf^{*}$ (stator flux) in TABLE VIII. If comparison between these $Vf^{*}$ and the rotor flux in TABLE VII is wanted, the peak value of the phase voltage and the frequency in rad/s need to be used to calculate $Vf^{*}$.

<table>
<thead>
<tr>
<th>Condition</th>
<th>Rated @ 1800RPM</th>
<th>Optimal @ 1800RPM</th>
<th>Optimal @ 1800RPM</th>
<th>Optimal @ 1800RPM</th>
<th>Optimal @ 1800RPM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Vf$ (V·s)</td>
<td>2.213</td>
<td>0.913</td>
<td>1.213</td>
<td>1.513</td>
<td>1.813</td>
</tr>
</tbody>
</table>

The torque and speed responses of using rated and optimal $Vf^{*}$ are shown in Fig. 48 and Fig. 49, respectively, along with the speed and torque references. It is seen that the torque responses track the reference excellently. The speed responses do not follow the reference as well as using the IFOC due to lack of speed regulation. Moreover, as expected, the slip increases with the increase of load torque when using rated $Vf^{*}$. If the optimal $Vf^{*}$ is used, the slip maintains a relatively large value. This is a tradeoff of improving efficiency using the LMC designed for the V/f control.
The simulation results of the machine efficiency and total loss are shown in Fig. 50 and Fig. 51. Using the optimal $Vf^*$, the machine efficiency is increased and the total loss is decreased. Similar to the LMC for IFOC, the room for efficiency enhancement increases at lower $T_L$. Note that the focus here is still at the steady-state comparison. Thus, the transient spikes in Fig. 50 and Fig. 51 are ignored.

![Torque response using optimal and rated $Vf^*$](image1)

**Fig. 48.** Torque response using optimal and rated $Vf^*$

![Speed response using optimal and rated $Vf^*$](image2)

**Fig. 49.** Speed response using optimal and rated $Vf^*$
4.2 System-level Model-based Loss Minimization Control of IMs Drives

The previous discussion shows that integrating core loss in the machine modeling and design of LMC can further increase the machine efficiency compared to the conventional copper-loss model. Considering the optimal operating point of an IM drive system could be different from that of an individual IM or drive, this section introduces the design of the system-level LMC for IM drives. An inverter-fed IM system is used and the goal is to improve the overall efficiency of the inverter and motor combination. Note that that a commercial drive will
also have a converter/rectifier stage, but modeling this stage will be similar to modeling the inverter stage.

The design of system-level LMC follows the same procedure as that for just the IM. The previous core-loss model is used to model the losses in IMs. Moreover, a power loss model of three-phase inverters, which is combined with the machine core-loss model, is applied to give the overall loss model of the inverter-fed IM drives. The LMC is also designed for the IFOC and V/f cases, where the optimal flux is determined based on the drive loss model for certain speed and torque requirement.

4.2.1 Modeling of Power Electronics Losses in Three-Phase Inverters

Common three-phase inverters have six active switches, e.g. IGBTs or MOSFETs, and six anti-parallel free-wheeling diodes. There are mainly two types of losses: conduction loss and switching loss. The conduction loss comes from both switches and diodes. The switching loss includes switch-on and switch-off losses from switches and the reverse-recovery loss from diodes. These losses can be modeled based on the datasheet of the power modules [114].

The $I_{CE}$-$V_{CE}$ and $I_{F}$-$V_{F}$ curves of Infineon BSM75GB60DL, which is the IGBT module used in the lab, are shown in Fig. 52 and Fig. 53, respectively, as example curves. These curves can be modeled as parabolic functions,

$$v_x = \beta_{1x} \cdot i_x^2 + \beta_{2x} \cdot i_x + \beta_{3x},$$

(4.24)

where $x$ can be $ce$ or $f$ to represent the switch or diode, respectively. $\beta_{1x}$, $\beta_{2x}$, $\beta_{3x}$ are the curve fitting coefficients. The values of $\beta_{1x}$, $\beta_{2x}$, $\beta_{3x}$ for the Infineon BSM75GB60DL IGBT module are summarized in TABLE IX. Moreover, assume the duty ratios and load current are

$$d_{ce} = \frac{1}{2} + \frac{MI}{2} \sin (\omega t + \phi),$$

(4.25)
\[ d_f = \frac{1}{2} \frac{MI}{2} \sin(\omega_f t + \phi), \]  
\[ (4.26) \]

\[ i_L = I_L \sin(\omega_c t), \]  
\[ (4.27) \]

where \( d_{ce} \) and \( d_f \) are the duty ratios of the IGBT and diode, respectively. \( \phi \) is the phase difference between the fundamental phase voltage and current. Here, assume the load current is purely sinusoidal. Therefore, the conduction loss of a single switch or diode can be calculated by

\[ P_c = \frac{1}{T} \int_{0}^{T/2} (v_i i_x d_x) dt = \delta_{1f} I_L^3 + \delta_{2f} I_L^2 + \delta_{3f} I_L, \]  
\[ (4.28) \]

where

\[ \delta_{i ce} = \frac{\beta_{i ce}}{3\pi} + \frac{3\beta_{i ce} \cdot MI \cos \phi}{32}, \]  
\[ (4.29) \]

\[ \delta_{2 ce} = \frac{\beta_{2 ce}}{8} + \frac{\beta_{2 ce} \cdot MI \cos \phi}{3\pi}, \]  
\[ (4.30) \]

\[ \delta_{3 ce} = \frac{\beta_{3 ce}}{2\pi} + \frac{\beta_{3 ce} \cdot MI \cos \phi}{8}, \]  
\[ (4.31) \]

\[ \delta_{1f} = \frac{\beta_{1 f}}{3\pi} - \frac{3\beta_{1 f} \cdot MI \cos \phi}{32}, \]  
\[ (4.32) \]

\[ \delta_{2f} = \frac{\beta_{2 f}}{8} - \frac{\beta_{2 f} \cdot MI \cos \phi}{3\pi}, \]  
\[ (4.33) \]

\[ \delta_{3f} = \frac{\beta_{3 f}}{2\pi} - \frac{\beta_{3 f} \cdot MI \cos \phi}{8}, \]  
\[ (4.34) \]

\( T \) is the fundamental period.
Fig. 52. The $I_C$-$V_{CE}$ curve in the datasheet of Infineon BSM75GB60DL.

Fig. 53. $I_F$-$V_F$ curve in the datasheet of Infineon BSM75GB60DL.
On the other hand, switching loss can be calculated based on the energy curves in the datasheet. The energy curves of the Infineon BSM75GB60DL module are shown in Fig. 54, which represent the energy consumption of one-time switch-on, switch-off and reverse-recovery action, respectively. Therefore, the accumulation of energy loss of each switching action for 1s is the switching power loss. The energy curves can be modeled as linear functions of $I_C$ or $i_L$ ($i_L$ is equal to the $I_C$ in Fig. 54)

$$E_{on} = E_{on,sl} \cdot i_L + E_{on,ini}, \quad (4.35)$$

$$E_{off} = E_{off,sl} \cdot i_L + E_{off,ini}, \quad (4.36)$$

$$E_{rev} = E_{rev,sl} \cdot i_L + E_{rev,ini}. \quad (4.37)$$

The curve-fitting coefficients of the energy curves are summarized in Table IX. Assume $E_{sw,sl}$ and $E_{sw,ini}$ are

$$E_{sw,sl} = E_{on,sl} + E_{off,sl} + E_{rev,sl}, \quad (4.38)$$

$$E_{sw,ini} = E_{on,ini} + E_{off,ini} + E_{rev,ini}, \quad (4.39)$$

which are defined for discussion convenience. Then, the total switching loss of an inverter is

$$P_{sw} = 6 \cdot \frac{f_{sw}}{2} \sum_{j=1}^{f_{sw}/2} [E_{sw,sl} \cdot i_L(j) + E_{sw,ini}], \quad (4.40)$$

where $f_{sw}$ is the switching frequency, $i_L(j)$ is the discrete representation of the load current at the switching instance $j$.  

<table>
<thead>
<tr>
<th>Coefficient $\beta$</th>
<th>$\beta_{ce1}$</th>
<th>$\beta_{ce2}$</th>
<th>$\beta_{ce3}$</th>
<th>$\beta_{f1}$</th>
<th>$\beta_{f2}$</th>
<th>$\beta_{f3}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>-1e-4</td>
<td>0.0233</td>
<td>0.7748</td>
<td>-4e-5</td>
<td>0.009</td>
<td>0.7851</td>
</tr>
</tbody>
</table>
\[ i_L(j) = I_L \sin \left( \frac{2\pi}{f_{sw}} j \right). \]  

Therefore,  

\[ P_{sw} = 6 \cdot \left( E_{sw,sl} \cdot I_L \cdot \frac{f_{sw}}{\pi} + E_{sw,ini} \cdot \frac{f_{sw}}{2} \right), \]  

(4.42)

Fig. 54. Energy curves in the datasheet of Infineon BSM75GB60DL

However, it is noted that the energy curves in the datasheet are tested at the specific condition shown in the datasheet. If the real operating condition is different from the one used in the datasheet, modification on (4.41) is needed \[115], \[116]. In our case, modifications on temperature and DC link voltage are applied. Specifically, to compensate the temperature difference,  

\[ E_{xx,ini}^{(real)} = E_{xx,ini}^{(datasheet)} \cdot \left( \frac{T_{real}}{T_{datasheet}} \right)^{D_{xx}}, \]  

(4.43)
where \( xx \) can be \( on \), \( off \) or \( rev \) to represent the switch-on, switch-off and reverse-recovery energy curves. \( T_{real} \) and \( T_{datasheet} \) are the temperature used in the real application and datasheet, respectively. Technically, the values of the temperature compensation coefficients, \( D_{on}, D_{off}, D_{rev} \), should be measured experimentally for the applied IGBT module, but the typical values for the second generation of the IGBTs are used, which are 0.51614, 0, 0.85722 [117]. On the other hand, the difference of the DC link voltage in the real application and datasheet can be compensated by

\[
\begin{aligned}
P_{sw(real)} &= P_{sw(datasheet)} \cdot \frac{V_{dc\_real}}{V_{dc\_datasheet}}. \\

\end{aligned}
\]

Therefore, the total loss of the three-phase inverter is

\[
\begin{aligned}
P_{in\_loss}(I_L, PF) &= 6 \cdot \left( \delta_{1ce} I_L^3 + \delta_{2ce} I_L^2 + \delta_{3ce} I_L + \delta_{1f} I_L^3 + \delta_{2f} I_L^2 + \delta_{3f} I_L \right) \\
&\quad + 6 \cdot \left( E_{sw,sl} \cdot I_L \cdot \frac{f_{sw}}{\pi} + E_{sw,ini\_real} \cdot \frac{f_{sw}}{2} \right) \cdot \frac{V_{dc\_real}}{V_{dc\_datasheet}}.
\end{aligned}
\]

where \( PF \) is power factor (\( \cos \phi \)), and

\[
\begin{aligned}
E_{sw,ini\_real} &= E_{on,ini} \left( \frac{T_{real}}{T_{datasheet}} \right)^{D_{on}} + E_{off,ini} \left( \frac{T_{real}}{T_{datasheet}} \right)^{D_{off}} + E_{rev,ini} \left( \frac{T_{real}}{T_{datasheet}} \right)^{D_{rev}}.
\end{aligned}
\]

Equation (4.45) shows that the inverter total loss is a function of \( I_L \) and \( PF \) with given curve-fitting coefficients, \( f_{sw}, V_{dc} \) and the temperature in real application and datasheet. On the other hand, \( I_L \) and \( PF \) can be represented by the synchronous-frame voltages,

\[
\begin{aligned}
I_L &= \sqrt{i_{qs} e^2 + i_{ds} e^2}, \\
PF &= \cos \left( \arctan \left( \frac{i_{ds} e}{i_{qs} e} \right) \right). \\
\end{aligned}
\]

Therefore, the inverter total loss can be expressed as a function of \( i_{qs} e \) and \( i_{ds} e \).
TABLE X. CURVE-FITTING COEFFICIENTS OF THE ENERGY CURVES

<table>
<thead>
<tr>
<th>Coefficient</th>
<th>$E_{on,sl}$ (mJ)</th>
<th>$E_{off,sl}$ (mJ)</th>
<th>$E_{rev,sl}$ (mJ)</th>
<th>$E_{on,ini}$ (mJ)</th>
<th>$E_{off,ini}$ (mJ)</th>
<th>$E_{rev,ini}$ (mJ)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>0.0068</td>
<td>0.0263</td>
<td>0.0077</td>
<td>0.1438</td>
<td>0.4305</td>
<td>1.7626</td>
</tr>
</tbody>
</table>

4.2.2 LMC of IFOC-Controlled Inverter-fed Motor Drives

The previous discussion shows that $P_{inv,loss}$ can be represented as a function of $i_{qs}e$ and $i_{ds}e$. Moreover, since $i_{qs}e$ and $i_{ds}e$ can be expressed by $\lambda_{dr}e^*$ as shown in (4.10) and (4.11), $P_{inv,loss}$ and the total loss of the inverter-fed motor drive, $P_{drive,loss}$, can be expressed as functions of $\lambda_{dr}e^*$. These functions of $\lambda_{dr}e^*$ are too complex. Thus, the numerical sweep is used to find the optimal $\lambda_{dr}e^*$, which is implemented through MATLAB code as shown in Appendix B.

Since the Infineon BSM75GB60DL IGBT module is much over-rated for the 1.5 HP IM, a 50HP IM is used in the simulation verification here to show effective impacts of the inverter losses. The machine parameters and the basic information of the 50 HP IM and the inverter are shown in TABLE XI. The IM is tested at 1800 RPM and three load torques: 21 N·m, 42 N·m and 63 N·m, where the numerical sweep changes $\lambda_{dr}e^*$ from rated value to 0.15 Wb when the flux is too weak to support the load. The step size of the flux sweep is 0.01 Wb. Rated $\lambda_{dr}e^*$ and the optimal $\lambda_{dr}e^*$ are summarized in TABLE XII. The previous Simulink model of the IFOC-controlled IM is applied again but with three modifications: 1) The machine parameters are changed from the 1.5 HP motor to the 50 HP motor; 2) A block to calculate the inverter loss is added to the simulation; and 3) Mechanical and stray losses are removed from the simulation due to lack of experimental data, but it does not affect showing a proof of concept.

Simulation results of the torque and speed responses are shown in Fig. 55 and Fig. 56, respectively. It is seen that both rated and the optimal $\lambda_{dr}e^*$ can lead to excellent tracking of speed and torque references, and no obvious difference in performance is observed between
using rated and the optimal $\lambda_{dr}^{*}$. In addition, the transient here is longer than using the 1.5HP IM, since the inertia of the 50 HP IM is much larger than that of the 1.5 HP motor, and the PI controllers also need more time to regulate the larger currents. The efficiency and the total loss of the inverter-fed motor drive are shown in Fig. 57 and Fig. 58, respectively. Again, the optimal $\lambda_{dr}^{*}$ leads to increased drive efficiency or decreased $P_{drive\_loss}$ at different low-load conditions. Moreover, the room for LMC shrinks as $T_L$ increases as before.

<table>
<thead>
<tr>
<th>TABLE XI. THE PARAMETERS AND BASIC INFORMATION OF THE 50 HP IM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameter</td>
</tr>
<tr>
<td>Value</td>
</tr>
<tr>
<td>Parameter</td>
</tr>
<tr>
<td>Value</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>TABLE XII. RATED AND THE OPTIMAL $\lambda_{dr}^{*}$ AT THE TESTING CONDITIONS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Condition</td>
</tr>
<tr>
<td>$\lambda_{dr}^{*}$ (Wb)</td>
</tr>
</tbody>
</table>

To show the impact of the inverter loss in changing the optimal flux point of the overall motor drive system, $P_{mach\_loss}$, $P_{inv\_loss}$, and $P_{drive\_loss}$ of the motor drive operating at 1800 RPM and $T_L=15$ N·m are shown in Fig. 59 with respect to different $\lambda_{dr}^{*}$. It is observed that the individual motor or inverter will give the optimal flux at $\lambda_{dr3}^{*}$ or $\lambda_{dr4}^{*}$, respectively. But the optimal flux of the overall system is at $\lambda_{dr5}^{*}$. Difference of drive loss using $\lambda_{dr3}^{*}$ and $\lambda_{dr5}^{*}$ is shown in the figure. Therefore, creating a system-level power loss model is important to guarantee the optimal efficiency of the overall motor drive systems. For other operating conditions or other motors and inverters, where the magnitude of $P_{mach\_loss}$ and $P_{inv\_loss}$ are more
comparable and $P_{\text{inv.loss}}$ is more significantly changed with flux, difference of drive loss using $\lambda_{dr1}^{*e*}$ and $\lambda_{dr5}^{*e*}$ could be much larger.

Fig. 55. The torque responses in the IFOC-controlled inverter-fed motor drive

Fig. 56. The speed responses in the IFOC-controlled inverter-fed motor drive

Fig. 57. The efficiency of using the optimal and rated $\lambda_{dr}^{*e*}$ in the IFOC-controlled inverter-fed motor drive
4.2.3 LMC of V/f-Controlled Inverter-fed Motor Drives

The design of the LMC for V/f-controlled inverter-fed motor drives follows the same procedure as that for the IFOC-controlled drives. The 50 HP IM and the Infineon BSM75GB60DL IGBTs are used again as well as the same $MI, f_{sw}, T_{real}$ and $V_{dc_{real}}$. The motor is also simulated at 1800 RPM and the three $T_L$: 21 N·m, 42 N·m and 63 N·m. Solving (4.17) to (4.21), then $P_{inv_loss}$ and $P_{drive_loss}$ can be represented as functions of $Vf^*$. Performing a numerical sweep on $Vf^*$, which is implemented using MATLAB code, the $Vf^*$ decreases from rated value with a step of 0.1 V·s. Rated and the optimal $Vf^*$ at the three testing conditions are summarized in TABLE XIII. It is seen that deeper flux weakening is available at lower $T_L$ as $Vf_{opt}^*$ decreases.
with the decrease of $T_L$ in TABLE XIII. The speed response, torque response, drive efficiency and the $P_{\text{drive\_loss}}$ of the V/f-controlled inverter-fed motor drive are shown in Fig. 61–Fig. 63, respectively. It is observed that the optimal $V_f^*$ gives higher drive efficiency than rated $V_f^*$ with unnoticeable extra cost of the dynamic performance.

**TABLE XIII. RATED AND OPTIMAL $V_f^*$ FOR THE V/f CONTROLLED INVERTER-FED MOTOR DRIVE**

<table>
<thead>
<tr>
<th>Condition</th>
<th>Rated @ 1800RPM</th>
<th>Optimal @ 1800RPM +21 N\cdot m</th>
<th>Optimal @ 1800RPM +42 N\cdot m</th>
<th>Optimal @ 1800RPM +63 N\cdot m</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_f^*$ (V\cdot s)</td>
<td>4.4264</td>
<td>2.0264</td>
<td>2.8624</td>
<td>3.4264</td>
</tr>
</tbody>
</table>

Fig. 60. The torque response of the V/f-controlled inverter-fed motor drive

Fig. 61. The speed response of the V/f-controlled inverter-fed motor drive
4.2.4 Experimental Verification of the Proposed LMC for Inverter-Fed Motor Drives

Experimental validation of the proposed system-level LMC algorithms is carried out on the setup shown in Fig. 21. The 1.5 HP IM and the Infineon BSM75GB60DL IGBT, which are used in the previous simulation verification, are used in the experiment. In the experimental verification, the flux of the IM is decreased from rated value until the efficiency drops significantly. The efficiency and total loss of the drive system at each flux point are measured. Then, the flux point that gives the maximum efficiency or minimum loss is defined as the measured optimal flux point of the drive system. Meanwhile, the proposed LMC is used to estimate the optimal flux point at the same condition. Finally, the measured and the estimated
optimal flux points are compared at several speed and load torques. The results of the comparison indicate the validity and accuracy of the proposed model-based system-level LMC.

The LMC designed for the V/f-controlled motor drive is examined first at two operating conditions: 1) 1800 RPM and 1.2 N·m; 2) 1200 RPM and 0.6 N·m, where the $V_f^*$ is decreased from 2.213 V·s with a step of 0.1 V·s. The previous discussion shows that $R_c$ and $L_m$ could change with flux. Therefore, no-load characterization tests are performed at five $V_f^*$, and the obtained $R_c$ and $L_m$ are used for curve fitting. The $R_c$ and $L_m$ at the two testing speeds and different $V_f^*$ are summarized in TABLE XIV.

**TABLE XIV. The $R_c$ and $L_m$ at 1800 RPM and 1200 RPM with Different $V_f^*$**

<table>
<thead>
<tr>
<th>$Spd^*$</th>
<th>1800 RPM</th>
<th>1200 RPM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Vf^*$ (V·s)</td>
<td>2.213</td>
<td>1.813</td>
</tr>
<tr>
<td>$R_c$ (Ω)</td>
<td>578.4</td>
<td>636.1</td>
</tr>
<tr>
<td>$L_m$ (H)</td>
<td>0.1645</td>
<td>0.1844</td>
</tr>
</tbody>
</table>

For the first operating condition, the measured and simulated $P_{mach\_loss}$ and $P_{inv\_loss}$ are compared in Fig. 64. The proposed model gives excellent estimation on the machine and inverter losses. Moreover, the measured and simulated $P_{drive\_loss}$ and efficiency are shown in Fig. 65 and Fig. 66, respectively. The larger dot indicates the optimal flux point in each curve. It is seen that the proposed LMC gives the optimal $Vf^*$ at 1.413 V·s, while the measured optimal $Vf^*$ is at 1.513 V·s. Therefore, the proposed LMC is quite accurate considering the power loss and efficiency near the optimal $Vf^*$ are relatively flat with respect to $Vf^*$. For example, the measured $P_{drive\_loss}$ at 1.513 V·s and 1.413 V·s are 201.75 W and 203.12 W, respectively. The difference of them is just 1.37 W or 0.68%, which can be easily caused by practical nonidealities. On the other hand, the measured $P_{drive\_loss}$ at rated $Vf^*$ is 235.68 W. Therefore, the proposed LMC, when makes $Vf^*$=1.413 V·s, can decease $P_{drive\_loss}$ by 32.56 W or 13.8% at this operating condition.
The comparison of $P_{\text{drive loss}}$ and efficiency at the second operating condition are shown in Fig. 67 and Fig. 68, which gives the similar conclusions to those for the previous operating condition. Based on Fig. 67, the measured and estimated $Vf_{\text{opt}}^*$ are 1.313 V·s and 1.113 V·s, respectively. Based on Fig. 68, the measured and estimated $Vf_{\text{opt}}^*$ are 1.413 V·s and 1.213 V·s, respectively. Fig. 67 and Fig. 68 give a slightly different $Vf_{\text{opt}}^*$, since the curves around $Vf_{\text{opt}}^*$ are very flat. Although the estimated $Vf^*$ is not exactly the same as the measured $Vf^*$, it can lead to significant loss reduction compared to using rated $Vf^*$ (from 167.88 W to 144.58 W, 13.9% loss reduction).

![Graph 64](image1.png)

**Fig. 64.** Comparison of the measured and simulated machine loss and inverter loss in V/f control at 1800 RPM and 1.2 N·m

![Graph 65](image2.png)

**Fig. 65.** Comparison of the measured and simulated drive total loss in V/f control at 1800 RPM and 1.2 N·m
Fig. 66. Comparison of the measured and simulated drive efficiency in V/f control at 1800 RPM and 1.2 N·m

Fig. 67. Comparison of the measured and simulated drive total loss in V/f control at 1200 RPM and 0.6 N·m

Fig. 68. Comparison of the measured and simulated drive efficiency in V/f control at 1200 RPM and 0.6 N·m
The LMC for the IFOC is supposed to be experimentally verified using the same procedure as before: Comparing the measured and estimated optimal $\lambda_{dr}^{*e}$. However, future work will address the IFOC experimental implementation. Presently, the torque and flux loops cannot be decoupled clearly. $T_e^*$ (output of speed PI) is not equal to $T_L$ and it changes with $\lambda_{dr}^{*e}$. This could be caused by the well-known parameter sensitivity issue of IFOC controller. Further work is required in improving the IFOC controller in the real setup. Once it is finished, the experimental verification of the LMC designed for the IFOC can be continued. On the other hand, if the $T_e^*$, which is the torque command generated in the experiment, is used in the simulation, the measured and the simulated $P_{drive_loss}$ are compared in Fig. 69. It is shown that the proposed drive loss model can estimate $P_{drive_loss}$ decently. Therefore, it is promising to use the proposed LMC to get the optimal $\lambda_{dr}^{*e}$ if $T_e^*$ can be regulated correctly to the desired value ($T_L$).

![Graph showing comparison of drive total loss](image)

Fig. 69. Comparison of drive total loss when using the $T_e^*$ obtained from the experiment in the simulation for LMC designed for IFOC

### 4.3 Improved Maximum Torque-per-Ampere Control of IMs by Considering Core loss

An improved maximum torque-per-ampere (MTPA) control is introduced in this chapter as another example of using the proposed core-loss model, which can be used as a general basis
for various control design by considering core loss. The specific interest here is the torque-per-ampere capability. New sets of references for current regulation are derived by considering core loss, which can give higher torque-per-ampere capability than the conventional MPTA control.

4.3.1 Derivation of the Proposed MTPA Control

The \( qd0 \)-frame torque and flux equations of the core-loss model are re-written here for reader’s convenience.

\[
T_e = \frac{3}{2} \frac{P}{2} \left( \lambda_{qr} ' i_{dr} ' - \lambda_{dr} ' i_{qr} ' \right), \quad (4.49)
\]

\[
\lambda_{qr} ' = L_{ss} ' i_{qr} ' + L_m \hat{i}_{qs}, \quad (4.50)
\]

\[
\lambda_{dr} ' = L_{ss} ' i_{dr} ' + L_m \hat{i}_{ds}. \quad (4.51)
\]

In the rotor-flux-oriented condition, \( \lambda_{qr}^e \) and \( i_{dr}^e \) are zero. Therefore,

\[
i_{qr}^e = -\frac{L_m}{L_{rr}} \hat{i}_{qs}^e, \quad (4.52)
\]

\[
T_e = \frac{3}{2} \frac{P}{2} \frac{L_m^2}{L_{rr}} \hat{i}_{qs}^e \cdot \hat{i}_{ds}^e. \quad (4.53)
\]

On the other hand, \( v_{qr}^e \) in the synchronous frame is

\[
0 = v_{qr}^e = R_e' i_{qr}^e + (\omega_e - \omega_r) \lambda_{dr}^e. \quad (4.54)
\]

Substituting (4.51) and (4.52) in (4.54) leads to the slip frequency,

\[
\omega_s = \frac{R_e'}{L_{rr}} \cdot \frac{\hat{i}_{qs}^e}{\hat{i}_{ds}^e}. \quad (4.55)
\]

Referring to Chapter 3 again, the stator currents in the synchronous frame and steady state are

\[
i_{qs}^e = \hat{i}_{qs}^e + \frac{L_m}{R_e} \omega_e \hat{i}_{ds}^e, \quad (4.56)
\]
\begin{equation}
\dot{i}_{ds}^e = \dot{i}_{ds}^e - \frac{L_{ms}}{R_c} \omega_e \dot{i}_{qs}^e.
\end{equation}

Thus, the current magnitude is

\begin{equation}
I_{mag} = \sqrt{i_{qs}^e + i_{ds}^e} = \sqrt{1 + \frac{L_{ms} \omega_e^2}{R_c^2} \left( \dot{i}_{qs}^e + \dot{i}_{ds}^e \right)}.
\end{equation}

Given certain torque and \(\omega_e\) and by considering (4.53) and (4.58) together, the true MTPA condition happens at

\begin{equation}
\dot{i}_{qs}^e = \dot{i}_{ds}^e,
\end{equation}

instead of at

\begin{equation}
i_{qs}^e = i_{ds}^e.
\end{equation}

This implies modification of current regulation in the proposed MTPA control, which would use new current references to increase machine’s torque-per-ampere capability. Applying (4.59) to (4.55), the \(\omega_s\) at the proposed MTPA condition is

\begin{equation}
\omega_{s,MTPA} = \frac{R'}{L_r',} = \frac{1}{\tau_r},
\end{equation}

where \(\tau_r\) is the machine’s rotor time constant. Therefore, the optimal \(\omega_s\) under the proposed MTPA control is still equal to the inverse of rotor time constant as in the conventional MTPA control. However, \(i_{qs}^e\) and \(i_{ds}^e\) are no longer regulated to the same value. Instead, they have to form a pattern so that (4.59) can be satisfied.

4.3.2 Improvement of the Torque-per-Ampere Capability Using the Proposed MPTA Control

To explore the possible enhancement of torque-per-ampere (TPA) capability using the proposed MTPA control, the torque generated from the 1) proposed and 2) the conventional MTPA control are compared, where the current magnitudes of the two conditions are kept the
same and the machine is not saturated. In the following derivation, subscripts 1 and 2 represent variables of conditions 1) and 2), respectively. The comparing process is shown in Fig. 70.

In the proposed MTPA control,

\[ \dot{i}_{q1}^e = \dot{i}_{d1}^e = \dot{i}_{s1}^e, \]  

(4.62)

where \( \dot{i}_{s1} \) is variable created for derivation convenience. Assume \( T_{e1} \) is the torque generated by the proposed MTPA control and apply (4.62) in (4.53), (4.56) and (4.57),

\[ T_{e1} = \frac{3}{2} P \frac{L_m}{2} \dot{i}_{s1}^e, \]  

(4.63)

\[ T_{e2} = \frac{3}{2} P \frac{L_m}{2} \dot{i}_{s2}^e, \]  

(4.64)

\[ i_{qs1}^e = \left(1 + \frac{L_m}{R_e} \omega_e\right) \dot{i}_{s1}^e, \]  

(4.65)

\[ i_{ds1}^e = \left(1 - \frac{L_m}{R_e} \omega_e\right) \dot{i}_{s1}^e. \]

Therefore, the corresponding current magnitude \( I_{mag1} \) is

\[ I_{mag1} = \sqrt{i_{qs1}^e \dot{i}_{qs1}^e + i_{ds1}^e \dot{i}_{ds1}^e} = 2 \dot{i}_{s1}^e \sqrt{1 + \frac{L_m^2}{R_e^2} \omega_e^2}. \]  

(4.66)

On the other hand, in the conventional MTPA control,
\( i_{qs}^e = i_{ds}^e \), \( (4.67) \)

\[
I_{mag2} = \sqrt{i_{qs}^{e2} + i_{ds}^{e2}} = \sqrt{2} i_{qs}^{e} = \sqrt{2} i_{ds}^{e}.
\]

\( (4.68) \)

Then, let the conventional MTPA control have the same current magnitude as the proposed MTPA control by equating (4.66) and (6.68). Thus,

\[
i_{qs}^{e} = i_{ds}^{e} = \hat{i}_{s1}^{e} \sqrt{1 + \frac{L_{ms}^2}{R_e^2 \omega_e^2}}.
\]

\( (4.69) \)

Apply (6.69) to (6.56) and (6.57), and solve for \( \hat{i}_{qs}^{e} \) and \( \hat{i}_{ds}^{e} \),

\[
\hat{i}_{qs}^{e} = \frac{R_e - L_{ms} \omega_e}{\sqrt{R_e^2 + L_{ms}^2 \omega_e^2}} \hat{i}_{s1}^{e},
\]

\( (4.70) \)

\[
\hat{i}_{ds}^{e} = \frac{R_e + L_{ms} \omega_e}{\sqrt{R_e^2 + L_{ms}^2 \omega_e^2}} \hat{i}_{s1}^{e}.
\]

\( (4.71) \)

Using (6.70) and (6.71) in (6.53), the torque generated by the conventional MTPA control, therefore, is

\[
T_{e2} = \frac{3 P L_{ms}^2}{2 L_{rr}} \frac{R_e^2 - L_{ms}^2 \omega_e^2}{R_e^2 + L_{ms}^2 \omega_e^2} \hat{i}_{s1}^{e2}.
\]

\( (4.72) \)

Finally, the torque ratio \( K \) of the proposed and the conventional MTPA control at the same current magnitude is

\[
K = \frac{T_{e1}}{T_{e2}} = \frac{T_{e_{proposed}}}{T_{e_{conventional}}} = \frac{R_e^2 + (L_{ms} \omega_e)^2}{R_e^2 - (L_{ms} \omega_e)^2}.
\]

\( (4.73) \)

Therefore, the corresponding percentage enhancement of the machine’s TPA capability is

\[
\Delta TPA = \left( \frac{T_{e1}/I_{mag1}}{T_{e2}/I_{mag2}} - 1 \right) \times 100\% = (K - 1) \times 100\%.
\]

\( (4.74) \)
It is shown that the torque ratio $K$ and $\Delta TPA$ are functions of machine parameters and $\omega_e$. Considering $L_{ms}$ is almost constant along with $\omega_e$ before saturation, a typical $K$-space with respect to $R_c$ and $\omega_e$ is shown in Fig. 71. Moreover, $\Delta TPA$ at an arbitrarily constant $R_c$ or $\omega_e$ are shown in Fig. 72. For 50/60Hz general-purpose IMs, since the product of $L_{ms}$ and $\omega_e$ is much smaller than $R_c$, the $\Delta TPA$ created by using the proposed MTPA control is minimal. For example, applying the parameters of a 1.5HP 60Hz IM in (4.74), the theoretical $\Delta TPA$ is only 1.5%. However, as $\omega_e$ increases or $R_c$ decreases, the $\Delta TPA$ becomes continuously more significant. Therefore, it indicates that the proposed MTPA control will be especially useful for certain types of IMs, even they may be available in the future, such as high-speed IMs and IMs of relatively small $R_c$ values (relatively large core loss).

![Graph](image.png)

Fig. 71. The torque ratio $K$ with respect to different $R_c$ and $\omega_e$
4.3.3 The Applicable Range of the Proposed MTPA Control

Similar to the conventional MTPA control, the proposed MTPA control also cannot be applied to the full range of load torque \( T_L \) due to flux saturation of the machine. Other control methods, such as IFOC of rated rotor flux, needs to replace the proposed MTPA control once \( T_L \) is too high. In the rotor-flux-oriented condition

\[
\lambda_{qr}^{\prime\prime} = 0, \tag{4.75}
\]

\[
\lambda_{dr}^{\prime\prime} = L_{m} \dot{I}_{dr}^{\prime\prime}, \tag{4.76}
\]

\[
\lambda_{Mag}^{\prime\prime} = \sqrt{\lambda_{qr}^{\prime\prime2} + \lambda_{dr}^{\prime\prime2}} = \lambda_{dr}^{\prime\prime}. \tag{4.77}
\]

Then applying (6.76), (6.77) and (6.59) in (6.53), the \( T_e \) of the proposed MTPA control relates to the magnitude of the rotor flux by

\[
T_e = \frac{3}{2} \frac{P}{2} \frac{\lambda_{Mag}^{\prime\prime2}}{L_{rr}^{\prime\prime}}. \tag{4.78}
\]
Considering that \( T_L \) is equal to \( T_e \) in steady state, thus the proposed MTPA control can support \( T_L \) less than the critical value that is calculated from rated rotor flux,

\[
T_{L\_critical} = \frac{3}{2} \frac{P \lambda_{rMag\_rated}^2}{L_{rr}}. \tag{4.79}
\]

Assuming that IFOC of rated flux replaces the proposed MTPA control when \( T_L \) is larger than \( T_{L\_critical} \), the general changes of \( \omega_s \) and \( \lambda_{rMag} \) in a full range of \( T_L \) are shown in Fig. 73.

![Fig. 73. The change of slip frequency and rotor flux magnitude with load torque](image)

4.3.4 Design of the Improved MTPA Controller

Based on the previous derivation, the current regulation of \( i_{qs}^* \) and \( i_{ds}^* \) in the conventional MTPA controller needs to be modified. In the proposed MTPA controller, \( i_{qs}^* \) and \( i_{ds}^* \) commands need to form a new pattern so that (4.59) can be satisfied. The proposed MTPA controller is shown in Fig. 74. The outer speed loop sets up the torque reference which generates identical \( i_{qs}^\hat{e} \) and \( i_{ds}^\hat{e} \) based on (4.53) and the proposed MTPA condition. Then, the corresponding \( i_{qs}^e^* \) and \( i_{ds}^e^* \) are calculated based on (4.56) and (4.57) for current regulation. The outputs of the current regulation are fed into two PI controllers to obtain the voltage references for the voltage-source inverter. On the other hand, \( \omega_s \) is set to the inverse of \( \tau_i \) as shown in (4.61). Note that
(4.55) is the requirement for rotor-flux-oriented control, while (4.61) is a more strict subtype of (4.55) for the proposed MTPA control. The conventional MTPA controller and the IFOC controller with fixed flux are shown in Fig. 75 and Fig. 76, respectively.

4.3.5 Simulation Verification and Results

To verify the proposed MTPA control, the proposed and conventional MTPA controllers, and the IFOC controller of rated flux are simulated and compared in MATLAB/Simulink. The
overall Simulink model is shown in Fig. 77, where the three controllers are shown in three subsystems of different colors. The IM is built based on the core-loss model, where the model parameters are extracted from the previous general-purpose 1.5 HP IM. The extracted $R_c$ is 530 $\Omega$. However, since $\Delta TPA$ increases with the decrease of $R_c$ as shown in Fig. 72, the $R_c$ is re-assigned to 200 $\Omega$ for a hypothetical IM to show more noticeable results. $I_{mag}$ at the same $T_L$ are compared using the three controllers. A stair-wise $T_L$ is applied to the machine operating at 1800 RPM, which is shown in Fig. 78. Theoretically, when supporting the same $T_L$ that is less than $T_{L, \text{critical}}$, the ratio of the current magnitudes by using the proposed and conventional MTPA control is

$$\frac{I_{mag, \text{proposed}}}{I_{mag, \text{conventional}}} = \frac{R_c^2 - (L_{mag} \omega_e)^2}{R_c^2 + (L_{mag} \omega_e)^2},$$

(4.80)

which can be derived from the previous analysis.

![Simulink model](image)

**Fig. 77.** Simulink model to compare the proposed and conventional MTPA controllers, and the IFOC controller

Several performance indices of the machine using the three controllers are compared in Fig. 79–Fig. 83. In Fig. 79, all the three controllers show good torque tracking that can support
the \( T_L \) well at different torque levels. In Fig. 80, as expected, \( \lambda_{rMag}' \) is constant with respect to \( T_L \) when using IFOC controller, whereas \( \lambda_{rMag}' \) changes in the same trend as \( T_L \) when using the proposed and conventional MTPA controllers. Moreover, \( \lambda_{rMag}' \) of the conventional MTPA controller is larger than using the proposed MTPA controller. Thus, for the third \( T_L \) in Fig. 79, which is close to but less than \( T_{L,\text{critical}} \), the conventional MPTA control starts to saturate the machine while the proposed MTPA control does not. On the other hand, the \( I_{mag} \) using the three controllers are shown in Fig. 81. For the same \( T_L \), the proposed MTPA controller requires least current, which verifies that the proposed MTPA control has higher TPA capability than the conventional MTPA control by considering core loss in the control design. Compared to the other two controllers, the IFOC controller asks for more current especially at very-low-torque conditions. This is due to the excessively redundant flux in the IFOC-controlled machine at low-load conditions, as shown in Fig. 80. The machine efficiencies of using the three controllers are shown in Fig. 82 and we only consider the efficiency at steady-state conditions. It shows that the proposed MTPA control leads to the highest machine efficiency, then followed by the conventional MTPA control. Moreover, the efficiency is significantly decreased in the IFOC and the decreasing rate depends on the redundant degree of the flux. In summary, due to more accurate estimation of machine losses, the proposed MTPA control shows higher TPA ratio and efficiency than those of the conventional MTPA control.

On the other hand, since no regulation is implemented to the machine flux in the IFOC of rated flux, it is expected that the dynamic performance of the IFOC controller is better than the proposed and conventional MTPA controllers. An example is shown in Fig. 83, which compares the transient response of \( T_e \) at 15s using the three controllers with the change of \( T_L \). The IFOC controller has no overshoot and the fastest regulation. But the performance of the MTPA
controllers are decent. Moreover, the proposed MTPA controller has almost the same overshoot but faster regulation than the conventional MTPA controller.

Fig. 78. The $T_L$ used in the simulation for all the three controllers

Fig. 79. The induced torque $T_e$ by the three controllers

Fig. 80. The magnitude of machine rotor flux using the three controllers
4.4 Summary

This chapter presents advanced model-based control to improve machines’ efficiency and performance (MTPA capability). The model-based LMC of IMs using the proposed core-loss model is first discussed for IFOC and V/f control. The closed-form solution for the optimal flux in IFOC is derived, whereas the optimal flux in V/f control is determined through numerical
sweep. Significant efficiency enhancement or loss reduction is observed using the designed LMC at low-load conditions, e.g. $T_L < T_{L_{\text{rated}}}$. The room for LMC is larger at lower $T_L$. Compared to the conventional copper-loss model, the proposed core-loss model can lead to further efficiency enhancement or loss reduction at low-load conditions when core loss is comparative to copper loss in magnitude.

The power electronic losses of inverters are derived and integrated with the core-loss model as well as the models of mechanical and stray losses in this chapter. The system-level power loss model is used to design LMC, using numerical sweep to find the optimal flux, for inverter-fed motor drive systems with IFOC and V/f control. Significant efficiency enhancement or loss reduction compared to using rated flux is observed again. Moreover, due to the integration of inverter losses, the optimal flux is shifted from the one decided by only machine losses. Experimental validation of the designed system-level LMC is carried out in V/f condition. More than 13% loss reduction is achieved in the experiment when the 1.5 HP motor is operated at 1800 RPM with 1.2 N·m and 1200 RPM with 0.6 N·m.

The core-loss model is also used to improve machines’ performance, e.g. MTPA, in this chapter. An improved model-based MTPA control is derived by considering core losses, which could lead to higher TPA ratio and efficiency than the conventional MTPA control. The increase of TPA ratio using the proposed and conventional MPTA control depends on $\omega_e$ and $R_c$. Possible applications of the proposed MPTA control are high-speed IMs and IMs with relatively large core loss (or small $R_c$).
CHAPTER 5
ADVANCED MODEL-BASED FAULT DIAGNOSIS AND FAULT-TOLERANT
CONTROL OF THREE-PHASE IM DRIVE SYSTEMS

5.1 Adaptive Modulation Time-Domain Fault Diagnosis of Three-Phase IMs

As introduced in Chapter 2, different IM faults can randomly happen due to constant wear and tear, heavy and changing loads, enduring operation, possible harsh environments and many other factors. Therefore, fault detection and diagnosis (FDD) is important for operating IMs, especially for safety-critical applications such as electric vehicles, elevators and escalators. A simple time-domain FDD method is proposed, which can detect all the four major types of IM faults: air-gap eccentricity fault (EF), bearing fault (BF), broken rotor bar fault (BRBF) and stator short winding fault (SSWF). The proposed FDD method requires minimal signal processing and no additional sensors except those already in machine drives for control purposes. No frequency-domain calculations, wavelet processing, machine learning, or other more complex algorithms is involved. Thus, the method can facilitate hardware implementation on devices of limited memory and processing capabilities. The proposed FDD method shows excellent fault sensitivity and robustness to noise and current harmonics for all the four major types of IM faults.

5.1.1 Methodology

5.1.1.1 Adaptive Modulation

The proposed method uses artificial modulating signals, which have the theoretical model-based fault frequencies shown in TABLE II, to modulate the stator current feedback. The current feedback signal into the controller is not disturbed, and modulation occurs in a side loop
that is non-intrusive. The method is online and self-adaptive to the change of speed and torque, as shown in Fig. 84. First, consider only the fundamental and fault signal in the current feedback. Assume the stator current fundamental (subscript \( e \)), fault signal (subscript \( f \)) and the artificial modulating signal (subscript \( md \)) are

\[
y_e(t) = A \sin(\omega_e t + \theta_{e0}), \tag{5.1}
\]

\[
y_f(t) = a \sin(\omega_f t + \theta_{f0}), \tag{5.2}
\]

\[
y_{md}(t) = X \sin(\omega_{md} t + \theta_{md0}), \tag{5.3}
\]

where \( A, a \) and \( X \) are the magnitudes of the frequency components. \( A \) is significantly larger than \( a \) to validate the proposed FDD method even when the fault signature is very small. \( \omega \) is the frequency in rad/s and \( \theta \) is the phase with subscript 0 indicating the initial time point. \( \omega_e \) is the synchronous frequency as before. \( \omega_{md} \) has the value of the theoretical fault frequency and is equal to \( \omega_f \) in an ideal case. However, in reality, \( \omega_{md} \) is always slightly different from \( \omega_f \) due to various reasons such as machine vibration, fluctuation of feedback signals, instrument inaccuracy, inaccurate machine parameter knowledge, and intrinsic inaccuracy of the estimator that is derived from simplified machine models, etc. The adaptive modulation on the current feedback results in

\[
y(t) = [y_e(t) + y_f(t)] \times y_{md}(t)
\]

\[
= \frac{AX}{2} \cos\left[\left(\omega_e - \omega_{md}\right) t + (\theta_{e0} - \theta_{md0})\right]
\]

\[
- \frac{AX}{2} \cos\left[\left(\omega_e + \omega_{md}\right) t + (\theta_{e0} + \theta_{md0})\right],
\tag{5.4}
\]

\[
+ \frac{aX}{2} \cos\left[\left(\omega_f - \omega_{md}\right) t + (\theta_{f0} - \theta_{md0})\right]
\]

\[
- \frac{aX}{2} \cos\left[\left(\omega_f + \omega_{md}\right) t + (\theta_{f0} + \theta_{md0})\right]
\]
where $y(t)$ is the modulation resultant that is also referred as modulated signal, and it contains four frequency components. The third component in (5.4), which has the frequency of $(\omega_f - \omega_{md})$, is selected as the fault-indicative component, since 1) it appears only when the fault occurs, and 2) this component is the slowest signal, ideally DC, in (5.4), and its frequency is far away from other components. No such extremely slow component exists when there is no fault except at uncommon near-zero speed. As for the fourth component that also contains $\omega_f$, it is not utilized since the sum of $\omega_f$ and $\omega_{md}$ may be confused by other frequencies.

Fig. 84. Adaptive modulation and signal processing

5.1.1.2 Signal Processing

The goal of signal processing is to attenuate relatively high-frequency fault-irrelevant components so that the fault-indicative component is more distinctive. To achieve this goal, the proposed method uses an average or Mean function, a low-pass filter (LPF), and a notch filter for the $(\omega_e - \omega_{md})$ component in (5.4), as shown in Fig. 84. Note that $\omega_e$ is obtained using phase-lock loop (PLL) on the current feedback, and $\omega_{md}$ is known. Moreover, no matter how significant the attenuation of the fault-irrelevant components is, the remnants of these components can still be relatively large and thus destructive compared to a small fault magnitude $a$. Therefore, it is expected that the proposed method, like other FDD methods, has limited fault sensitivity. This
limitation depends on the relative amplitudes of the fault-irrelevant components and their relative separation on the frequency spectrum from the fault-indicative component. It also depends on the performances of the Mean function and the filters. Generally, smaller and further fault-irrelevant components along with sharper filters can increase the fault sensitivity. The three typical types of modulated signals after the signal processing are shown in Fig. 85, which are no-fault-indicative component, small fault-indicative component with large ripple, and large fault indicative component with small ripple, respectively.

Fig. 85. Different types of processed signals: a) no fault-indicative component; b) small fault-indicative component with relatively large ripple; c) large fault-indicative component with relatively small ripple
5.1.1.3 Fault Detection and Diagnosis

The previous signal processing results in a slow sinusoidal signal coupled with higher-frequency ripples when a fault happens. Otherwise, the signal has only high-frequency ripple centered around zero. Since the fault-indicative signal is the slowest component of the modulated signal, this component can be detected by checking the zero-crossing points of the modulated signal after processing. Assume that the time between the simulation or experimental runtime and the instance of the last zero-crossing point is $T_i$. An example of $T_i$ for different runtimes of a fault-indicative signal is shown in Fig. 86, and $T_i$ for a no-fault signal is shown in Fig. 87. Clearly, the possible maximum $T_i$, which is about half period of the slowest frequency component, is much larger when a fault occurs. Therefore, $T_i$ is continuously monitored during simulation or experimentation. If $T_i$ is larger than a pre-defined threshold $T_{i\text{-th}}$, it means that the very-low-frequency fault-indicative component has been detected. Otherwise, no fault is flagged. $T_{i\text{-th}}$ can be set by experience. For example, if $T_{i\text{-th}}$ is set to 0.5 s, then any real signal that is within 1 Hz from the estimated fault frequency would flag a fault. However, note from Fig. 85(b) that signal ripples will significantly decrease the maximum $T_i$ when the fault magnitude is insufficient. This sets the fault sensitivity limit of the proposed method.

![Fig. 86. $T_i$ of a fault-indicative signal at runtime of 10.2s (a) and 10.7s (b)](image-url)
In addition to the fundamental component, real stator current feedback also contains harmonics, noise and possibly a small DC offset. These practical factors add additional ripples after signal processing, but they also can be greatly attenuated by the Mean function and LPF.

Fig. 87. $T_t$ of a no-fault signal ($T_t$ here is about 0.01s, very small)

5.1.2 Simulation Verification

A MATLAB/Simulink model is built to test the proposed fault detection method in simulation, which is shown in Fig. 88. This is a general IM drive system that is equipped with fault detection capability using the proposed FDD method. The IM is controlled using IFOC combined with current hysteresis control to generate switching signals for the inverter. The conventional copper-loss model of IMs is used. Moreover, a fan-type load is used, which determines $T_L$ based on the speed feedback. In the Fault Detection block, the adaptive modulation is applied to the current feedback, where the artificial modulating signal is generated based on the speed feedback, current feedback (for PLL) and equations in TABLE II. All the four major types of faults are tested in the simulation verification.

The subsystem of the Fault Detection block is shown in Fig. 89, where the BRBF is used as an example. By running a MATLAB m-file in advance, the integer $q$ is randomly assigned either a zero or one, which determines whether an additional frequency component (not
necessarily a fault) is injected \((q=1)\) or not \((q=0)\) in the current feedback. Moreover, a random drift frequency \(d\), which decides whether the injected frequency component is a fault component or not, is added to the theoretical fault frequency. For example, we choose \(T_{t,th}\) to be 0.5s and \(d\) to be a random number between -2Hz to 2Hz. If \(d\) is within +/-1Hz, then the injected frequency component is a fault component, since \(T_{t,th}\) is 0.5s. Otherwise, the additional frequency component is not a fault component. Such setting of \(q\) and \(d\) is used to check whether the proposed method will flag a false alarm when a suspicious fault-irrelevant signal exists. On the other side, the modulated signal goes through a moving Mean function, a notch filter and a LPF to attenuate fault-irrelevant components. As stated before, the Mean function and the filters will affect the fault sensitivity of the proposed method. Here, we select the fundamental frequency of the Mean function to be half of \(\omega_e\) and a second-order LPF. The \(Q\) factor of the notch filter and the LPF are 0.707 for critical damping. Finally, the latest zero-crossing point is detected and compared with the runtime of the simulation. \(T_i\) and \(T_{t,th}\) are compared continuously and the flag value is saved in the variable \(r\) \((1\) for fault, \(0\) for no fault). The correctness of the proposed method is determined by inspecting the \(q, d, r\) values, which is shown in Fig. 90.

![High-level block diagram of the simulation verification of the proposed FDD method](image)

Fig. 88. High-level block diagram of the simulation verification of the proposed FDD method
The proposed FDD method is verified for all the four major types of IM faults at different speeds, fault magnitudes, noise and DC offsets. $T_{c,th}$ and the maximum/minimum value of $d$ are selected to be 0.5 s and +/- 2 Hz, respectively. The results are shown in TABLE XV to TABLE XVIII. The FDD correctness in each cell of these tables are obtained through 300 simulation runs showing in Fig. 90.

![Diagram of Fault Detection Block](image)

**Fig. 89.** The subsystem of the Fault Detection block in Fig. 88

![Flowchart](image)

**Fig. 90.** Decision of FDD correctness in the simulation verification
It is shown in TABLE XV and TABLE XVIII that the proposed FDD method is applicable to different faults at various speed and $T_L$ (fan-type load) conditions, and has excellent fault sensitivity. Moreover, the method is robust to current and speed noises, current harmonics and small DC offset in feedback. Generally, the successful fault detection rate increases with the increase of the fault magnitude and the decrease of the fault-irrelevant interferences. Note that, since the successful detection rate is a statistical result obtained from the 300 simulation runs for each condition, the rate is not perfectly accurate unless the number of simulation runs is significantly large, or ideally infinite. Thus, there are several small rebounds in the tables that do not follow the major trend. Moreover, since different faults have different characteristic frequencies, their interaction with the same signal processing system results in different relative attenuation of the fault-indicative component and fault-irrelevant components. This leads to different successful fault detection rates for different types of faults of the same magnitude.

**TABLE XV. FAULT DETECTION CORRECTNESS WITHOUT NOISE AT 1800 RPM**

<table>
<thead>
<tr>
<th>Fault Magnitude w.r.t Current Fundamental</th>
<th>5%</th>
<th>2%</th>
<th>1%</th>
<th>0.5%</th>
<th>0.2%</th>
<th>0.1%</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{ef1}$</td>
<td>99.7</td>
<td>99.7</td>
<td>99.3</td>
<td>97.7</td>
<td>94.3</td>
<td>74.3</td>
</tr>
<tr>
<td>$f_{ef2}$</td>
<td>99.3</td>
<td>99.3</td>
<td>99.3</td>
<td>99.0</td>
<td>97.3</td>
<td>98.0</td>
</tr>
<tr>
<td>$f_\varepsilon$</td>
<td>99.0</td>
<td>100</td>
<td>99.0</td>
<td>98.0</td>
<td>97.3</td>
<td>93.0</td>
</tr>
<tr>
<td>$f_{sr}$</td>
<td>99.7</td>
<td>99.3</td>
<td>99.0</td>
<td>99.3</td>
<td>98.3</td>
<td>95.0</td>
</tr>
<tr>
<td>$f_{brhf}$</td>
<td>100</td>
<td>99.0</td>
<td>99.3</td>
<td>98.7</td>
<td>97.7</td>
<td>98.0</td>
</tr>
<tr>
<td>$f_{swf}$</td>
<td>100</td>
<td>99.0</td>
<td>99.3</td>
<td>99.3</td>
<td>98.0</td>
<td>94.0</td>
</tr>
</tbody>
</table>

5.1.2 Experimental Verification

The previous dSPACE platform is used for the experimental verification of the proposed FDD method. The verification uses real speed and current feedback information in creating modulating signals, while the fault signal is not from a broken motor, but is virtually generated.
through real feedback signals. On the other hand, V/f control is used instead IFOC for easy control purpose. The constructed virtual panel in dSPACE ControlDesk is shown in Fig. 91. On this virtual panel, the speed and V/f ratio are set to control the IM. Different faults, fault magnitudes, magnitudes of the artificial modulating signal and drift frequencies can be tested in the verification. Moreover, important information can be read from the virtual panel, such as current and speed feedback, PLL output, theoretical and “real” fault frequencies, modulated signals before and after signal processing, and most importantly, the fault flag.

**TABLE XVI. Fault Detection Correctness without Noise at 600 RPM**

<table>
<thead>
<tr>
<th>Fault Magnitude w.r.t Current Fundamental</th>
<th>5%</th>
<th>2%</th>
<th>1%</th>
<th>0.5%</th>
<th>0.2%</th>
<th>0.1%</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_{ef1} )</td>
<td>99.7</td>
<td>99.7</td>
<td>99.7</td>
<td>99.7</td>
<td>99.7</td>
<td>97.0</td>
</tr>
<tr>
<td>( f_{ef2} )</td>
<td>99.7</td>
<td>99.3</td>
<td>99.3</td>
<td>99.3</td>
<td>99.0</td>
<td>98.0</td>
</tr>
<tr>
<td>( f_r )</td>
<td>99.3</td>
<td>95.7</td>
<td>97.0</td>
<td>93.0</td>
<td>79.7</td>
<td>76.3</td>
</tr>
<tr>
<td>( f_{or} )</td>
<td>99.3</td>
<td>99.7</td>
<td>99.7</td>
<td>99.0</td>
<td>99.3</td>
<td>99.3</td>
</tr>
<tr>
<td>( f_{brbf} )</td>
<td>99.3</td>
<td>99.7</td>
<td>99.0</td>
<td>99.3</td>
<td>99.3</td>
<td>99.0</td>
</tr>
<tr>
<td>( f_{sswf} )</td>
<td>99.7</td>
<td>99.7</td>
<td>99.7</td>
<td>99.7</td>
<td>99.3</td>
<td>96.7</td>
</tr>
</tbody>
</table>

**TABLE XVII. The Fault Detection Correctness with Noise at 1800 RPM and 0.5% Fault Magnitude (S: speed; C: current)**

<table>
<thead>
<tr>
<th>Fault Magnitude</th>
<th>0.75% (S), 2% (C) Noises, No DC Offset</th>
<th>1.5% (S), 5% (C) Noises, No DC Offset</th>
<th>3% (S), 5% (C) Noises, No DC Offset</th>
<th>3% (S), 10% (C) Noises, 5% DC Offset</th>
<th>3% (S), 10% (C) Noises, 5% DC Offset</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_{ef1} )</td>
<td>99.0</td>
<td>99.7</td>
<td>98.7</td>
<td>99.0</td>
<td></td>
</tr>
<tr>
<td>( f_{ef2} )</td>
<td>98.0</td>
<td>97.7</td>
<td>98.3</td>
<td>98.3</td>
<td></td>
</tr>
<tr>
<td>( f_r )</td>
<td>86.0</td>
<td>83.7</td>
<td>78.3</td>
<td>76.3</td>
<td></td>
</tr>
<tr>
<td>( f_{or} )</td>
<td>98.7</td>
<td>98.3</td>
<td>98.3</td>
<td>98.0</td>
<td></td>
</tr>
<tr>
<td>( f_{brbf} )</td>
<td>97.7</td>
<td>95.7</td>
<td>95.7</td>
<td>95.3</td>
<td></td>
</tr>
<tr>
<td>( f_{sswf} )</td>
<td>99.7</td>
<td>98.3</td>
<td>97.7</td>
<td>97.7</td>
<td></td>
</tr>
</tbody>
</table>
The experimental verification is performed by checking the Flag signal on the virtual panel after the Fault Trigger is enabled. The default value of the Flag is zero, and it changes to one when a fault is detected. $T_{th}$ of 0.5s is selected in the hybrid experimental verification, which means any large enough frequency component that is less than +/-1Hz away from the theoretical fault frequency will flag a fault. It also means the responding time of the fault detection is about 0.5s (small fault magnitude may increase the responding time). In other words, the fault will be detected about 0.5s after the fault occurs, as the model must wait at least $T_{th}$ time before flagging a fault. An example of fault detection is shown in Fig. 92. The bottom signal is the Fault Trigger, which is arbitrarily enabled at around 13.1s when its value changes from zero (disabled) to one (enabled). After about 0.5s, the Flag signal changes from zero to one which indicates catching a fault.

**TABLE XVIII. The Fault Detection Correctness with Noise at 600 RPM and 0.5% Fault Magnitude (S: speed; C: current)**

<table>
<thead>
<tr>
<th></th>
<th>0.75% (S), 2% (C)</th>
<th>1.5% (S), 5% (C)</th>
<th>3% (S), 10% (C)</th>
<th>3% (S), 10% (C)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Noises, No DC Offset</td>
<td>Noises, No DC Offset</td>
<td>Noises, No DC Offset</td>
<td>Noises, 5% DC Offset</td>
</tr>
<tr>
<td>$f_{ef1}$</td>
<td>99.3</td>
<td>95.7</td>
<td>93.0</td>
<td>90.0</td>
</tr>
<tr>
<td>$f_{ef2}$</td>
<td>97.7</td>
<td>96.7</td>
<td>95.7</td>
<td>93.7</td>
</tr>
<tr>
<td>$f_{fr}$</td>
<td>94.7</td>
<td>94.7</td>
<td>90.3</td>
<td>90.0</td>
</tr>
<tr>
<td>$f_{fe}$</td>
<td>96.0</td>
<td>97.7</td>
<td>93.3</td>
<td>95.7</td>
</tr>
<tr>
<td>$f_{bhf}$</td>
<td>98.7</td>
<td>94.7</td>
<td>93.3</td>
<td>91.7</td>
</tr>
<tr>
<td>$f_{swf}$</td>
<td>100</td>
<td>97.0</td>
<td>91.7</td>
<td>95.3</td>
</tr>
</tbody>
</table>
5.2 High-Performance Synchronous-Frame Multi-Controller Drive for Fault-Tolerant Control of IMs with Sensor Failures

A high-performance synchronous-frame multi-controller IM drive is proposed to deal with sudden sensor failures in closed-loop controllers. This study was initially collaborated with Michael Stettenbenz, who is a master student in APEDL, and preliminary results were published in [118]. In this dissertation, the improvement from the initial work is presented. The direct
torque control (DTC) and IFOC controllers have been improved and much better simulation results are obtained. Detailed illustrations of each controller and the constructed drive are provided. Both simulation and experimental verification of the proposed drive are given. Unlike other multi-controller drives in literature, the proposed drive does not require monitoring or controlling the phase of the rotor flux, or any other phase, to smooth hand-off transients when switching between controllers. The drive intrinsically satisfies the requirement of synchronization of different controllers. The smooth switching can be achieved using a simple rate limiter on the synchronous-frame voltage commands. Moreover, the proposed drive and control can be equally used in sensor failure and recovery conditions. The high-level block diagrams of the proposed synchronous-frame drive and the drive using conventional abc-frame switching are shown in Fig. 93 and Fig. 94, respectively.

Note that the simpler copper-loss model of three-phase IMs is used to design the proposed drive. Therefore, although the same variables from the core-loss model are used in this sub-section, the voltage, current, flux, torque and speed relationships of these variables may change from those in the core-loss model.

Fig. 93. The proposed synchronous-frame multi-controller drive
5.2.1 Synchronous-Frame Controllers in the Proposed Multi-Controller Drive

Three of the most common controllers for IMs, DTC, IFOC and V/f controllers, are integrated in the proposed drive. The hierarchy of the three controllers is shown in Fig. 95 along with the switching criteria based on sensor failure or recovery. The proposed drive uses speed encoder, stator current, and voltage sensors. The DTC controller is executed when all the sensors are available, then the IFOC controller is engaged in the case of voltage sensor failure. V/f controller is used as the last resort to keep the continuity of operation if neither of the closed-loop controllers can be used. Note that the DTC and IFOC controllers can be interchanged in Fig. 95 with corresponding adjustment of sensor information. However, the present hierarchy of controllers is applied for illustration convenience based on the number of sensors in each controller. The same smooth hand-off switching approach can be applied if the DTC and IFOC controllers are interchanged. The conventional DTC and IFOC controllers are modified to provide \( v_{qs}^e \), \( v_{ds}^e \). Then, \( v_{qs}^e \) and \( v_{ds}^e \) are used to generate \( v_{abc}^e \) for sinusoidal PWM (SPWM) generator, which uses synchronous-frame phase information, \( \theta_e^e \), from each controller.
5.2.1.1 Synchronous-frame Closed-Loop DTC Controller

The conventional DTC controller directly regulates a machine’s torque and flux using a switching table, which is based on stator flux and torque feedback. DTC can achieve fast torque response without requiring field orientation, but it generally suffers from relatively large torque ripple due to the embedded torque hysteresis control. The DTC-SVPWM control method uses the space vector PWM (SVPWM) technique to replace the switching table and decrease the machine’s torque ripple. Both conventional DTC and DTC-SVPWM are implemented in the stationary frame. A modified synchronous-frame DTC controller is shown in Fig. 96. This DTC controller requires three-phase voltage and current feedbacks \( (v_{abcs}, i_{abcs}) \) for flux and torque estimation. Moreover, it requires speed feedback for speed control of the machine. \( v_{abcs} \) and \( i_{abcs} \) are first transformed into stationary frame to get the corresponding \( d \)- and \( q \)-axis voltages and currents \( (v_{ds}, v_{qs}, i_{ds}, i_{qs}) \). Then, the stationary-frame fluxes \( (\lambda_{ds}, \lambda_{qs}) \) can be estimated from

\[
\lambda_{ds} = \int (v_{ds} - R_e \cdot i_{ds}) \quad (5.5)
\]

\[
\lambda_{qs} = \int (v_{qs} - R_e \cdot i_{qs}) \quad (5.6)
\]

where the superscript \( s \) represents stationary frame. Moreover, \( T_e \) can be estimated from

\[
T_e = \frac{3P}{4} \left( \lambda_{ds} i_{qs} - \lambda_{qs} i_{ds} \right) \quad (5.7)
\]
In Fig. 96, $T_e$ is compared with the $T_e^*$ (ideally, $T_e^*$ is equal to $T_L$), which is generated from the speed control loop via a PI controller. The comparison gives the desired slip frequency, $\omega_{s\_DTC}^*$, which is used to obtain the desired synchronous-frame frequency and phase, $\omega_{e\_DTC}^*$ and $\theta_{e\_DTC}^*$, in the DTC controller. On the other hand, the magnitude of the stator flux command, $|\lambda_s|^*$, is set to rated value. Then, $|\lambda_s|^*$ and $\theta_{e\_DTC}^*$ are used to generate the $d$- and $q$-axis stator flux commands in the stationary frame

$$\lambda_{ds\_DTC}^* = -|\lambda_s|^* \sin(\theta_{e\_DTC}^*), \quad (5.8)$$

$$\lambda_{qs\_DTC}^* = |\lambda_s|^* \cos(\theta_{e\_DTC}^*). \quad (5.9)$$

Note that $q$-axis is used as the phase angle reference following the custom in [14]. Then, $\lambda_{ds\_DTC}^*$ and $\lambda_{qs\_DTC}^*$ are regulated through two PI controllers that generate $v_{ds\_DTC}^*$ and $v_{qs\_DTC}^*$, respectively. To achieve smooth switching between different controllers later, $v_{ds\_DTC}^*$ and $v_{qs\_DTC}^*$ are first transformed to $v_{ds\_e\_DTC}^*$ and $v_{qs\_e\_DTC}^*$ before they are eventually transformed to $v_{abcs\_e\_DTC}^*$ for SPWM generation.
5.2.1.2 Synchronous-frame Closed-Loop IFOC Controller

IFOC controllers are already designed in synchronous frame. The machine’s torque and flux are controlled separately through current regulation on $i_{qs}^e$ and $i_{ds}^e$, respectively. $\lambda_{qr}^e$ is set to zero in order to decouple the torque and flux control loops. Based on the conventional copper-loss model of IMs, the field orientation is achieved through proper adjustment of synchronous-frame phase that is calculated by

$$\theta_{e_{-IFOC}}^e = \int \left( \frac{R_r^e}{L_m^e} i_{qs}^e + \omega_e \right).$$

(5.10)

Many conventional IFOC controllers use current hysteresis control to regulate $i_{qs}^e$ and $i_{ds}^e$, and to generate non-PWM switching signals. However, to achieve the smooth hand-off transient between different controllers, a voltage-type IFOC controller is used and shown in Fig. 97.

In Fig. 97, the magnitude of rotor flux, $|\lambda_r^e|$ (equal to $\lambda_{dr}^e$ since $\lambda_{qr}^e$ is forced to zero), is set to rated value, which is used to determine $i_{ds}^{e*}$,

$$i_{ds}^{e*} = \frac{|\lambda_r^e|}{L_{m}^e},$$

(5.11)

On the other hand, the speed control loop sets up the desired $T_e^*$ through a PI controller, which is used to calculate $i_{qs}^{e*}$,

$$i_{qs}^{e*} = \frac{4L_{te}^e}{3PL_{m}^2} \frac{T_e^*}{i_{ds}^{e*}}.$$

(5.12)

Then, the current regulation of $i_{ds}^e$ and $i_{qs}^e$ are achieved using PI controllers instead of current hysteresis loops, to result in $v_{ds_{-IFOC}}^{e*}$ and $v_{qs_{-IFOC}}^{e*}$. 

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5.2.1.3 Synchronous-frame Open-Loop Vf Controller

The common V/f controller already satisfies the requirement for the smooth hand-off transition, which is shown in Fig. 98. Compared to the previous two closed-loop controllers, the open-loop V/f controller is much simpler and does not require any feedback signals, but its response is relatively sluggish due to coupled torque and flux control. Moreover, it lacks speed regulation and reference-tracking accuracy. The synchronous frequency, $\omega_{e\_vf}$, is calculated directly from the speed command, which is shown in (4.22) but is repeated here for convenience.

$$
\omega_{e\_vf} = \frac{2\pi}{60} \cdot \frac{P}{2} \cdot \text{Spd}^*,
$$

where it is assumed that $\omega_{e\_vf}$ and $\omega_r$ are close enough. The Vf* is assigned to rated value. $v_{ds\_vf}^*$ is set to zero and $v_{qs\_vf}^*$ can be calculated by

$$
v_{qs\_vf}^* = \sqrt{2} \cdot \frac{P}{60} \cdot \frac{2}{\text{Spd}^*} \cdot \text{Vf}^*.
$$

$v_{ds\_vf}^*$ and $v_{qs\_vf}^*$ are again transformed to $v_{abcs}^*$ fed into the SPWM generator.
Note that all flux commands are set to rated values in this work, but they can be updated using other flux-determining algorithms if needed for efficiency enhancement, flux weakening, or other purposes. However, this point is not the focus of this work.

5.2.2 Synchronous-Frame Multi-Controller Drive with Smooth Transition Between Controllers

5.2.2.1 The Proposed Synchronous-Frame Multi-Controller Drive

As shown in Fig. 96–Fig. 98, the DTC, IFOC and V/f controllers are modified to have a similar structure in the synchronous frame: each controller has a speed command and a flux command or setpoint (in red) as inputs. Moreover, each controller has a pair of synchronous-frame voltage commands, $v_{ds}^*$ and $v_{qs}^*$ (in green), which can be transformed into $abc$-frame commands based on the corresponding synchronous speed and phase information, $\omega_e^*$ and $\theta_e^*$ (in blue). Therefore, these controllers can be used to build the proposed synchronous-frame multi-controller drive, which is shown in Fig. 93. The flux setpoints and feedback signals are avoided in Fig. 93 to highlight the main structure of the drive. Moreover, the “Sensor Condition Monitor” block provides command for controller selection depending on the sensor information (failure and recovery). Note that we do not include any detection of sensor conditions in this work, but readers are referred to [99], [119]. Instead, manual selection and switching of controllers are used to emulate sensor failure and recovery, since the focus of this paper is on the fault-tolerant capability and the response at hand-off transients after a fault happens. A simple “Rate Limiter”
on $v_{ds,qs}^*$ is used to achieve the smooth hand-off transition between different controllers, which is enabled only when a change of controller-selection command is detected.

5.2.2.2 Smooth Transition Between Different Controllers

The reason to modify all the controllers into the unified synchronous-frame form is to get DC-type voltage commands, $v_{ds}^*$ and $v_{qs}^*$, which are much easier to control compared to normal AC-type commands, $v_{abcs}^*$. To understand the large hand-off transients when switching between different controllers, a multi-controller drive using the same DTC, IFOC and V/f controllers as before is considered, but the controllers are switched in the normal abc-frame as shown in Fig. 94. Take the switching from the IFOC controller to the V/f controller as an example, and assume operation in steady state,

$$v_{ds_{-_IFOC}}^* = v_{ds_{-_Vf}}^* + \Delta v_{ds}^*,$$  \hspace{1cm} (5.15)

$$v_{qs_{-_IFOC}}^* = v_{qs_{-_Vf}}^* + \Delta v_{qs}^*,$$  \hspace{1cm} (5.16)

$$\theta_{e_{-_IFOC}}^* = \theta_{e_{-_Vf}}^* + \Delta \theta_e^*,$$  \hspace{1cm} (5.17)

where $\Delta v_{ds}^*$, $\Delta v_{qs}^*$ and $\Delta \theta_e^*$ are the differences of synchronous-frame $d$-, $q$-axis voltages and phases of the two controllers, respectively. Then, the corresponding $v_{as}^*$ out of the two controllers are (assuming balance condition where $\theta$-axis voltages are zero)

$$v_{as_{-_Vf}}^* = v_{qs_{-_Vf}}^* \cos(\theta_{e_{-_Vf}}^*) + v_{ds_{-_Vf}}^* \sin(\theta_{e_{-_Vf}}^*),$$  \hspace{1cm} (5.18)

$$v_{as_{-_IFOC}}^* = \left( v_{qs_{-_IFOC}}^* + \Delta v_{qs}^* \right) \cos(\theta_{e_{-_IFOC}}^* + \Delta \theta_e^*) + \left( v_{ds_{-_IFOC}}^* + \Delta v_{ds}^* \right) \sin(\theta_{e_{-_IFOC}}^* + \Delta \theta_e^*).$$  \hspace{1cm} (5.19)

Therefore, when switching from the IFOC controller to the V/f controller, the change of $v_{as}^*$ is

$$\Delta v_{as}^* = v_{as_{-_IFOC}}^* - v_{as_{-_Vf}}^*$$

$$= v_{qs_{-_Vf}}^* \left[ \cos(\theta_{e_{-_IFOC}}^*) - \cos(\theta_{e_{-_Vf}}^*) \right] + v_{ds_{-_Vf}}^* \left[ \sin(\theta_{e_{-_IFOC}}^*) - \sin(\theta_{e_{-_Vf}}^*) \right] \right.$$  

$$+ |\Delta v_s^*| \sin(\theta_{e_{-_IFOC}}^* + \gamma)$$  \hspace{1cm} (5.20)
where

\[ |\Delta v_s^*| = \sqrt{\Delta v_{ds}^{e*2} + \Delta v_{qs}^{e*2}}, \]  

(5.21)

\[ \gamma = \tan^{-1} \left( \frac{\Delta v_{qs}^{e*}}{\Delta v_{ds}^{e*}} \right). \]  

(5.22)

The first two terms on the right side of (5.20) change with the phase difference between \( \theta_{e_{IFOC}}^* \) and \( \theta_{e_{vf}}^* \) at the controller hand-off instance, while the third term changes with the magnitude of \( \Delta v_{ds}^{e*} \) and \( \Delta v_{qs}^{e*} \). All the three terms periodically change with time and thus \( \Delta v_{as}^{e*} \) changes with the hand-off instance. Similar changes happen to \( v_{bs}^{e*} \) and \( v_{cs}^{e*} \) as well. The instantaneous change of the voltage commands is the reason for large transient responses.

The proposed synchronous-frame multi-controller drive can alleviate the sudden changes in \( v_{abcs}^{e*} \) and the corresponding large transient responses. In Fig. 93, although \( \omega_e^{e*} \) changes when switching controllers, \( \theta_e^{e*} \) does not change instantaneously due to the integration operation. In other words, the phase of the original controller is inherited by the controller after switching. Therefore, the synchronization of controllers at the hand-off instance is intrinsically and perfectly satisfied. Correspondingly, the first two terms on the right side of (5.20) are zero at the controller hand-off instance. On the other hand, the simple “Rate Limiter” can restrict the instantaneous value of \( |\Delta v_s^*| \) at the controller hand-off instance. This decreases the magnitude of the third term on the right side of (5.20). Note that a simple rate limiter can be used here to achieve the smooth hand-off transition due to the advantages of DC-type voltage commands in the synchronous frame. The rate limiter will not work for AC-type voltage commands.

5.2.2.3 Switching to A Closed-Loop Controller

To minimize “dwell time” at controllers’ hand-off transient, the three controllers in the proposed drive are set to be concurrently active, but only one controller is executed. Thus, the PI
controllers in the nonexecuted DTC and/or IFOC controller will keep regulating, but will not succeed. As a result, the outputs of the PI controllers at the hand-off instance could be very large and different from the steady-state values that they should be. This fact could therefore deteriorate hand-off transients when switching to a closed-loop controller using the proposed drive. To solve this problem, resetting PI controllers is needed. For PI controllers generating $T_e^*$, the initial value is reset to zero at the hand-off instance. For PI controllers generating voltage commands, the initial values are reset to be the values of the same variables from the controller before switching. Specifically, for example, when switching from the V/f controller to the IFOC controller, the values of $v_{ds,vf}^*$ and $v_{qs,vf}^*$ that are right before the hand-off instance are used as the initial values of $v_{ds,IFOC}^*$ and $v_{qs,IFOC}^*$ after switching controllers. This is shown in Fig. 99 (the purple part). When switching to the DTC controller, since the outputs of voltage PI controllers in the DTC controller are $v_{ds}^*$ and $v_{qs}^*$, an extra synchronous-to-stationary transform is needed in the V/f and IFOC controllers, which transforms $v_{ds}^*$ and $v_{qs}^*$ to $v_{ds}^*$ and $v_{qs}^*$. An example of switching from the V/f controller to the DTC controller is shown in Fig. 100. Note that we cannot perform this reset action if the controllers are switched in $abc$-frame, since we assume no communication to the internal of each controller in that case. Therefore, due to PI-related issues, the nonexecuted closed-loop controller(s) has(have) to be deactivated or idle in $abc$-frame switching. This will leads to significant transients when activating them.
5.2.3 Simulation Verification of the Proposed Drive and Smooth Transition Between Different Controllers

The proposed drive and smooth hand-off transition are first tested in simulation using Simulink. The DTC, IFOC and V/f controllers are built in Simulink to control a 1.5HP 4-pole general-purpose IM. Then, the controllers are integrated to form abc- and synchronous-frame switching as shown in Fig. 93 and Fig. 94, respectively. All the six switching types, including both sensor failure and recovery conditions, between any two of the three controllers are tested. Manual switching of controllers is used in the simulation and later experimental tests without integrating sensor condition monitor, but it will not affect the validation of the proposed drive and control as discussed before.
The IM is operated at 1800RPM and about 50% $T_L$ (3.08N·m), while controllers are switched at steady state. Since the results of the six switching types are similar and support the same conclusions, only the results of switching from 1) The V/f controller to the DTC controller (an open-loop to a closed-loop controller); 2) The IFOC controller to the V/f controller (a closed-loop to an open-loop controller); 3) THE DTC controller to the IFOC controller (a closed-loop to a closed-loop controller) are provided, which are shown in Fig. 101–Fig. 106. For each switching type, the scaled voltage command ($V_{abc}^*$), $i_{abc}$, $T_e$ and $Spd$ are plotted and compared for the $abc$- and synchronous-frame switching. The controllers are arbitrarily switched at 4s, 20s, and 10s for the previous three switching types after the machine reaches steady state.
Fig. 101. Transients of abc-frame switching from V/f to DTC controller: (a) voltage commands; (b) current feedback; (c) torque response; (d) speed response

Fig. 102. Transients of abc-frame switching from V/f to DTC controller: (a) voltage commands; (b) current feedback; (c) torque response; (d) speed response

Fig. 103. Transients of abc-frame switching from IFOC to V/f controller: (a) voltage commands; (b) current feedback; (c) torque response; (d) speed response

Fig. 104. Transients of synchronous-frame switching from IFOC to V/f controller: (a) voltage commands; (b) current feedback; (c) torque response; (d) speed response
Fig. 105. Transients of $abc$-frame switching from DTC to IFOC controller: (a) voltage commands; (b) current feedback; (c) torque response; (d) speed response.

Fig. 106. Transients of synchronous-frame switching from DTC to IFOC controller: (a) voltage commands; (b) current feedback; (c) torque response; (d) speed responses.

It is seen from the simulation results that an instantaneous change happens in $V_{abcs}^*$ when switching in the $abc$-frame, whereas the voltage change is not observed when using the synchronous-frame switching. Therefore, significant stator current and torque responses are generated at the hand-off transient in the $abc$-frame switching, as expected. These large transients not only create large instantaneous electrical and mechanical stress on the machine, which could shorten its lifetime or damage it on site, but also may trigger the machine’s or drive’s protection circuitry (e.g. over-current protection), which could shut down the machine instantaneously. The proposed drive, however, has minimum current and torque transients. Great improvement is also seen from the speed response when switching from the DTC controller to the IFOC controller. Note that the synchronous-frame switching may take a little longer to steady state than the $abc$-frame switching due to the rate limiter. For example, this can be seen by comparing the torque responses in Fig. 103 and Fig. 104. There is, similar to other control
systems, an intrinsic trade-off between settling time and magnitude of transients. However, this short delay is much safer for the machine and inverter than the large transient responses. The delay can also be controlled by adjusting the slew rate of the rate limiter.

5.2.4 Experimental Verification of the Proposed Drive and Smooth Transition Between Different Controllers

Experimental tests that compare the abc- and synchronous-frame switching of the controllers are performed to further validate the proposed drive and its smooth hand-off transition. Switching between the IFOC controller and the V/f controller is applied, as a proof of concept. The experimental verification is carried out at different speed and load torque conditions, which are summarized in TABLE XIX.

<table>
<thead>
<tr>
<th>Switching Type</th>
<th>Speed (RPM)</th>
<th>Load Torque (N·m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 IFOC to V/f</td>
<td>1800</td>
<td>3.08</td>
</tr>
<tr>
<td>2 V/f to IFOC</td>
<td>1800</td>
<td>1.05</td>
</tr>
<tr>
<td>3 IFOC to V/f</td>
<td>600</td>
<td>0.53</td>
</tr>
<tr>
<td>4 V/f to IFOC</td>
<td>600</td>
<td>0.53</td>
</tr>
</tbody>
</table>

The experimental setup is shown in Fig. 21. The IFOC and V/f controllers are built in Simulink and then loaded onto dSPACE 1104 platform. A virtual panel shown in Fig. 107 is created in dSPACE ControlDesk, which provides the speed command, switches between the controllers, and displays interested machine information in real time. A 1.5HP IM is loaded by a Kollmorgen servomotor which is used as a dynamometer, and can provide speed feedback. The current feedback are obtained from a three-phase inverter. The speed and current feedback are scoped to monitor the hand-off transients.
Fig. 107. The virtual panel for experimental validation of the proposed fault-tolerant control

The experimental results of the four testing conditions are shown in Fig. 108, Fig. 110, Fig. 112 and Fig. 113, respectively, while the zoom-in versions of Fig. 108 and Fig. 110 are also provided in Fig. 109 and Fig. 111 emphasizing the details at the hand-off transient. In each figure, channels 1 and 2 are the speed feedback and one phase of the current feedback, which are scaled by 1/270 and 1/6 of their real values, respectively. Channel 3 shows the signal that decides the executing controller: 0 represents the IFOC controller and 1 represents the V/f controller. Change of the level in channel 3 means a change of controller. It is clear in Fig. 108 to Fig. 113 that the proposed drive can achieve much smoother hand-off transients than the normal abc-frame switching in all of the conditions. Moreover, there is no sudden change of current magnitude at the switching instance using the synchronous-frame switching. On the other hand, the switching between the IFOC and V/f controllers at 1800RPM and rated $T_L$ using the
synchronous-frame switching is shown in Fig. 114. Note that this operating condition is not available for the \(abc\)-frame switching, since the large current transient will trigger the overcurrent protection of the inverter and shutdown the setup. This figure indicates that the proposed drive can provide smooth hand-off transition all the way up to rated operating condition and can avoid immediate shutdown of the machine.

Fig. 108. Switching transients from IFOC to V/f controller at 1800 RPM: (a) switching in \(abc\)-frame; (b) switching in the proposed synchronous frame
Fig. 109. Zoom-in version of switching transients from IFOC to V/f controller at 1800 RPM: (a) switching in $abc$-frame; (b) switching in the proposed synchronous frame

Fig. 110. Switching transients from V/f to IFOC controller at 1800 RPM: (a) switching in $abc$-frame; (b) switching in the proposed synchronous frame
Fig. 111. Zoom-in version of switching transients from V/f to IFOC controller at 1800 RPM: (a) switching in abc-frame; (b) switching in the proposed synchronous frame

Fig. 112. Switching transients from IFOC to V/f controller at 600 RPM: (a) switching in abc-frame; (b) switching in the proposed synchronous frame
Fig. 113. Switching transients from V/f to IFOC controller at 600 RPM: (a) switching in abc-frame; (b) switching in the proposed synchronous frame

Fig. 114. Hand-off transients at 1800 RPM and rated torque using synchronous-frame switching:
(a) IFOC to V/f controller; (b) V/f to IFOC controller
5.3 Summary

This chapter presents a model-based FDD method of IM faults. The presented method can be applied to all four major types of IM faults. The virtual modulating signals of model-based theoretical fault characteristic frequencies are applied to the current feedback, which can generate fault-indicative super low-frequency component if a fault exists. This method is simple and only requires multiplication processing and simple filters. It is also nonintrusive and does not affect main control loop. The accuracy and robustness of the proposed method are excellent.

This chapter also presents a model-based fault-tolerant control of IMs under sensor failures. The control is based on the proposed multi-controller drive which uses DTC, IFOC and V/f controllers as replacements of each other. The DTC and IFOC controllers are properly modified as voltage-type synchronous-frame controllers, which generate DC-type voltage commands in the synchronous frame and thus can be feasibly controlled using a Rate Limiter. The proposed drive can be equally used in sensor failure and sensor recovery conditions. The proposed drive with $qd0$-frame switching is compared with the conventional $abc$-frame switching in simulation and experiment. Significant reduction of hand-off transients when switching between controllers is observed using the proposed drive with $qd0$-frame switching. Since the hand-off transients could break power switches and/or trigger over-current protection of the drive, which will shut down the machine, the proposed drive and control can avoid these schemes and maintain the continuity of operation.
6.1 A Comparison of Rotor Bar Material of Squirrel-Cage IMs for Efficiency Enhancement Purposes

Except applying a better control algorithm, the machine efficiency can also be increased by using new material. A study along this line has been explored, where the efficiency of a squirrel-cage IM is increased by replacing aluminum or copper rotor bars with silver rotor bars. The per-phase equivalent circuit model shown in Fig. 9 is used in this study to estimate losses and efficiency of machines using different rotor bars, while the excitation and load are identical for different machines.

Aluminum has been used as the traditional rotor bar material in IMs due to its high conductivity, low price and weight, as well as its decent physical properties to meet thermal, metallurgical and mechanical requirements. To pursue higher machine efficiency, a more expensive material, copper, has been proposed for rotor bars over the past two decades due to having over 60% higher conductivity compared to aluminum, such as in MVIMs. It is believed that the long-term operation benefits of using copper rotor bars can pay back the extra purchasing cost of copper rotor machines [120]–[122]. On the other hand, the high melting point of copper leads to serious thermal stress on casting dies which greatly shortens their lifetime and causes manufacturing difficulty, especially for medium- and small-sized IMs [123]. Some metallurgy technologies targeting on this problem have been developed during these years [124],
Following up with this line, silver is explored as a hypothetical rotor bar material, since it has even higher conductivity than copper.

In the per-phase equivalent circuit model, the torque, output power, copper loss, core loss and the efficiency are calculated through (6.1) to (6.5)

\[
T_e = \frac{3V_{th}^2 R'_{r} / s}{\alpha_e \left[ \left( R_{th} + R'_{r} / s \right)^2 + \left( X_{th} + X'_{r} \right)^2 \right]}, \quad (6.1)
\]

\[
P_{out} = 3I_r^2 \left( \frac{1-s}{s} \right) R'_{r}, \quad (6.2)
\]

\[
P_{Cu} = 3 \left| \left( R'_{r} / s + jX'_{r} \right) I_r \right|^2 \left( \frac{j r_{e-ph} X_m}{r_{e-ph} + j X_m} \right) + I_r \left( R_{r} + I_r^2 R'_{r} \right), \quad (6.3)
\]

\[
P_{core} = 3 \left( R'_{r} / s + jX'_{r} \right) I_r \left( \frac{j r_{e-ph} X_m}{r_{e-ph} + j X_m} \right)^2, \quad (6.4)
\]

\[
\eta = \frac{P_{out}}{P_{out} + P_{Cu} + P_{core}} \times 100\%, \quad (6.5)
\]

where \( V_{th}, X_{th} \) and \( R_{th} \) are the Thevenin voltage, reactance and resistance looking from the magnetizing branch to the input side. \( \eta \) is the machine efficiency and \( I_r' \) is the RMS value of the rotor current. \( X_m \) and \( X'_{r} \) are the magnetizing reactance and rotor reactance in the per-phase equivalent circuit.

Three real IMs (1.5 HP, 3 HP and 10 HP) of aluminum rotor bars are used for the case study. Their parameters, which are needed for (6.1) to (6.5), are obtained through machine characterization tests. As shown in (6.2) to (6.5), \( P_{out}, P_{Cu}, P_{core} \) and \( \eta \) are functions of \( I_r' \), while \( I_r' \) can be calculated from (6.2) considering that the \( P_{out-slip} \) relationship is available from the
The $T_c$–slip curve of the machine based on (6.1). Therefore, the machine power losses and $\eta$ versus $P_{out}$ can be calculated. The results for the 1.5 HP machine are shown in Fig. 115–Fig. 117 as examples. Moreover, the $T_c$–s curve and the $P_{out}$–s curve are plotted to analyze machine performances, which are shown in Fig. 118 and Fig. 119.

![Copper-loss curve of the 1.5HP machine](image1)

**Fig. 115. Copper-loss curve of the 1.5HP machine**

![Core-loss curve of the 1.5HP machine](image2)

**Fig. 116. Core-loss curve of the 1.5HP machine**

To explore the impact of rotor bar material on $\eta$ and machine performance, $R_r'$ is scaled with respect to the conductivity relationship of silver versus aluminum and copper versus aluminum. Repeating the MATLAB code with the new rotor resistances, $R_r'_{Ag}$ and $R_r'_{Cu}$, the machine losses and $\eta$ as well as the torque and output power performances for silver and copper rotor bars are plotted in Fig. 115–Fig. 119 and compared to the aluminum case. The resultant
curves for the 3 HP and 10 HP machines have similar shapes as that for the 1.5 HP one, but with different numerical values.

As before, $P_{Cu}$ is found to increase significantly with load. Silver has the smallest $P_{Cu}$ in all conditions due to its highest conductivity. The $P_{Cu}$ difference of the three materials are more
evident with the increase of $T_L$. Thus, high conductivity materials have better application in heavily loaded machines from energy and efficiency perspectives. On the other hand, $P_{core}$ is relatively less affected by load. Rotor bar material barely has impact on $P_{core}$. The slight increase of $P_{core}$ due to using high-conductivity rotor bar material is caused by the small increase of magnetizing branch voltage while $R_c$ remains the same. As shown in Fig. 117, silver rotor machine has the highest $\eta$ in all conditions followed by the copper type. Aluminum is the least efficient among the three. The same conclusions apply for the 3 HP and 10 HP machines. To obtain quantitative analysis of $\eta$ enhancement, $\eta$ of the three IMs under four $T_L$ are summarized in TABLE XX for comparison. It is found that the degree of $\eta$ enhancement increases with the load of the machine, or the power ratings of machines while they are operating at the same percentage load. On the other hand, it is found that the $\eta$ enhancement of silver versus copper is smaller than 0.12 percentage points, which is as expected since they only have roughly 6% conductivity difference.

As for the machine performance aspects, the $T_e$–speed curve and $P_{out}$ capability are the two characteristics of main interest. Fig. 118 shows that increasing the conductivity of rotor bar material will push the pull-out torque point to the synchronous speed and make the linear region steeper, while the value of pull-out torque keeps the same. Therefore, increasing rotor bar conductivity can help machine maintain rated torque without losing much speed to almost mimic synchronous motor operation in the IM’s linear region. Moreover high rotor bar conductivity is found to decrease the starting torque of the machine. Thus, it can relieve issues related to large in-rush current and give a smoother startup transient. On the other hand, Fig. 119 shows that increasing rotor bar conductivity can improve the maximum $P_{out}$ which appears at lower slip now. Moreover, $P_{out}$ of high conductivity rotor bar machines is larger than $P_{out}$ of the relatively
low conductivity rotor bar machines at low slip conditions, but is smaller at high slip conditions. The crossover point depends on the machines’ relative conductivities.

It is seen from TABLE XX that the efficiency enhancement from the copper rotor machine to the silver rotor machine is tiny, while silver is significantly more expensive than copper or aluminum. Therefore, the long-term benefit of the silver rotor machine due to accumulative operation savings seems not be able to pay back the extra initial purchasing cost for general-purpose IMs during their lifetime. However, this study is still inspiring to have silver or silver alloys in consideration and the payback time may be significantly reduced due to the optimization of rotor bar shapes and sizes as well as the changes of material and electricity prices. Moreover, the silver rotor IMs could be useful for energy-limited applications where refueling is difficult and costive, and thus even tiny efficiency enhancement is desired, such as in aerospace or other tough environments.

**TABLE XX. Efficiency of Machines Using Different Rotor Bar Materials**

<table>
<thead>
<tr>
<th>Machine</th>
<th>Load Percentage</th>
<th>Al</th>
<th>Cu</th>
<th>Ag</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.5 HP</td>
<td>100%</td>
<td>83.01</td>
<td>84.10</td>
<td>84.18</td>
</tr>
<tr>
<td></td>
<td>75%</td>
<td>83.79</td>
<td>84.54</td>
<td>84.59</td>
</tr>
<tr>
<td></td>
<td>50%</td>
<td>83.03</td>
<td>83.48</td>
<td>83.51</td>
</tr>
<tr>
<td></td>
<td>25%</td>
<td>77.51</td>
<td>77.69</td>
<td>77.71</td>
</tr>
<tr>
<td></td>
<td>100%</td>
<td>86.94</td>
<td>88.14</td>
<td>88.23</td>
</tr>
<tr>
<td></td>
<td>75%</td>
<td>87.87</td>
<td>88.71</td>
<td>88.78</td>
</tr>
<tr>
<td></td>
<td>50%</td>
<td>87.67</td>
<td>88.19</td>
<td>88.23</td>
</tr>
<tr>
<td></td>
<td>25%</td>
<td>83.88</td>
<td>84.10</td>
<td>84.12</td>
</tr>
<tr>
<td>3HP</td>
<td>100%</td>
<td>85.86</td>
<td>87.30</td>
<td>87.41</td>
</tr>
<tr>
<td></td>
<td>75%</td>
<td>87.54</td>
<td>88.53</td>
<td>88.60</td>
</tr>
<tr>
<td></td>
<td>50%</td>
<td>87.87</td>
<td>88.47</td>
<td>88.51</td>
</tr>
<tr>
<td></td>
<td>25%</td>
<td>84.40</td>
<td>84.65</td>
<td>84.67</td>
</tr>
</tbody>
</table>
6.2 Torque Enhancement and Re-rating of Medium-Voltage IMs Using Nano-structured Stator Winding Insulation

Except analytical models, finite-element-analysis (FEA) models are also explored to improve the performance of IMs. Specifically, increasing the torque, current and power ratings of medium voltage IMs using new nano-structured insulation material for stator winding. Increasing power density is critical for large propulsion motors as well as for aviation and aerospace applications. It is reported in Office of Naval Research (ONR) Next Gen Integrated Power System Roadmap that improving dielectric insulation could enable payload efficiency and affordable high power density of integrated propulsion motors [126], [127]. This study presents a multi-physics FEA model that is used to study the impacts of a new nano-structured insulation material on large propulsion motors, such as medium-voltage IMs (MVIMs). This work is collaborated with Prof. Yang Cao’s group at UConn. My focus is on creating the multi-physics FEA simulation, which can provide the infrastructure to evaluate different insulation materials, even hypothetical ones, at various operation conditions. The focus of Prof Yang Cao’s group is on developing and studying new nano-structured material.

The multi-physics FEA simulation can give numerous electromagnetic and thermal inspections before a complex and costly real motor is built and tested. Such simulation and the associated analyses are not trivial since: 1) Parameters and public resources to build a FEA model of an MVIM are scarce; 2) Winding insulation is typically ignored in most FEA simulations of electric machines; 3) Integrated electromagnetic and thermal analyses are not often co-simulated and simultaneously analyzed. Due to the expected better characteristics of the new insulation material, such as aging, breakdown voltage, corona resistance and especially the greatly increased thermal conductivity, the machine with the new insulation material could
tolerate more stator current than a similar machine with conventional micaceous insulation without exceeding the temperature limit. This will lead to higher machine current, torque, and power ratings as well as higher torque density and machine efficiency of the machine using the new insulation material.

6.1.1 Multi-Physics Finite-Element-Analysis Simulation

6.1.1.1 Electromagnetic Model

The electromagnetic simulation is performed using ANSYS Maxwell. A self-designed MVIM is used as the subject of this work. The machine design starts by using the RMxpert tool in ANSYS Maxwell, which is a template-based tool for fast design of electric machines. Machine structural parameters, electrical setups and operating condition, etc. are input to the RMxpert model that can automatically generate corresponding machine geometry and excitation. Moreover, it can roughly estimate electrical properties, such as rated current, losses, efficiency, etc., and mechanical properties, such as speed and torque, etc., of the machine based on the simplified equivalent circuit of the machine. Part of the RMxpert model parameters in this work are referred to [128], whereas other parameters are tested and decided so that the results of the designed RMxpert MVIM model are similar to those in [128]. The parameters from [128] are shown in TABLE XXI, while the parameters decided from our tests are summarized in TABLE XXII. The parameters of stator and rotor slots in TABLE XXI and TABLE XXII are referred to Fig. 120. The RMxpert model is shown in Fig. 121.
Fig. 120. Parameters for stator and rotor slots: (a) stator slot; (b) rotor slot

**TABLE XXI. THE PARAMETERS REFERRED TO [128]**

<table>
<thead>
<tr>
<th>Machine</th>
<th>Number of Poles</th>
<th>24</th>
<th>Reference Speed</th>
<th>193 RPM</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Machine → Stator</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Outer Diameter</td>
<td>1170 mm</td>
<td>Inner Diameter</td>
<td>950 mm</td>
<td></td>
</tr>
<tr>
<td>Length</td>
<td>1000 mm</td>
<td>Number of Slots</td>
<td>144</td>
<td></td>
</tr>
<tr>
<td><strong>Machine → Stator → Slot</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Hs0</td>
<td>1 mm</td>
<td>Hs1</td>
<td>2.5 mm</td>
<td></td>
</tr>
<tr>
<td>Hs2</td>
<td>75 mm</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Machine → Stator → Winding</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Coil Pitch</td>
<td>5</td>
<td>Number of Strands</td>
<td>1</td>
<td></td>
</tr>
<tr>
<td><strong>Machine → Rotor</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of Slots</td>
<td>108</td>
<td>Outer Diameter</td>
<td>948 mm</td>
<td></td>
</tr>
<tr>
<td>Inner Diameter</td>
<td>180 mm</td>
<td>Length</td>
<td>1000 mm</td>
<td></td>
</tr>
<tr>
<td><strong>Machine → Rotor → Winding</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Bar Conductor</td>
<td>Copper</td>
<td>End Length</td>
<td>25 mm</td>
<td></td>
</tr>
<tr>
<td>End Ring Width</td>
<td>25 mm</td>
<td>End Ring Height</td>
<td>25 mm</td>
<td></td>
</tr>
<tr>
<td><strong>Machine → Rotor → Slot</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>None</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Analysis → Setup</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Rated Output Power</td>
<td>4500 HP</td>
<td>Rated Voltage</td>
<td>4160 V</td>
<td></td>
</tr>
<tr>
<td>Rated Speed</td>
<td>193 RPM</td>
<td>Operating Temperature</td>
<td>75 °C</td>
<td></td>
</tr>
<tr>
<td>Winding Connection</td>
<td>Wye</td>
<td>Frequency</td>
<td>40 Hz</td>
<td></td>
</tr>
</tbody>
</table>
Fig. 121. The RMxpert model of the MVIM

### TABLE XXII. THE PARAMETERS DECIDED FROM OUR TESTS

<table>
<thead>
<tr>
<th>Machine</th>
<th>Stray Loss Factor</th>
<th>0.009</th>
<th>Frictional Loss</th>
<th>283 W</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Windage Loss</td>
<td>0 W</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Machine → Stator</th>
</tr>
</thead>
<tbody>
<tr>
<td>Steel Type</td>
</tr>
<tr>
<td>Slot Type</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Machine → Stator → Slot</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bs1</td>
</tr>
<tr>
<td>Bs2</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Machine → Stator → Winding</th>
</tr>
</thead>
<tbody>
<tr>
<td>Winding Type</td>
</tr>
<tr>
<td>Winding Layers</td>
</tr>
<tr>
<td>Parallel Branches</td>
</tr>
<tr>
<td>Conductors per Slot</td>
</tr>
<tr>
<td>Wire Size</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Machine → Rotor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slot Type</td>
</tr>
<tr>
<td>Steel Type</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Machine → Rotor → Winding</th>
</tr>
</thead>
<tbody>
<tr>
<td>None</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Machine → Rotor → Slot</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hs0</td>
</tr>
<tr>
<td>Hs1</td>
</tr>
<tr>
<td>Hs01</td>
</tr>
<tr>
<td>Hs2</td>
</tr>
<tr>
<td>Bs0</td>
</tr>
<tr>
<td>Bs1</td>
</tr>
<tr>
<td>Bs2</td>
</tr>
<tr>
<td>Rs</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Analysis → Setup</th>
</tr>
</thead>
<tbody>
<tr>
<td>None</td>
</tr>
</tbody>
</table>

143
Then, a 2D FEA model of the MVIM is created based on the RMxprt model using the built-in function in ANSYS Maxwell. The features of the 2D FEA model, such as mesh, excitation, boundary conditions, operation condition, etc., are automatically set up during the creation. Therefore, tediously drawing and setting up of the FEA model can be avoided. On the other hand, the magnetic symmetry based on the number of magnetic poles in the MVIM is recognized during the creation. Thus, only part of the cross-section of the MVIM is created in the 2D FEA model with proper electromagnetic boundary conditions, which is enough to analyze the complete electromagnetic distribution inside the machine. The created 2D FEA model is shown in Fig. 122. The zoomed-in view of the stator slot is also shown in Fig. 122. Apparently, the automatically generated FEA 2D model does not include insulation layer. Therefore, an insulation layer is drawn and added to the model in Fig. 122 with properly shrunk coil cross-section. The updated 2D FEA model with insulation layer is shown in Fig. 123, where spacers and wedges are also added to form a more realistic stator winding system.

Different materials are assigned to different parts of the 2D FEA model, which are summarized in TABLE XXIII. Especially, the proposed nano-structured material and the conventional micaceous material are assigned to the insulation layer for the comparative study. The major characteristics of the two insulation materials are compared in TABLE XXIV. Note that the focus of this work is on the relative comparison of the two insulation materials on the machine’s performances, where the only difference between the two simulation models is the material property of the insulation layer.
Fig. 122. The 2D FEA model of the MVIM without insulation layer

Fig. 123. The updated 2D FEA model of the MVIM with insulation layer, spacer and wedge

**TABLE XXIII. THE APPLIED MATERIAL IN THE 2D FEA MODEL**

<table>
<thead>
<tr>
<th>Machine Part</th>
<th>Stator Core</th>
<th>Stator Winding</th>
<th>Rotor Core</th>
<th>Rotor Bar</th>
</tr>
</thead>
<tbody>
<tr>
<td>Material</td>
<td>M36 Steel</td>
<td>Copper</td>
<td>M36 Steel</td>
<td>Copper</td>
</tr>
<tr>
<td>Machine Part</td>
<td>Insulation</td>
<td>Spacer</td>
<td>Wedge</td>
<td></td>
</tr>
<tr>
<td>Material</td>
<td>Mica or Proposed Insulation</td>
<td>FR4-epoxy</td>
<td>PTFE</td>
<td></td>
</tr>
</tbody>
</table>

6.1.1.2 Thermal Model

The thermal model is constructed using ANSYS Steady-state Thermal analysis system. It inherits the geometry from the 2D FEA model. In the thermal simulation, one of the most
important properties is the material’s thermal conductivity, which are summarized in TABLE XXV for our simulation. Another important setting is the boundary condition for different boundaries of the model. The applied boundaries are shown in TABLE XXVI and Fig. 5.

**TABLE XXIV. The Major Dielectric and Physical Properties of the Proposed and The Conventional Insulation Materials**

<table>
<thead>
<tr>
<th></th>
<th>Mica</th>
<th>The Proposed Material</th>
</tr>
</thead>
<tbody>
<tr>
<td>Relative Permittivity</td>
<td>5.7</td>
<td>5</td>
</tr>
<tr>
<td>Relative Permeability</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Bulk Conductivity</td>
<td>0 Simens/m</td>
<td>0 Simens/m</td>
</tr>
<tr>
<td>Dielectric Loss Tangent</td>
<td>3%</td>
<td>2.5%</td>
</tr>
<tr>
<td>Core Loss Model</td>
<td>None</td>
<td>None</td>
</tr>
<tr>
<td>Mass Density</td>
<td>2500 kg/(m³)</td>
<td>~1500 kg/(m³)</td>
</tr>
<tr>
<td>Composition</td>
<td>Solid</td>
<td>Solid</td>
</tr>
</tbody>
</table>

**TABLE XXV. The Thermal Conductivity (W/(m·°C)) Of Different Machine Parts**

<table>
<thead>
<tr>
<th>Machine Part</th>
<th>Stator and Rotor Cores</th>
<th>Stator Winding and Rotor Bar</th>
<th>The Proposed Insulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thermal Conductivity</td>
<td>40 S/m</td>
<td>400 S/m</td>
<td>0.7 S/m</td>
</tr>
<tr>
<td>Machine Part</td>
<td>Shaft</td>
<td>Mica Insulation</td>
<td>Wedge</td>
</tr>
<tr>
<td>Thermal Conductivity</td>
<td>60.5 S/m</td>
<td>0.25 S/m</td>
<td>1.4 S/m</td>
</tr>
<tr>
<td>Machine Part</td>
<td>Spacer</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Thermal Conductivity</td>
<td>0.294 S/m</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**TABLE XXVI. The Applied Boundary Conditions In The Thermal Simulation**

<table>
<thead>
<tr>
<th>Boundary Type</th>
<th>Major Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Convection (Steel/Air)</td>
<td>Film Coefficient: 1050 W/(m²·°C); Ambient: 22 ºC</td>
</tr>
<tr>
<td>Radiation (Steel/Air)</td>
<td>Emissivity: 0.85; Ambient: 22 ºC</td>
</tr>
<tr>
<td>Convection (Steel/Air)</td>
<td>Film Coefficient: 1050 W/(m²·°C); Ambient: 22 ºC</td>
</tr>
<tr>
<td>Convection (Copper/Air)</td>
<td>Film Coefficient: 1160 W/(m²·°C); Ambient: 22 ºC</td>
</tr>
<tr>
<td>Convection (Wedge/Air)</td>
<td>Film Coefficient: 100 W/(m²·°C); Ambient: 22 ºC</td>
</tr>
<tr>
<td>Fixed Temperature</td>
<td>Temperature: 40 ºC</td>
</tr>
</tbody>
</table>
6.1.1.3 Multi-physics Simulation

The Maxwell 2D model and the Steady-state Thermal model are linked in ANSYS Workbench, which is an interface to link different analysis systems. Different analysis systems can share information by simply connecting proper components of the analysis systems, as shown in Fig. 125. In our case, the Maxwell and Thermal models share the same geometry, while the machine losses calculated in Maxwell are used as internal heat sources in the thermal simulation. The overview of the multi-physics simulation is shown in Fig. 126.
6.1.2 Simulation Results

The common electrical and mechanical properties of the MVIM, such as current, losses, torque, etc., can be plotted with respect to simulation runtime, while the magnetic distribution inside the MVIM can be displayed for a specific simulation instance when the field is saved. The results of electromagnetic analysis for the machines using the proposed and conventional insulation materials are shown in Fig. 127–Fig. 129. It is found that the small changes of dielectric characteristics of the insulation material do not have noticeable impacts on the machine’s electromagnetic and mechanical properties.

![Fig. 127. Comparison of magnetic flux density (B) distribution at simulation runtime = 0.5s using:](image)

(a) mica; (b) proposed insulation
Fig. 128. Comparison of torque at steady state using: (a) mica; (b) proposed insulation

Fig. 129. Comparison of machine total loss at steady state using: (a) mica; (b) proposed insulation

The temperature distribution inside the MVIMs using the conventional micaceous and the proposed insulation materials are shown in Fig. 130, where zoomed-in views of stator slot temperature are also provided. It is observed that the proposed insulation material can decrease the temperature inside the machine.

Fig. 130. Comparison of temperature distribution using: (a) mica; (b) proposed insulation
6.1.3 Machine Re-Rating Process and Results

Due to the decreased temperature distribution inside the MVIM, it is expected to be able to push more current into the machine without exceeding the original machine’s temperature limit. Thus, the machine’s torque and power ratings as well as power density and payload efficiency are expected to increase as well. Iterative testing is performed and the procedure is shown in Fig. 131. Basically, the load torque is increased (decrease speed command) in the Maxwell simulation, which generates more losses (heat sources), until the machine temperature reaches the value of using the conventional insulation at rated load. Then, the over-rated torque and current will be the new ratings of the machine using the proposed insulation material. It is shown in Fig. 132 that the machine’s temperature reaches the original limit shown in Fig. 130(a) when the load torque is increased from 166.4 kN⋅m by 189.8 kN⋅m (14%). The corresponding enhancement of rated current is 26%.

![The Re-rating Process of the MVIM](image)

Fig. 131. The machine re-rating procedure
6.3 Summary

This chapter presents enhancement of IMs’ efficiency and performance (torque rating and power density) through model-based material changes. Silver is hypothetically explored as rotor bar material of IMs. Different \( R_r \) values are used for aluminum, copper and silver in the per-phase equivalent circuit of IMs to study the effect of rotor bar material on efficiency. Less than 0.12\% efficiency enhancement is observed when replacing copper rotor bar with silver rotor bar. Also, silver is much more expensive than copper. Therefore, silver rotor bar is not a good choice for general-purpose IMs presently, but it may be useful for energy-limited applications where refueling is difficult and costive, such as in aerospace.

This chapter also presents a multi-physics FEA model of MVIMs. The model uses co-simulation on electromagnetic and thermal analyses to study the effects of stator winding insulation material on machine performances, where a nano-structured insulation material is compared with the conventional micaceous insulation material. The copper and core losses calculated in the electromagnetic simulation are used as the internal heat sources in the thermal simulation. It is found that insulation material barely has effects on the electromagnetic...
distribution inside the machine. However, the nano-structured insulation material can significantly decrease the machine’s temperature due to its higher thermal conductivity than mica. Thus, using the nano-structured insulation material could push more current into the machine without exceeding the temperature limit when using mica. As a result, it is found that the torque rating and power density of the MVIM could be increased by 14% when using the nano-structured insulation material, and the corresponding efficiency enhancement is 26%.
CHAPTER 7
CONCLUSIONS AND CONTRIBUTIONS

This dissertation presented advanced models and model-based control and diagnostics to improve induction motor drives’ efficiency, availability and performance. Different power loss models of three-phase induction motors and drive systems are reviewed and compared. Various loss minimization control methods of induction motors and drive systems are classified using different criteria, and compared mainly based on the convergence rate, steady-state error and parameter dependence. Four major types of induction machine faults and their causes are reviewed. Moreover, a state-of-the-art review on the fault detection and diagnosis methods of induction machines is given, where recent research trends and developments as well as gaps and new ideas are discussed. Different fault-tolerant control of induction motor drive systems under sensor failures are also reviewed. Based on the gaps in the literature and needs in real applications, new models, controls, fault detection and diagnosis, and material configurations are provided in this dissertation.

An advanced dynamic core-loss model, which can perform $qd0$-frame analysis and vector control in any reference frame, is proposed, elaborately derived, tested, and validated in simulation and experiments under various operating conditions. The mechanical loss is modelled as a linear function of speed. The stray loss is modelled as a linear function of torque squared, where the linearity values are saved in a look-up table for different speeds. The accuracy of the loss estimation is shown to change with a machine’s power rating and operating condition. For example, in Fig. 29, the estimation errors of copper and core losses are 6.0W(4.9%) and 1.7W(4.8%) at the rated load torque, and 1.5W(5.7%) and 0.6W(1.75%) at the 25% of the rated load torque. The minimum accuracy for all the tested operating conditions and the three
induction machines is greater than 93%. This core-loss model can be used as a basis for various control designs and machine analysis approaches, which could provide improved results by considering copper, core, mechanical, and stray losses.

A system-level power loss model for inverter-fed induction motor drives, which can be extended to full inverter-fed induction motor drives, is proposed and integrated with V/f control and IFOC for loss minimization control (LMC) of the overall drive system. Specifically, the LMC decides the optimal flux through controlling $Vf^*$ and $\lambda_{dr^*e^*}$ in V/f control and IFOC, respectively. Based on a systematic series of experimental tests on mechanical and stray losses, these two losses are treated as constant during LMC due to their low sensitivity with respect to flux for the load range that is interesting to LMC. The room for useful LMC operation varies for machine ratings and operating conditions. Generally, higher machine ratings and higher loads will have less room for LMC operation. For the tested 1.5 HP inverter-fed inverter motor drive under V/f control, the power loss of the system is decreased by more than 13% at 1800 RPM with 1.2 N·m, and 1200 RPM with 0.6 N·m conditions using the proposed LMC.

An improved MTPA controller, which considers core loss in the control design, is proposed and discussed. A new set of reference commands for current regulation in the synchronous frame is given, which can lead to higher TPA ratio than the conventional MTPA. The increase of the TPA ratio depends on the values of $R_c$ and $\omega_e$, which could be large for high-speed induction machines or induction machines with relatively large core loss (small $R_c$ value).

An adaptive time-domain FDD method, which requires no extra hardware other than that needed for closed-loop regulatory control, is invented and comprehensively tested. The FDD method shows excellent fault sensitivity and robustness to noise at different speeds, torques, fault magnitudes, magnitudes of the modulating signal, and drift frequencies. The response time of the
method is fast. It is expected that, in the future, the proposed method can be extended to other types of faults that have periodic mathematical expressions of fault frequencies. Moreover, adaptive modulation can be applied to other feedback signals, such as vibration signal, in a similar approach to the one used in current feedback.

A synchronous-frame multi-controller drive for fault-tolerant control of induction machines with sensor failures is proposed in this dissertation. The proposed drive uses concurrently active DTC, IFOC and V/f controllers to back up each other in case the executing controller fails due to sensor failure. The proposed drive can also be used equally in sensor recovery conditions to retrieve better drive performance. The conventional DTC and IFOC controllers are modified and then integrated with the conventional V/f controller in the synchronous frame. A simple rate limiter is applied to smooth transient responses during switching between different controllers, which takes advantages of DC-type synchronous-frame voltage commands from different controllers. Simulation and experimental verification of the proposed drive are performed, which show significant improvement of hand-off transition compared to the normal $\text{abc}$-frame switching. The presented synchronous-frame switching idea can also be extended to other types of machines that can be controlled by different synchronous-frame controllers.

Different material to enhance induction machine performance are also studied through finite element models. First, different rotor bar material are evaluated for better efficiency of the machine by design choice. Results showed that copper provides the highest efficiency per cost. Also, a multi-physics finite-element-analysis model of medium-voltage induction machines is designed and introduced to study different stator winding insulation materials on machine performance. The construction details of the model are provided. Moreover, the comparison of
the machines using the proposed and the conventional insulation materials is given in terms of their electrical, magnetic, mechanical and thermal properties. It is found that the minor difference between the proposed and the conventional insulation materials do not significantly change the machine’s electromagnetic and mechanical properties. However, the increased thermal conductivity of the proposed insulation material greatly decreases the machine’s maximum temperature given the same heat sources and thermal boundaries. Therefore, the torque and current ratings of the MVIM can be increased by 14% and 26% purely by replacing the insulation material.
LIST OF PUBLICATIONS

Journal and Magazine Articles:


Conference Papers:


APPENDIX A

THE NUMERICAL SWEEP USED FOR LMC OF V/f-CONTROLLED IMs

%%%Input the speed command and load torque
disp('****please input the speed command and load torque****')
speed=input('speed(RPM)=');
TL=input('TL(N.m)=');

%% DC test
Rs=((12.18+12.59+12.47)/3/4.299)/2;

%% No load test
noload_V=184.29;
noload_I=1.492733;
noload_P=49.51333;
lockrotor_V=43.20097;
lockrotor_I=3.984;
lockrotor_P=132.4322;
Pmech=87.866;
kStray1=1.1845;
kStray2=9.1957;

%% No load test
vll_nl=noload_V; %measured average line-to-line voltage
vph_nl=vll_nl/sqrt(3); %calculated average phase voltage
ill_nl=noload_I; %measured average line-to-line current
iph_nl=ill_nl; %calculated average phase current
pph_nl=noload_P/3; %measured per-phase input power
rc_ph(1)=vph_nl^2/pph_nl; %first calculation of rc_ph
xm(1)=vph_nl^2/sqrt((vph_nl*iph_nl)^2-pph_nl^2);

%% Lock rotor test
vll_lr=lockrotor_V; %measured average line-to-line voltage
vph_lr=vll_lr/sqrt(3); %calculated average phase voltage
ill_lr=lockrotor_I; %measured average line-to-line current
iph_lr=ill_lr; %calculated average phase current
pph_lr=lockrotor_P/3; %measured per-phase input power
Req(1)=pph_lr/(iph_lr^2); %first calculation of Rs+Rr'
rr_prime(1)=Req(1)-Rs; %first calculation of Rr'
xeq(1)=sqrt((vph_lr*iph_lr)^2-pph_lr^2)/(iph_lr^2); %first calculation of xls+xlr'
xls(1)=xeq(1)*0.4; %first calculation of xls based on NEMA standard of the machine
xlr_prime(1)=xeq(1)*0.6; %first calculation of xlr' based on NEMA standard of the machine

kk=10; %number of iteration; To make the calculated parameters more accurate
for i=2:kk
    % NL test
    pm(i)=pph_nl-iph_nl^2*Rs;
    qm(i)=sqrt((vph_nl*iph_nl)^2-pph_nl^2)-iph_nl^2*xls(i-1);
    theta=acos(pph_nl/(vph_nl*iph_nl));
    u(i)=vph_nl-(iph_nl*cos(-theta)+j*iph_nl*sin(-theta))*(Rs+j*xls(i-1));
    rc_ph(i)=abs(u(i))^2/pm(i);
xm(i)=abs(u(i))^2/qm(i);

%LR test
alpha=acos(pph_lr/(vph_lr*iph_lr));
is_lr=iph_lr*cos(-alpha)+j*i ph_lr*sin(-alpha);
vm_lr(i)=vph_lr-is_lr*(Rs+j*xls(i-1));
im_lr(i)=vm_lr(i)/(rc_ph(i)+j*xm(i));
pmm(i)=abs(vm_lr(i))^2/rc_ph(i);
qmm(i)=abs(vm_lr(i))^2/xm(i);
ir_lr(i)=is_lr-im_lr(i);
rr_prime(i)=(pph_lr-iph_lr^2*Rs-pmm(i))/(abs(ir_lr(i))^2);
xlr_prime(i)=(sqrt((vph_lr*iph_lr)^2-pph_lr^2)-iph_lr^2*xls(i-1)-
qmm(i))/(abs(ir_lr(i))^2);
xls(i)=xlr_prime(i)*0.4/0.6;
end

% Machine parameters
P=4; % number of poles
J=0.089/2; % inertia
Lls=xls(kk)/(2*pi*speed/30);
Lm=xm(kk)/(2*pi*speed/30);
Lms=Lm*2/3;
Rr_prime=rr_prime(kk);
Llr_prime=xlr_prime(kk)/(2*pi*speed/30);
Lss=Lls+Lm;
Lrr_prime=Llr_prime+Lm;
LM=[Lss 0 0 0 Lm 0 0 Ls 0 Lms 0 0 0 Lm 0 0 Lrr_prime 0 0 0 Lm 0 0 Lrr_prime 0 0 0 Lrr_prime];
inverse_LM=inv(LM);
Rc=rc_ph(kk)*2/3;

we=speed/60*2*pi*P/2;%synchronous speed
Fstray=kStray1*TL^2+kStray2;
MI=0.9;
DC=417.3;

%Perform symbolic calculation
syms iqr idr iqshat idshat wr
Eff=zeros(1,30);
Pcu=zeros(1,30);
Pcore=zeros(1,30);
Pcc=zeros(1,30);
vff=zeros(1,30);
RPM=zeros(1,30);

for iii=1:30;
  vf=2.213-(iii-1)*0.05%sweep vf from rated value
  vff(1,iii)=vf;
  vqs = vf*speed*DC/2/(2.213)/1800*MI;

  % Solve the simplified linear machine constitution equations
  eqns=[vqs==vph_lr-is_lr*Rs-vls*Lms*xm(i)+vff(i,iii)-vph_lr*iph_lr+iquestat+idplat+is plat+ilm]
\[(R_s - w_e^2 \cdot L_{ls} \cdot L_{ms}/R_c) \cdot i_{dshat} - (R_s \cdot L_{ms} \cdot w_e/R_c + w_e \cdot L_{ls} + w_e \cdot L_m) \cdot i_{qshat} - w_e \cdot L_m \cdot i_{qr},\]

\[R_{r_prime} \cdot i_{qr} - (w_e - w_r) \cdot (L_{lr_prime} + L_m) \cdot i_{drr} + (w_e - w_r) \cdot L_m \cdot i_{dshat},\]

\[R_{r_prime} \cdot i_{drr} - (w_e - w_r) \cdot (L_{lr_prime} + L_m) \cdot i_{iqr} - (w_e - w_r) \cdot L_m \cdot i_{iqshat},\]

\[T_L + (P_{mech} + P_{stray})/(w_e/(P/2)) = 3 \cdot P \cdot L_m/4 \cdot (i_{qshat} \cdot i_{dr} - i_{dshat} \cdot i_{qr})];\]

\[\text{machine=solve(eqns,iqr,idr,iqshat,idshat,wr);}\]

% The solution has two sets of answers, using \(k\) to take the correct one (for some MATLAB version, \(k\) should be 1, but for others, \(k\) should be 2. An easy way to find the \(k\) value is % assigning \(v/f\) a value and then see the results' values. Here, the results % have explicit values instead of being a function of \(v/f\).

\[k=1;\]
\[I_{qr}=vpa(\text{real(machine.iqr}(k)),6);\]
\[I_{dr}=vpa(\text{real(machine.idr}(k)),6);\]
\[I_{qshat}=vpa(\text{real(machine.iqshat}(k)),6);\]
\[I_{dshat}=vpa(\text{real(machine.idshat}(k)),6);\]
\[W_R=vpa(\text{real(machine.wr}(k)),6);\]

\[I_{qs}=I_{qshat} + L_{ms}/R_c \cdot w_e \cdot I_{dshat};\]
\[I_{ds}=I_{dshat} - L_{ms}/R_c \cdot w_e \cdot I_{qshat};\]

% \(RPM(1,iii)=W_R;\)
% \(P_{cu}(1,iii)=\text{real}(1.5 \cdot (R_s \cdot (I_{qs}^2 + I_{ds}^2) + R_{r_prime} \cdot (I_{qr}^2 + I_{dr}^2)));\)
% \(P_{core}(1,iii)=\text{real}(1.5 \cdot w_e^2 \cdot L_m \cdot L_{ms}/R_c \cdot (I_{dshat}^2 + I_{dshat} \cdot I_{dr} + I_{qshat}^2 + I_{qshat} \cdot I_{qr}));\)
% \(P_{cc}(1,iii)=\text{real}(P_{cu}(1,iii) + P_{core}(1,iii));\)
% \(P_{total}=P_{cu}(1,iii) + P_{core}(1,iii) + P_{mech} + P_{stray};\)
% \(Eff(1,iii)=\text{real}((T_L \cdot W_R)/(T_L \cdot W_R/P + P_{total}) \cdot 100);\)

% if \(Eff(1,iii)<Eff(1,iii - 1),\) break, end % if the efficiency starts dropping, breaks the sweep to save time

end

% get the optimal \(v/f\) under this step size
[Eff,IND]=max(Eff);
\[v_{f_best}=2.213 - (\text{IND} - 1) \cdot 0.05\]
APPENDIX B

THE NUMERICAL SWEEP USED FOR LMC OF IFOC-CONTROLLED IM DRIVES

%%%Input the speed command and load torque

\[
\text{disp('****please input the speed command and load torque****')}
\]

\[
\text{TL=input('TL(N.m)=');}
\]

%Machine parameters
\[
P=4; \quad \text{% number of poles}
\]
\[
J=1.662; \quad \text{% inertia}
\]
\[
R_s=0.087;
\]
\[
L_{ls}=0.302/(2*\pi*60);
\]
\[
L_m=13.08/(2*\pi*60);
\]
\[
L_{ms}=L_m*2/3;
\]
\[
R_r'=0.228;
\]
\[
L_{lr}'=0.302/(2*\pi*60);
\]
\[
L_{ss}=L_{ls}+L_m;
\]
\[
L_{rr}'=L_{lr}'+L_m;
\]
\[
\text{LM}=[L_{ss} \ 0 \ 0 \ 0 \ 0; \ 0 \ L_{ss} \ 0 \ 0 \ L_m; \ 0 \ 0 \ L_m \ 0 \ 0; \ 0 \ L_{ms} \ 0 \ 0 \ 0; \ 0 \ L_{ms} \ L_{rr}'; \ 0 \ 0 \ 0 \ 0 \ L_{lr}'];
\]
\[
\text{inverse}_{-}\text{LM} = \text{inv}(\text{LM});
\]
\[
\text{Rc}=200;
\]

\[
\text{we}=60*2*\pi; \quad \text{%synchronous speed}
\]
\[
\text{MI}=0.9;
\]
\[
\text{DC}=834.64;
\]
\[
\text{DC} _{data} = 300;
\]
\[
\text{speed}=1800;
\]
\[
\text{fsw}=10000;
\]
\[
\text{torque}=\text{TL};
\]

\[
\text{Eff} = \text{zeros}(1,80);
\]
\[
\text{Pcu} = \text{zeros}(1,80);
\]
\[
\text{Pcore} = \text{zeros}(1,80);
\]
\[
\text{Pcc} = \text{zeros}(1,80);
\]
\[
\text{Pinv} = \text{zeros}(1,80);
\]
\[
\text{Ptotal} = \text{zeros}(1,80);
\]
\[
\text{Pcd} = \text{zeros}(1,80);
\]
\[
\text{Psw} = \text{zeros}(1,80);
\]
\[
\text{fluxx} = \text{zeros}(1,80);
\]

%%%Find the optimal flux_dr from matlab calculation #Method 2

\[
\text{for } \text{iii}=1:81;
\]
\[
\text{flux_dr} = 0.9528-(\text{iii}-1)*0.01
\]
\[
\text{fluxx}(1,\text{iii})=\text{flux_dr};
\]

\[
\text{Iqshat} = \text{torque} / (3*P/4*L_m*\text{flux_dr}/L_{rr}');
\]
\[
\text{Idshat} = \text{flux_dr}/L_m;
\]
\[
\text{Iqr} = -L_m/L_{rr}'+\text{torque} / (3*P/4*L_m*\text{flux_dr}/L_{rr}');
\]
\[
\text{Idr} = 0;
\]
\[
\text{Iqs} = \text{Iqshat} + \text{Lms}/\text{Rc} * \text{we} * \text{Idshat};
\]
\[
\text{Ids} = \text{Idshat} - \text{Lms}/\text{Rc} * \text{we} * \text{Iqshat};
\]
\[
\text{IL} = \sqrt{\text{Iqs}^2+\text{Ids}^2};
\]
PF=cos(atan(Ids/Iqs));

%calculate inverter conduction loss
beta_ce1=-1e-4;
beta_ce2=0.0233;
beta_ce3=0.7748;
del_ce1=beta_ce1/3/pi+3*beta_ce1*MI*PF/32;
del_ce2=beta_ce2/8+beta_ce2*MI*PF/3/pi;
del_ce3=beta_ce3/2/pi+beta_ce3*MI*PF/8;
Pcd_ce=del_ce1*IL^3+del_ce2*IL^2+del_ce3*IL;

beta_f1=-4e-5;
beta_f2=0.009;
beta_f3=0.7851;
del_f1=beta_f1/3/pi-3*beta_f1*MI*PF/32;
del_f2=beta_f2/8-beta_f2*MI*PF/3/pi;
del_f3=beta_f3/2/pi-beta_f3*MI*PF/8;
Pcd_f=del_f1*IL^3+del_f2*IL^2+del_f3*IL;

Pcd(1,iii)=6*(Pcd_ce+Pcd_f);

%calculate inverter switching loss
Temp=25;
Temp_datsht=125;
TP1=0.51614;
TP2=0;
TP3=0.85722;
Eon_sl=0.0068;
Eoff_sl=0.0263;
Erev_sl=0.0077;
Eon Ini=0.1438*(Temp/Temp_datsht)^TP1;
Eoff Ini=0.4305*(Temp/Temp_datsht)^TP2;
Erev Ini=1.7626*(Temp/Temp_datsht)^TP3;
Psw(1,iii)=1/1000*6*((Eon sl+Eoff sl+Erev sl)*IL*fsw/pi+(Eon Ini+Eoff Ini+Erev Ini)*fsw/2)*DC/DC_datsht;

Pinv(1,iii)=Pcd(1,iii)+Psw(1,iii);

%Machine losses and total efficiency
Pcu(1,iii)=1.5*(Rs*(Iqs^2 + Ids^2) + Rr_prime*(Iqr^2 + Idr^2));
Pcore(1,iii)=1.5*we^2*Lm*Lms/Rc*(Idshat^2 + Idshat*Idr + Iqshat^2 + Iqshat*Iqr);
Pcc(1,iii)=Pcu(1,iii)+Pcore(1,iii);
Ptotal(1,iii)=Pcu(1,iii)+Pcore(1,iii)+Pinv(1,iii);
Eff(1,iii)=(TL*we/2)/((TL*we/2)+Ptotal(1,iii))*100;

end

[EFF,IND]=max(Eff);
flux_best=0.9528-(IND-1)*0.01
REFERENCES


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