

**Accurate Modeling and High Performance Control of Permanent Magnet
Synchronous Machines**

by

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DEDICATION

To,

Lord Sri Krishna
Sri Guru and Sri Gauranga

“Whatever you do, whatever you eat, whatever you offer or give away, and whatever austerities you perform – do that, O son of Kuntī, as an offering to Me.” BG 9.27

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LIST OF ACRONYMS

PMSMs Permanent Magnet Synchronous Machines

PM Permanent Magnet

NL Non-Linear

LUT Look Up Table

FEA Finite Element Analysis

ABSTRACT

AC motors have long been used in manufacturing, oil and gas refineries, and power-generation plants, serving as critical components in sectors that form the backbone of national economies. With increasing interest in renewable energy and vehicle electrification, Permanent Magnet Synchronous Machines (PMSMs) have become a popular choice due to their high efficiency and high power density. Recent advances in cooling techniques now make it possible to design PMSMs with very high operating current densities, under which they exhibit significant nonlinear (NL) behavior due to saturation of their soft magnetic material and temperature variations of permanent magnets (PM) that significantly affects the magnetic properties of the machine, and hence its torque production capability. Accurately capturing these phenomena in machine models is therefore crucial for effective control.

This dissertation advances the state of the art in modeling and control of Permanent Magnet Synchronous Machines. We present a novel magnetics model for PMSMs that accurately captures both the nonlinear and rotor-temperature-dependent behavior of PMSMs. The proposed model decouples the nonlinear effects of magnetic saturation and the rotor temperature effect on permanent magnet excitation. The result is an accurate model suitable for use in real-time control that works over a broad operating range, is computationally efficient, and relatively simple to implement. Further, leveraging the ability to capture the non-linear characteristics of PMSMs accurately, we develop a control scheme for accurate torque regulation and an adaptive parameter estimation strategy for PM excitation. Together, the proposed model, control algorithm, and adaptive estimation strategy form a comprehensive framework for high-performance electric-drive systems.

The first part of this dissertation introduces the new proposed magnetics model for PMSM and demonstrates the accuracy of the proposed model over the standard model from the literature. The motivation of modeling PM excitation as a current source in the proposed model and its benefits is studied by considering a simple single coil magnetic device. Results validating the proposed model both in simulation and experimentally is also presented. The later part of the thesis focuses on demonstration of the use of the proposed PMSM model for real-time control purposes. An adaptive parameter estimation strategy for PM current parameter is developed. Finally, an accurate torque controller is proposed that uses flux

regulator instead of current regulator highlighting the high performance real time controller implementation of the proposed model.

CHAPTER 1

Problem Statement and Research Objectives

This dissertation consists of advancing the frontier of modeling and control of Permanent Magnet Synchronous Machines (PMSMs). With an increased focus on PMSMs for vehicle electrification applications, accurate modeling and the associated control of PMSMs over a wide range of operating conditions is necessary. Specifically, we focus on capturing the non-linear characteristics of PMSMs due to magnetic saturation and the effects of rotor temperature variation. Further, leveraging the ability to capture the non-linear characteristics of PMSMs accurately, we develop a control scheme for accurate torque regulation and an adaptive parameter estimation strategy for the Permanent Magnet (PM) excitation. The proposed model, associated control algorithm, and adaptive estimation strategy altogether form a comprehensive package for high-performance electric drives, and is generic in the sense that it can be used for all types of PMSM designs.

1.1 Motivation and Overview

AC motors have been around for a long time, historically used in the manufacturing sector, oil/gas refineries, and power generation plants. They are a critical component in sectors that form the backbone of the economy of a nation. With increasing interest in fields like renewable energy, electric vehicles, and actuators for robotic applications, PMSMs have become a popular choice due to their high efficiency and high power density [7, 53, 35, 5, 52, 24, 42, 21, 40]. With recent advances in cooling techniques, it is now possible to develop PMSM designs with very high operating current density [50, 25, 14].

At these high current excitation levels, PMSMs tend to exhibit significant nonlinear (NL) behavior due to saturation of their soft magnetic material, as shown in Fig. 1.1-1.2. Thus a linear magnetics model no longer accurately captures the machine behavior in the entire range of operation and necessitates the need of a NL model.

Furthermore, with the possibility of operating PMSMs at very high current excitation, a wide range of temperature variation of the permanent magnets can occur. The flux den-

sity/field intensity relationships of Neodymium-Iron-Boron (NdFeB) permanent magnets (PMs) can change significantly with rotor temperature [3], as shown in Fig. 1.3.

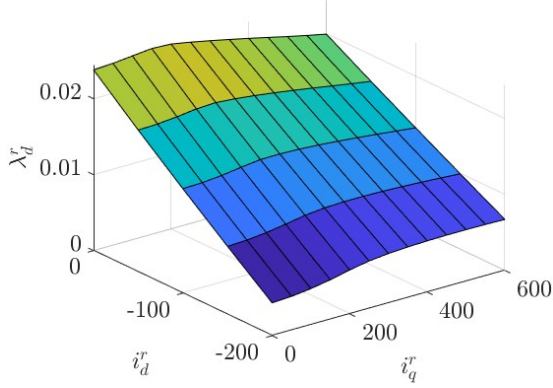


Figure 1.1: λ_d^r of Interior Permanent Magnet Synchronous Machine as a function of i_d^r and i_q^r (generated from FEA simulation)

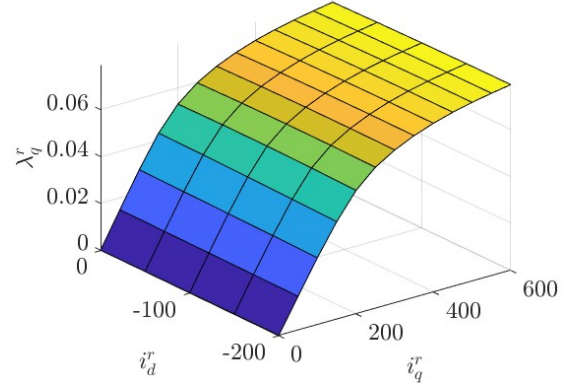


Figure 1.2: λ_q^r of Interior Permanent Magnet Synchronous Machine as a function of i_d^r and i_q^r (generated from FEA simulation)

In [27, 28, 26, 51, 43, 23], the authors highlight the effects of saturation and temperature variations under different operating conditions. From Figs. 1.4-1.9 [27], it is clear that both saturation and temperature variation have a significant impact on the magnetic properties of the machine and the torque production capacity, thereby affecting the overall operation drastically. Furthermore, we can observe from Figs. 1.7-1.9 [27] (where the τ_i 's refer to different temperatures) that different operating regions and electric drive constraints are also affected by temperature variation. The non-linear magnetic properties due to saturation and effects of magnet temperature variation eventually lead to poor response if not accounted for accurately in the controller. Specifically, non-linear magnetic properties due to saturation and effects of temperature variation of PM has a significant impact on:

1. Torque production capacity and
2. Safe electric drive operating regions/system limits.

Thus, it becomes very important to capture these properties of the machine for designing optimal control strategies and develop accurate models for efficient performance.

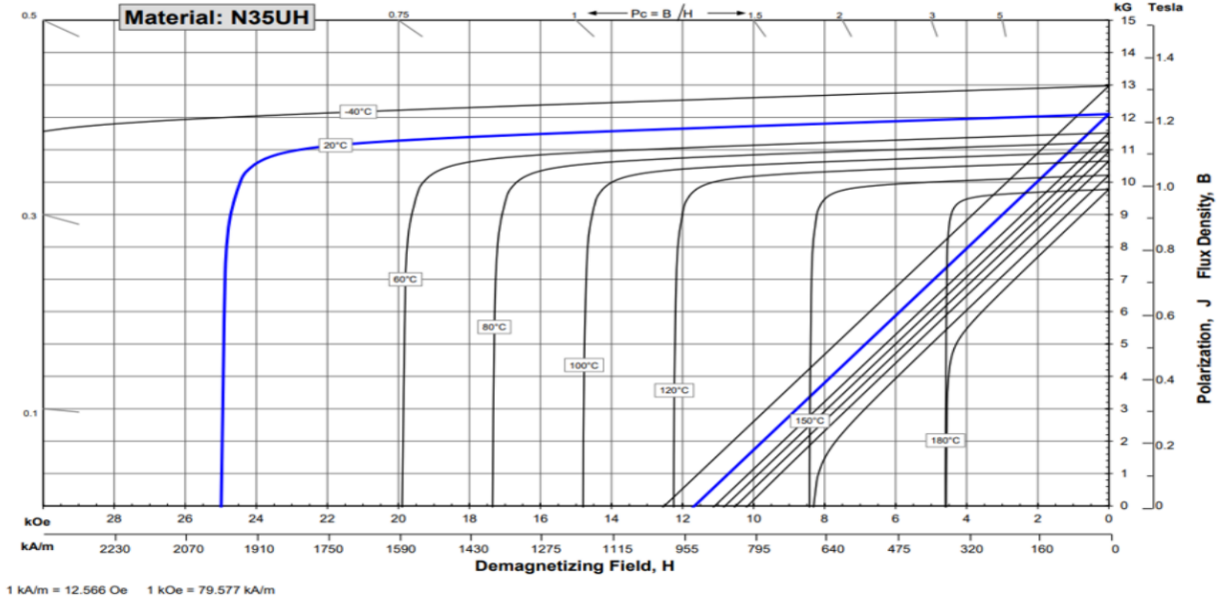


Figure 1.3: $B - H$ curves of NdFeB Permanent Magnet at different temperatures [49]

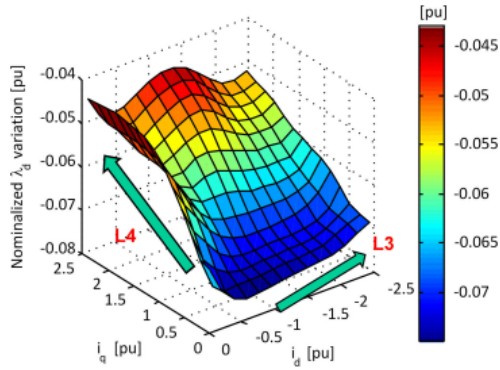


Figure 1.4: Nominalized λ_d^r variation from 50 to 120 °C, per unit [27]

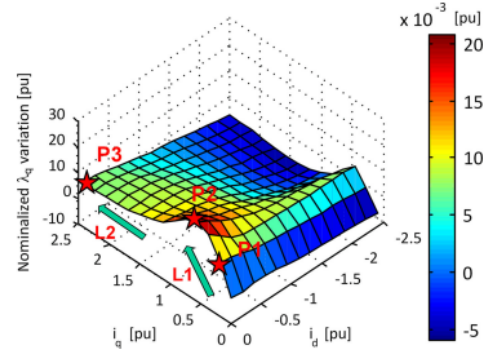


Figure 1.5: Nominalized λ_q^r variation from 50 to 120 °C, per unit [27]

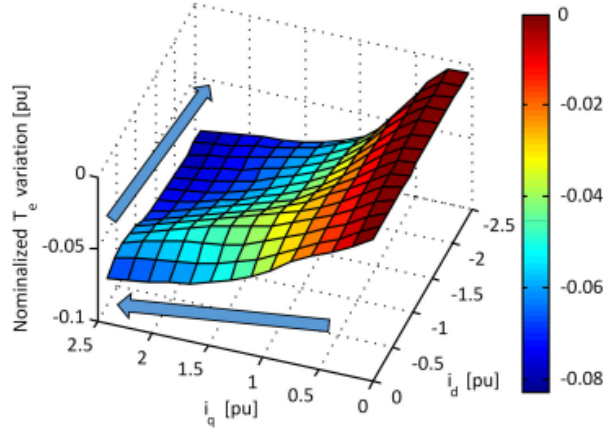


Figure 1.6: Nominalized torque variation from 50 to 120 °C, per unit [27]

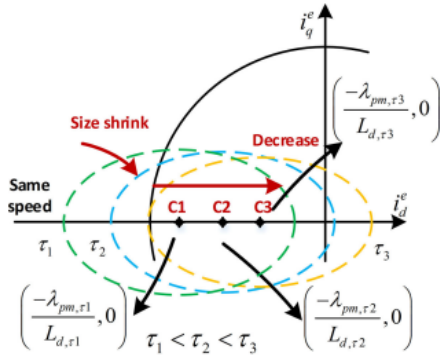


Figure 1.7: Voltage limit ellipses variation of IPM machines for different temperatures (τ_i 's refer to different magnet temperatures)[27]

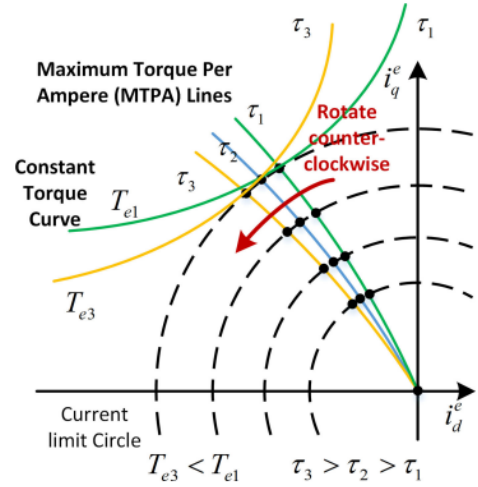


Figure 1.8: MTPA curves variation of IPM machines at different temperatures (τ_i 's refer to different magnet temperatures) [27]

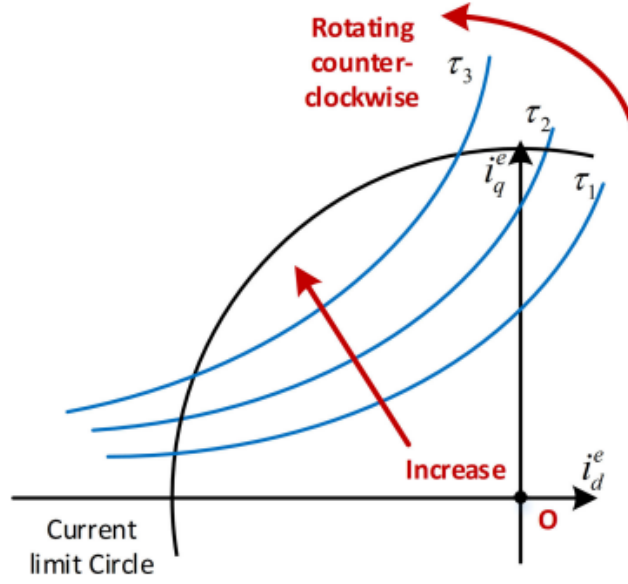


Figure 1.9: Constant torque curves variation of IPM machines for different temperatures (τ_i 's refer to different magnet temperatures) [27]

1.2 Background and Research Gap

As we have established the importance of having accurate models of PMSMs, different models exist for different purposes, such as control, analysis, and design. For the purpose of this thesis, we focus on models of PMSMs appropriate for real-time control purposes.

Magnetics model of PMSMs can be broadly classified as:

- Linear models are typically inaccurate, especially when the machine is magnetically saturated,
- Non-linear lumped-parameter magnetics model, and
- Finite Element Analysis (FEA) models which are accurate but computationally intensive.

The most common control-oriented model for PMSMs is the linear magnetics model. In order to include the effects of temperature variation in linear models the permanent magnet is typically modeled as a flux linkage source that is a function of rotor temperature (Λ_{PM}) [36, 37, 44] as shown below.

$$\vec{\lambda}^r = \mathbf{L} \vec{i}^r + \vec{\lambda}_{pm}^r(T_r); \quad (1.1)$$

$$\begin{aligned} \mathbf{L} &= \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix}, & \vec{\lambda}^r &= \begin{bmatrix} \lambda_d^r \\ \lambda_q^r \end{bmatrix}, & \vec{i}^r &= \begin{bmatrix} i_d^r \\ i_q^r \end{bmatrix}, \\ \vec{\lambda}_{pm}^r(T_r) &= \begin{bmatrix} \Lambda_{PM}(T_r) \\ 0 \end{bmatrix}. \end{aligned} \tag{1.2}$$

As stated previously, linear models are inaccurate when the machine iron is magnetically saturated [44, 33]. Many different control-oriented models of PMSMs exist in the literature that attempt to model the nonlinear magnetic properties under certain assumptions [22, 33, 48, 13, 39, 46, 4, 19, 11, 17, 31, 9]. However, these approaches do not take into account the effects of rotor temperature.

When taking rotor temperature effects into account, we typically focus on the effects of temperature variation on the NdFeB permanent magnet excitation while neglecting the effects of temperature on the soft magnetic material properties. This is because the NdFeB magnets are much more sensitive to changes in temperature because of their lower Curie temperature compared to that of soft magnetic materials [18]. Many approaches that capture rotor temperature effects therefore model the inductances [10, 30] as a function of rotor reference frame currents i_d^r and i_q^r to capture the saturation phenomenon, while the permanent magnet excitation is modeled as a flux linkage source $\Lambda_{PM}(T_r)$ whose value varies with rotor temperature. This separation of magnetic saturation and rotor temperature effects in the model is useful as it allows for the PM flux linkage to be treated as a parameter that can be adaptively estimated in real time. However, it has been shown that this approach becomes inaccurate in regions of heavy saturation and significant variation in rotor temperature [39]. A popularly used NL model from literature is shared below:

$$\vec{\lambda}^r = \mathbf{L}(\vec{i}^r)\vec{i}^r + \vec{\lambda}_{pm}^r(T_r), \tag{1.3}$$

$$\begin{aligned} \mathbf{L}(\vec{i}^r) &= \begin{bmatrix} L_d(i_d^r) & L_{dq}(i_d^r, i_q^r) \\ L_{qd}(i_d^r, i_q^r) & L_q(i_q^r) \end{bmatrix}, \\ \vec{\lambda}_{pm}^r(T_r) &= \begin{bmatrix} \Lambda_{PM}(T_r) \\ 0 \end{bmatrix} \end{aligned} \tag{1.4}$$

where,

$$\begin{aligned}
L_d(i_d^r) &= \frac{\lambda_d^r(i_d^r, 0) - \Lambda_{pm}(T_r)}{i_d^r}, \\
L_q(i_q^r) &= \frac{\lambda_q^r(0, i_q^r)}{i_q^r}, \\
L_{dq}(i_d^r, i_q^r) &= \frac{\lambda_d^r(i_d^r, 0) - \lambda_d^r(i_d^r, i_q^r)}{i_q^r}, \\
L_{qd}(i_d^r, i_q^r) &= \frac{\lambda_q^r(0, i_q^r) - \lambda_q^r(i_d^r, i_q^r)}{i_d^r}
\end{aligned}$$

In an attempt to improve accuracy, [28] proposes forming three-dimensional Look-Up Tables (LUTs) with currents as a function of both flux linkages and rotor temperature, i.e. $\vec{i}^r = f(\vec{\lambda}^r, T_r)$. Another approach is to calculate the inductance matrix as a function of rotor temperature and currents from three-dimensional LUTs. A few authors [26, 8] have also suggested to include compensation terms to take into account rotor temperature variation in the LUT, which primarily captures saturation. All of these approaches suffer from the need for additional model parameters that depend upon rotor temperature. This requires either the knowledge of the rotor temperature, through measurement or estimation, or the application of adaptive techniques that estimate rotor temperature or multiple parameters simultaneously, which relies on the persistent excitation condition to ensure parameter convergence [20].

Another approach to improving accuracy involves using Finite Element Analysis (FEA) models, or reduced-order models formed using FEA [29, 38, 6]. This approach also requires knowledge of rotor temperature. Moreover, obtaining accurate FEA models of actual PMSMs requires knowledge of the dimensional parameters and material properties of the machine, information which is not often available to the controls engineer.

An alternate approach to modeling permanent magnet excitation is through the use of a current source rather than a flux-linkage source, as discussed in [45, 12, 21]. However, the models in these papers all assume linear magnetics, hence they do not address the issue of accurate modeling of the effects of magnetic saturation and rotor temperature variations.

1.3 Contributions

We present a novel magnetics model for PMSMs that accurately captures both the nonlinear and rotor-temperature-dependent behavior of PMSMs by modeling the PM as a current source as opposed to the typical flux linkage source. This is the first time that it is shown

that, by modeling PM as a current source, a nonlinear magnetics model for PMSM can be developed that decouples the nonlinear effects of magnetic saturation and the rotor temperature effect on permanent magnet excitation. In this model, magnetic saturation can be captured through the use of two two-dimensional LUTs, and the permanent magnet excitation can be captured in a single parameter, the permanent magnet current. This parameter can then be adaptively estimated in real time, removing the need for measurement or estimation of the rotor temperature. The result is an accurate model suitable for use in real-time control that works over a broad operating range, and is computationally efficient and relatively simple to implement. Since the proposed model decouples the effects of saturation and temperature variation in the current-fluxlinkage ($\vec{i}^r$ - $\vec{\lambda}^r$) relationship, the model provides accurate torque prediction as opposed to other models, especially in regions where the machine exhibits non-linear behavior.

An adaptive estimation technique for PM excitation is presented leveraging the decoupling feature of magnetic saturation from effects of rotor temperature variation in the proposed NL magnetics model. It has been shown previously that modeling PM excitation as current source is more accurate than the usual way of modeling it has a flux linkage source. The proposed model consists of two components; a non-linear current-flux linkage relationship that captures the saturation phenomenon and PM current parameter that is a function of the rotor temperature which captures the variation in rotor temperature. The proposed adaptive estimation strategy is validated both in simulation and experimental hardware setup for different operating conditions. It is thus demonstrated that the proposed estimation strategy accurately estimates PM current parameter which thereby can enable in precise torque regulation.

Finally, leveraging the proposed model and the adaptive estimation strategy for PM excitation, a comprehensive accurate torque regulator is developed that ensures high performance torque control of PMSM. This proposed torque regulator demonstrates the effectiveness of the proposed model in real-time control purpose. The proposed torque regulator comprises of 4 blocks, namely:

- Torque reference to flux linkage mapping using the proposed NL machine model,
- Flux regulator,
- NL Luenberger observer and
- PM current parameter estimator.

Typically Field Oriented Control (FOC) scheme is developed by implementing a current regulator but we analyze and demonstrate the superiority of implementation of flux regulator

as part of the final contribution. To the best knowledge of the author, this is the first work that develops a FOC scheme solely by flux regulation and experimentally validates the same across the complete operating region of the electric drive. Thus, the main distinction of the proposed torque regulator is the implementation flux regulation as opposed to current regulation. It is shown that with accurate flux linkage estimation, flux regulator provides better transient response and better immunity from noise to control performance that enables better bus voltage utilization thereby leading to a high performance control scheme. The benefits of the proposed torque controller are highlighted using simulation results and the immunity to noise in controller performance is demonstrated using experimental results.

1.4 Outline

The Chapter 2 discusses the basic concepts and the fundamentals on electric drives. The proposed PMSM model is the topic of Chapter 3. The proposed PMSM model is introduced and justified in Section 3.2. A comparison of the proposed model with a representative nonlinear model that uses a PM flux linkage parameter is conducted in 3.4.2 using FEA simulation results for different machine designs. A similar comparison is conducted using experimental data in 3.4.3, followed by concluding remarks in 3.5.

A proposed adaptive estimation strategy for PM excitation is presented in Chapter 4. The proof of convergence of the proposed estimator along with simulation and experimental results are shared in this chapter.

The design architecture of the proposed high performance torque controller is shared in Chapter 5. The chapter also briefly highlights the optimization problem to solve for flux linkage commands using the NL torque mapping from the proposed model. The improvement in controller performance is shown using simulation results and experimental results highlighting the controller performance in presence of noise in the system is also presented.

Future work and further extensions of this thesis is shared in Chapter 6.

CHAPTER 2

Fundamentals on Electric Machines and Drives

2.1 Introduction

This chapter introduces the fundamentals of electric machines and electric drives. A brief summary on the fundamentals of magnetics is provided, followed by the basics of three-phase electric drives. A simple linear model of Permanent Magnet Synchronous Machines (PMSMs) for control applications is presented. Finally, theory on voltage source inverters (VSIs) and field oriented control (FOC) schemes are presented.

2.2 Fundamentals of Magnetism

The magnetic field generated by a current-carrying conductor is determined by “Ampere’s circuital law” as shown below:

$$\oint \vec{H} \cdot d\vec{l} = I \quad (2.1)$$

where I is the total current through a closed loop which is equal to the line integral of magnetic field intensity ($\vec{H}$) over the loop (“Ampere’s loop”). For a magnetic material, Magnetic field intensity and magnetic flux density ($\vec{B}$) are related as follows:

$$\vec{B} = \mu_0 (\vec{H} + \vec{M}) \quad (2.2)$$

where $\vec{M}$ is the magnetization, which is a property of material referring to the density of the magnetic moments in the material, and μ_0 is the permeability of free space. For a “soft”

magnetic material, assuming linear magnetics,

$$\vec{M} = \chi \vec{H} \quad (2.3)$$

$$\vec{B} = \mu_0 \mu_r \vec{H} \quad (2.4)$$

For permanent magnets (PM), i.e. materials where magnetization already exists without any external excitation, the relationship of $\vec{M}$ assuming linear magnetics is given as:

$$\vec{M} = \vec{M}_r + \chi_m \vec{H} \quad (2.5)$$

where $\vec{M}_r$ is the remanent magnetization of the PM. The B-H relationship for PM is given by:

$$\vec{B} = \vec{B}_r + \mu_m \vec{H} \quad (2.6)$$

where $\vec{B}_r$ is the remanent flux density of the PM and μ_m is its effective permeability. The temperature dependence of these magnetic properties will be discussed in the next chapter along with the non-linear behavior of magnetic materials.

2.3 Three-Phase Electric Machines and Drives

An electric drive consists of an electric machine (operating as either a motor or generator), associated power electronic circuitry, and a controller as shown in Fig. 2.1. An electric machine is an electromechanical device that converts electrical energy into mechanical energy, and vice-versa. It typically consists of a rotating part, often referred to as a rotor, and a stationary part called a stator. Three-phase electric machines are not easy to control, especially over a wide speed range. It requires power electronic circuitry to modulate the input electrical power to the machine depending on the region of operation and a controller to regulate the variable of interest (e.g. speed, torque/current).

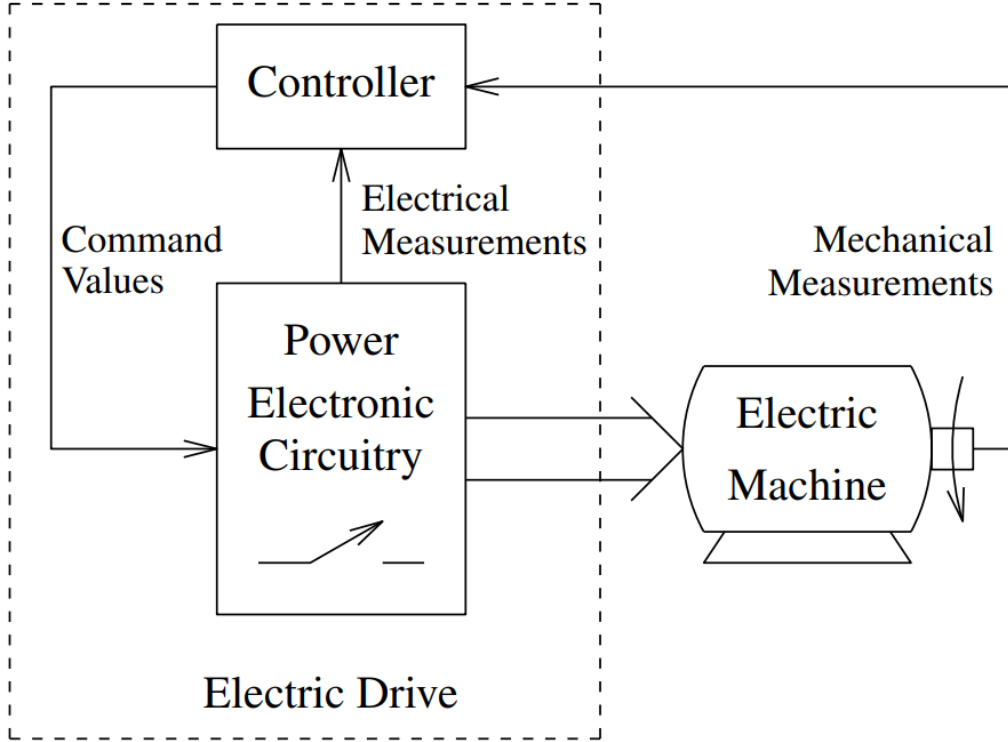


Figure 2.1: Simple magnetic device with permanent magnet [18]

2.4 Transformations

The most commonly followed approach of controlling a three-phase electric machine is to transform the three-phase quantities into equivalent two-phase quantities and a “zero-sequence” component. The zero-sequence component is basically the average value of the three-phase variables and, in balanced operation, the zero-sequence component is zero in case of the machine currents. It is also important to note that the zero sequence component does not contribute to torque production, i.e. the zero-sequence current does not impose a magnetic field on the rotor. This transformation is referred to as the Clarke Transformation and is as follows:

$$\begin{bmatrix} x_d \\ x_q \\ x_0 \end{bmatrix} = \mathbf{T}_{23} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix},$$

where

$$\mathbf{T}_{23} = \begin{bmatrix} \frac{2}{3} & \frac{-1}{3} & \frac{-1}{3} \\ 0 & \frac{\sqrt{3}}{3} & \frac{-\sqrt{3}}{3} \\ \frac{1}{3} & \frac{1}{3} & \frac{1}{3} \end{bmatrix}$$

The Inverse Clarke Transformation can be thus shown as:

$$\begin{aligned} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} &= \mathbf{T}_{23}^{-1} \begin{bmatrix} x_d \\ x_q \\ x_0 \end{bmatrix} \\ &= \mathbf{T}_{32} \begin{bmatrix} x_d \\ x_q \\ x_0 \end{bmatrix}, \end{aligned}$$

where

$$\mathbf{T}_{32} = \begin{bmatrix} 1 & 0 & 1 \\ \frac{-1}{2} & \frac{\sqrt{3}}{2} & 1 \\ \frac{-1}{2} & \frac{-\sqrt{3}}{2} & 1 \end{bmatrix}$$

Another transformation that is commonly used in vector control of AC machines is the Park Transformation. The Park Transformation converts the stationary frame signals (signals that are typically AC in steady state, i.e. sinusoids) into a synchronous reference frame where they are constant in steady state. A typical implementation of Field Oriented Control (FOC) for PMSM entails current regulation in a synchronous reference frame. The Park Transformation and its inverse is given by:

$$\begin{aligned} \begin{bmatrix} x_d^{syn} \\ x_q^{syn} \end{bmatrix} &= \begin{bmatrix} \cos(\theta_{syn}) & \sin(\theta_{syn}) \\ -\sin(\theta_{syn}) & \cos(\theta_{syn}) \end{bmatrix} \begin{bmatrix} x_d \\ x_q \end{bmatrix} \\ \begin{bmatrix} x_d \\ x_q \end{bmatrix} &= \begin{bmatrix} \cos(\theta_{syn}) & -\sin(\theta_{syn}) \\ \sin(\theta_{syn}) & \cos(\theta_{syn}) \end{bmatrix} \begin{bmatrix} x_d^{syn} \\ x_q^{syn} \end{bmatrix} \end{aligned}$$

$$\begin{bmatrix} x_d^r \\ x_q^r \end{bmatrix} = \begin{bmatrix} \cos(\theta_{re}) & \sin(\theta_{re}) \\ -\sin(\theta_{re}) & \cos(\theta_{re}) \end{bmatrix} \begin{bmatrix} x_d \\ x_q \end{bmatrix}$$

$$\theta_{re} = \frac{N_p}{2}\theta_r \quad (2.7)$$

The application of the Clarke transform (converting 3 phase current into equivalent two phase current) and Park transform (converting 2 phase current from stationary frame to synchronous frame) is shown in Fig. 2.2 below. In synchronous machines, the rotor reference frame is typically used as the synchronous frame, i.e.

$$\theta_{syn} = \frac{N_p}{2}\theta_r = \theta_{re}. \quad (2.8)$$

Another notation of the variables in the rotor reference frame (designated by a superscript 'r'), and the variables in the stationary reference frame is as follows:

$$\vec{x}^r = e^{-\mathbf{J}\theta_{re}}\vec{x}, \quad \vec{x} = e^{\mathbf{J}\theta_{re}}\vec{x}^r. \quad (2.9)$$

where

$$\mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$

is the 90° counterclockwise rotation matrix, N_p is the number of poles of the machine, and θ_r is the mechanical angle of the rotor. The matrix exponentials $e^{-\mathbf{J}\theta_{re}}$ and $e^{\mathbf{J}\theta_{re}}$ represent the Park and Inverse Park Transforms, respectively.

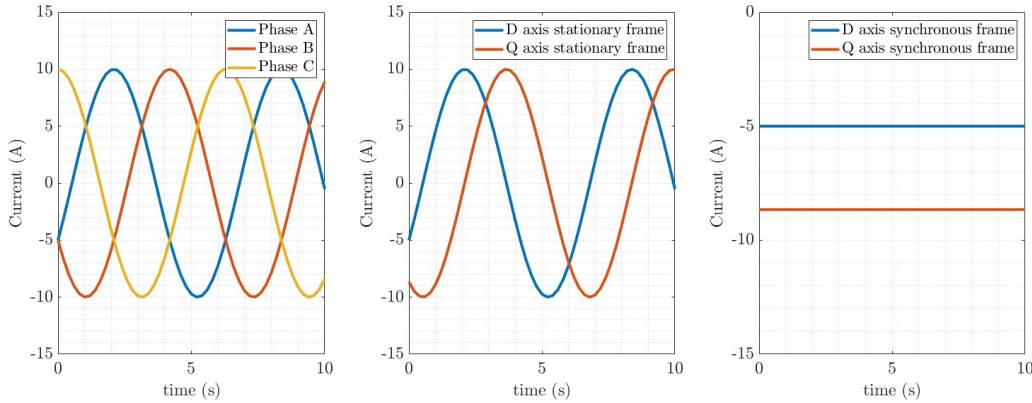


Figure 2.2: Application of Clarke and park transformations, left: three-phase variables, center: two-phase quantities in stationary frame after Clarke transformation, right: two-phase quantities in synchronous reference frame after Park transformation

2.5 Permanent Magnet Synchronous Machines Theory

A smooth-air-gap model of a Permanent Magnet Synchronous Machine (PMSM), as shown in Fig.2.3, is typically considered for the purposes of real-time control implementation. The expressions for a linear model of PMSM is provided in subsequent sections under the following assumptions:

1. Effects due to the slots in the stator core are neglected.
2. Magnetic saturation, core losses and temperature effects are neglected in the linear model.

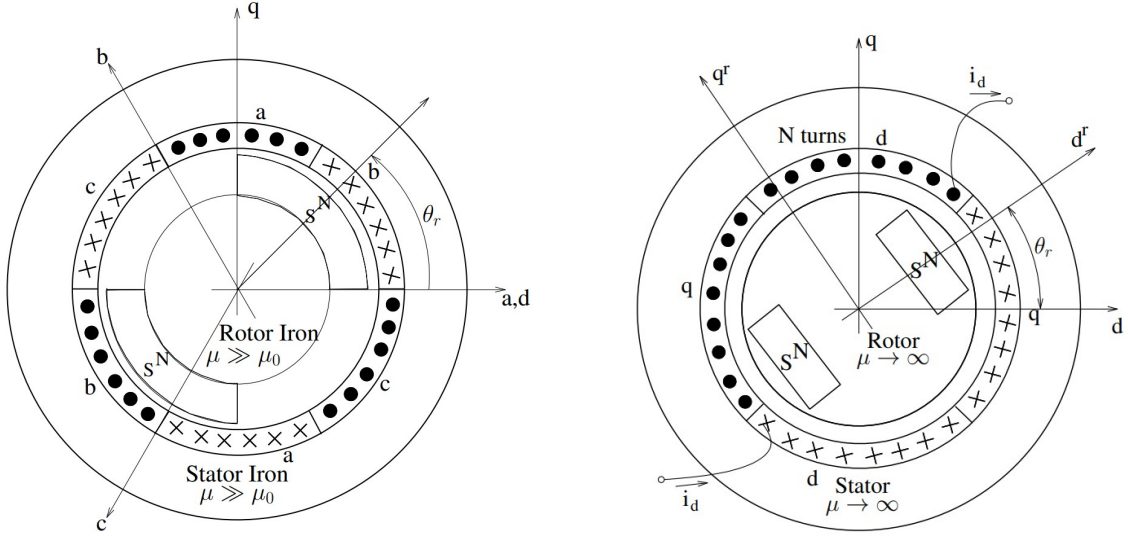


Figure 2.3: Two phase equivalent smooth airgap models of two different PMSM designs, left: Surface Mount Permanent Magnet (SMPM) machine, right: Interior Permanent Magnet (IPM) machine [18]

2.5.1 Flux-Linkage/Current Relationships

The flux linkage-current relationship in the synchronous reference frame (i.e. rotor reference frame) of PMSM is depicted by modeling the PM as a flux linkage source as shown below:

$$\lambda_d^r = L_d i_d^r + \Lambda_{PM}, \quad (2.10)$$

$$\lambda_q^r = L_q i_q^r, \quad (2.11)$$

where L_d and L_q are the inductances along the direct and quadrature axes of the rotor reference frame.

2.5.2 Stator Voltage Equations

The dynamic stator voltages in the rotor reference frame are given by:

$$v_d^r = Ri_d^r - \omega_{re}L_qi_q^r + L_d\frac{di_d^r}{dt}, \quad (2.12)$$

$$v_q^r = Ri_q^r + \omega_{re}(L_d i_d^r + \Lambda_{PM}) + L_q\frac{di_q^r}{dt}, \quad (2.13)$$

where

$$\omega_{re} = \frac{d\theta_{re}}{dt} = \frac{N_p}{2}\omega_r \quad (2.14)$$

is called the electrical rotor velocity and R is the stator resistance.

2.5.3 Electromagnetic Torque

Using the above magnetics model, the three-phase torque is given by [18]:

$$\tau_{3ph} = \frac{3}{2}\tau_{2ph} = \frac{3N_p}{4}\vec{\lambda}^r \otimes \vec{i}^r = \frac{3N_p}{4} \left[\Lambda_{PM}i_q^r + (L_d - L_q)i_d^r i_q^r \right] \quad (2.15)$$

2.6 Power Electronics: Voltage Source Inverters

A power electronic circuit, namely a three-phase voltage-source inverter (Fig. 2.4) is used to generate and drive the PMSM to the desired operating condition. A three-phase voltage-source inverter consists of 3 half bridge circuits comprised of power semiconductor transistors (e.g., MOSFET, IGBT). These power semiconductor transistors are operated as "switches", where upon providing a gate signal the transistor starts conducting current while an absence of a gate signal leads to no flow of current in an ideal scenario. Thus, depending upon the desired operating condition, the three-phase voltage-source inverter is used to supply the necessary excitation in an average-value sense. The transistors are switched at a relatively high frequency (e.g., 10kHz) such that the output value can be varied as desired in an average-value sense. Different modulation schemes exist, of which Space Vector Modulation (SVM) is a commonly used approach.

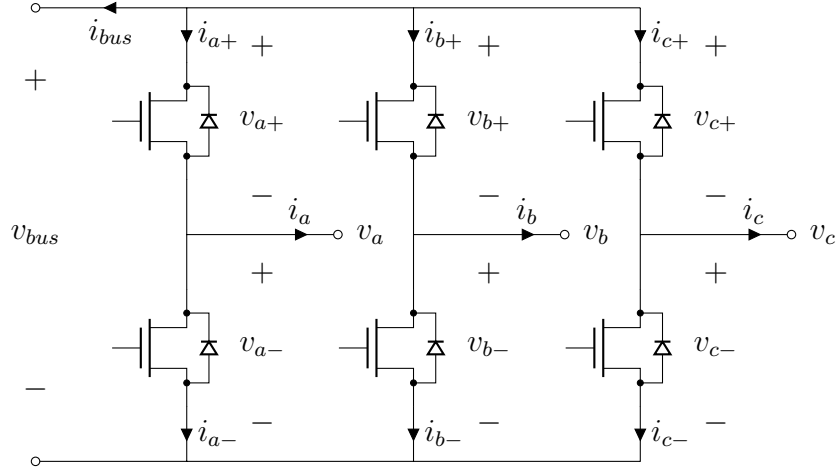


Figure 2.4: Three-phase voltage source inverter

2.7 Control of PMSM machines

A block diagram of the control scheme for a three-phase PMSM machine using its equivalent two-phase model is shown below in Fig. 2.5. The control algorithm is implemented in a synchronous reference frame; a more detailed breakdown of the control algorithm is provided in the next section. Three-phase line currents are measured and transformed into equivalent two-phase currents. The outputs of the controller are voltage commands that are used to compute the duty cycles for each half bridge circuit in the three phase inverter using a modulation scheme (e.g., Space Vector Modulation).

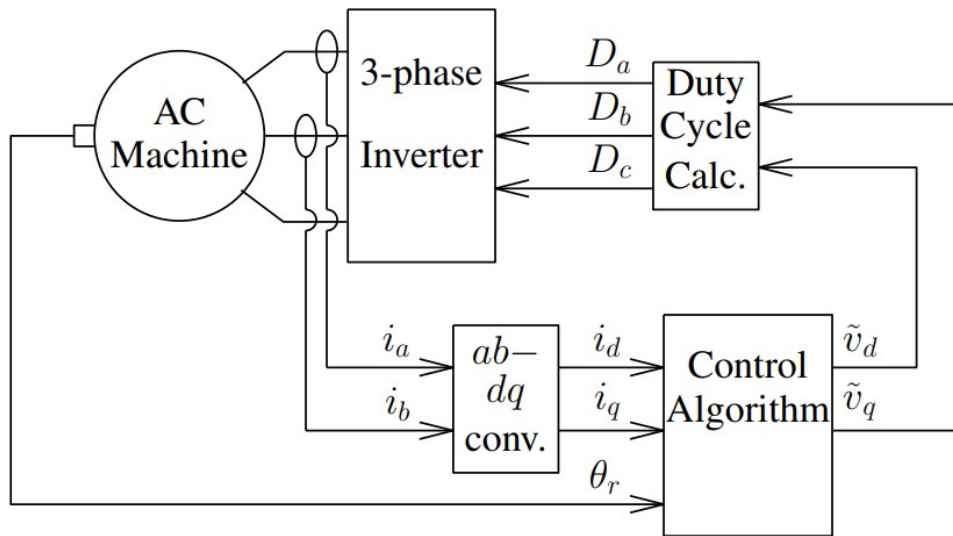


Figure 2.5: Controller block diagram [18]

2.7.1 Field Oriented Control (FOC) Scheme

A simplified block diagram of the FOC algorithm for synchronous machines is shown in Fig. 2.6. The two-phase currents in the stationary frame are transformed into the synchronous reference frame. A synchronous frame current regulator consisting of PI regulators is implemented for each axis of the rotor reference frame. Advanced topics on controller implementation like discrete time implementation, dead-time compensation, speed dependent cross-coupling and rotor angle advancement will be discussed in later chapters.

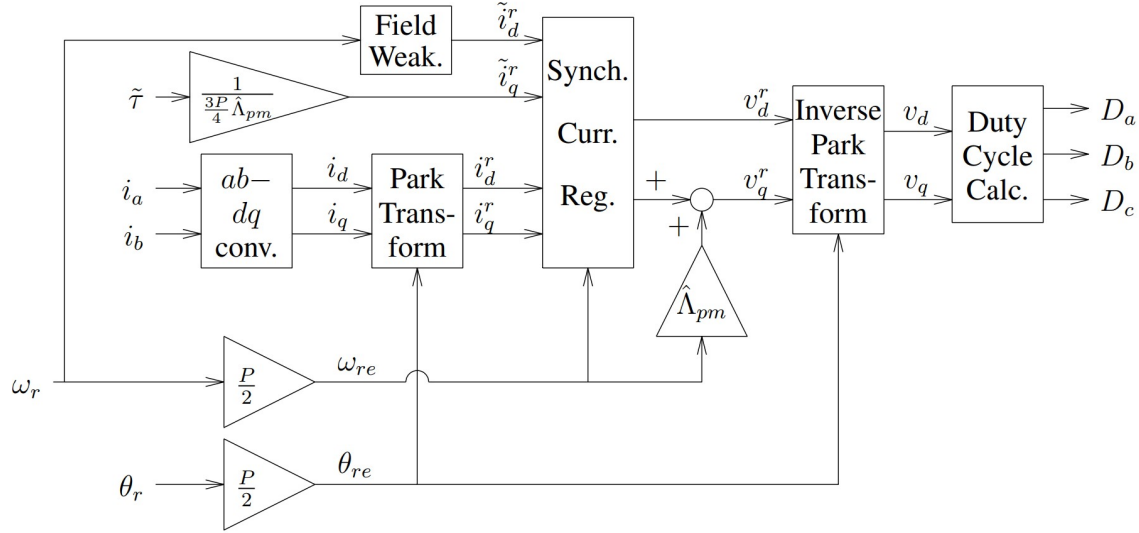


Figure 2.6: Field Oriented Control block diagram [18]

CHAPTER 3

Non-Linear Magnetism Model for Permanent Magnet Synchronous Machines Capturing Saturation and Temperature effects

3.1 Introduction

This chapter presents a nonlinear magnetism model for Permanent Magnet Synchronous Machines that accurately captures the effects of magnetic saturation in the machine iron and variations in rotor temperature on the permanent magnet excitation. The proposed model considers the permanent magnet as a current source rather than the more commonly used flux-linkage source. A comparison of the two modeling approaches is conducted using Finite Element Analysis for different machine designs as well as experimental validation, where it is shown that the proposed model has substantially better accuracy. The proposed model decouples magnetic saturation and rotor temperature effects in the current/flux-linkage relationship, allowing for adaptive estimation of the permanent magnet excitation. Table I compares the structure of the proposed model with models from literature.

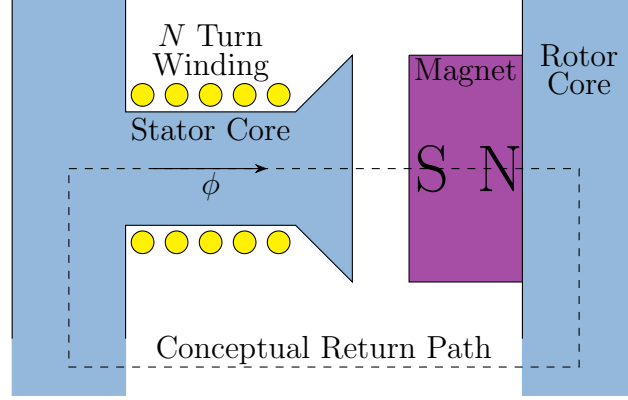


Figure 3.2: Simplified, single-coil magnetics model of PMSM machine

TABLE I
COMPARISON OF STRUCTURE OF THE PROPOSED MODEL WITH MODELS IN LITERATURE

Reference	Model	Effects of Rotor Temperature	Parameter Count	Look-Up Tables	Comments
[20], [21], [22]	Linear	Yes	3 (L_d, L_q and Λ_{pm})	none	Does not capture magnetic saturation
[42], [43], [8]	Linear	No	3 (L_d, L_q and I_{pm})	none	Does not capture magnetic saturation
[24], [23], [25], [26] [27], [28], [29], [30] [31], [32], [33], [34]	NL	No	1 (Λ_{pm})	Two 2D	Does not capture rotor temperature effects
[35], [36]	NL	Yes	1 (Λ_{pm})	Two 2D	Inaccurate in the region of deep magnetic saturation and significant rotor temperature variation [27]
[16], [37]	NL	Yes	-	Minimum of two 3D	Requires higher dimensional LUT; Rotor temperature needs to be measured or estimated
Proposed model	NL	Yes	1 (I_{pm})	Two 2D	Decouples magnetic saturation and effects of rotor temperature variation; Knowledge of rotor temperature unnecessary through online I_{pm} estimation

Figure 3.1: Comparison of structure of the proposed model with models in literature

3.2 Motivation

The motivation for the proposed model is illustrated by considering the single-coil magnetic device shown in Fig. 3.3 which resembles a simplified model of a PMSM.

We begin by creating a magnetic circuit model of the device. The N -turn coil is represented by a magnetomotive force (MMF) source Ni , where i is the current flowing in the coil. The stator and rotor cores consist of a soft magnetic material exhibiting magnetic saturation, and so are represented by a nonlinear reluctance $\mathcal{R}_c(\phi)$ whose value varies with flux level ϕ . The PM is assumed to possess linear magnetic properties with an incremental permeability μ_m and a remanent flux density B_r that varies with rotor temperature T_r , and so it is modeled as a constant reluctance $\mathcal{R}_{pm}$ and flux source $\Phi_{pm}(T_r)$, as shown in Fig. 3.3.

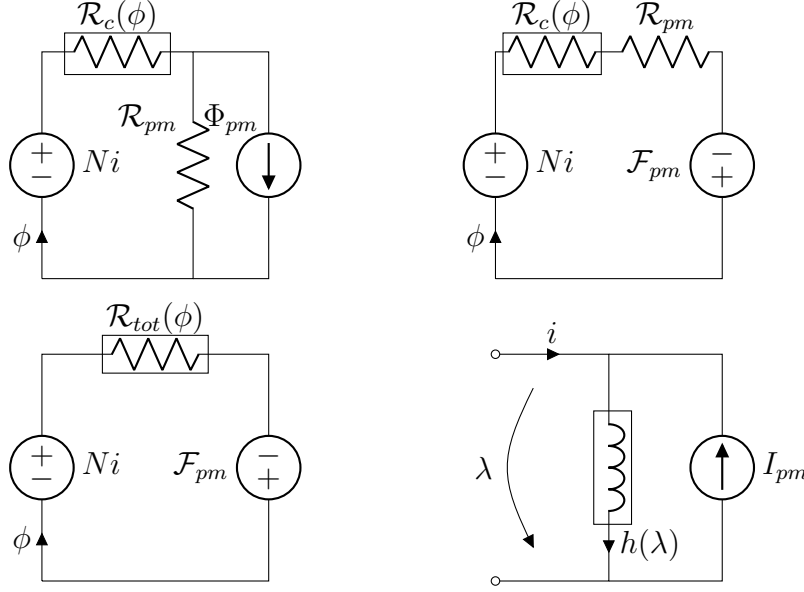


Figure 3.3: Derivation of proposed magnetics model for single-coil device. Top left: magnetic circuit model. Top right: Thevenin equivalent of permanent magnet. Bottom left: Combination of magnet and core reluctances. Bottom right: Resulting electric circuit model.

An equivalent magnetic circuit model can be created using the Thevenin equivalent of the PM, where the PM excitation is modelled as MMF source $\mathcal{F}_{pm}(T_r) = \mathcal{R}_{pm}\Phi_{pm}(T_r)$ rather than a flux source. Simplifying the magnetic circuit further, the two reluctances are now in series and can be combined into a single nonlinear reluctance, $\mathcal{R}_{tot}(\phi)$. Analysis of the resulting magnetic circuit yields:

$$Ni + \mathcal{F}_{pm}(T_r) = \mathcal{R}_{tot}(\phi)\phi. \quad (3.1)$$

If we denote $f(\phi) = \mathcal{R}_{tot}(\phi)\phi$, and note that $\phi = \frac{\lambda}{N}$, we can represent the current as follows:

$$i = \frac{1}{N} \left(f \left(\frac{\lambda}{N} \right) - \mathcal{F}_{pm}(T_r) \right) \quad (3.2)$$

$$= h(\lambda) - I_{pm}(T_r). \quad (3.3)$$

This expression can be represented in an electrical circuit model as a nonlinear inductance, with current/flux-linkage relationship $i = h(\lambda)$, in parallel with a current source $I_{pm}(T_r)$ representing the PM excitation, as shown in Fig. 3.3. As the value of I_{pm} is directly proportional to the remanent flux density B_r of the magnet, it will share the same dependence upon the rotor temperature.

The same expression (3.3) can be derived by following a more detailed analysis of the

single coil scalar model, as shared below where the non-linear magnetic behavior of the core and linear magnetic property of the permanent magnet are:

$$B_c(H_c) = \mu_0(H_c + M(H_c)) \quad (3.4)$$

$$B_{PM} = \mu_0 H_{PM} + B_r(T_r) \quad (3.5)$$

$$B_g = \mu_0 H_g \quad (3.6)$$

where $B_c(H_c)$ and H_{PM} represent magnetic flux density of the core as function of magnetic field intensity and magnetic field intensity of permanent magnet respectively. Since we know magnetic flux density as a function of magnetic field intensity $B_c(H_c)$ (Fig. 3.4), and assuming no core loss in the system, we can calculate the inverse function $H_c(B_c)$ (Fig. 3.5), as the slope of the B-H curve is always positive. It is important to note that $M_c(B_c)$ denotes the magnetization of the core as function of flux density. Typically, we denote magnetization as a function of magnetic field intensity $M_c(H_c)$ (Fig. 3.6), but since we know magnetic flux density as a function of magnetic field intensity $B_c(H_c)$, and assuming no core loss, we can calculate magnetization as function of flux density, $M_c(B_c)$ (Fig. 3.7). Let i denote the

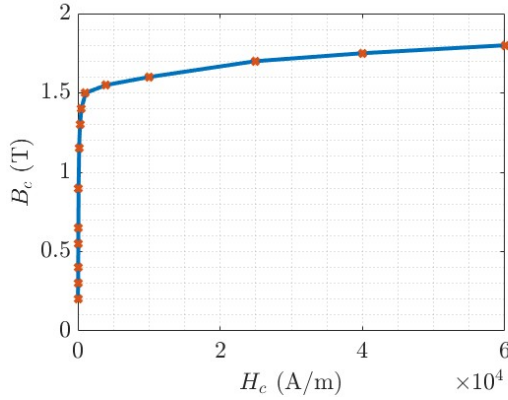


Figure 3.4: B-H curve of electrical steel

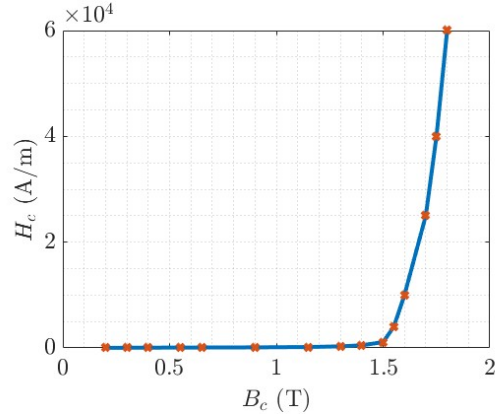


Figure 3.5: H-B curve of electrical steel

total current flowing through N turns of the coil and l_c , l_g and l_{pm} represent the effective length of core, air-gap and PM. The cross sectional area is given by A_c . Assuming no leakage flux, the flux density in the air gap, core and PM is the same, that is $B_g = B_c = B_{PM} = B$.

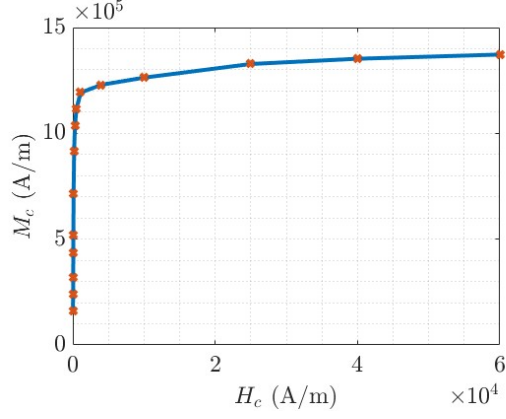


Figure 3.6: M-H curve of electrical steel

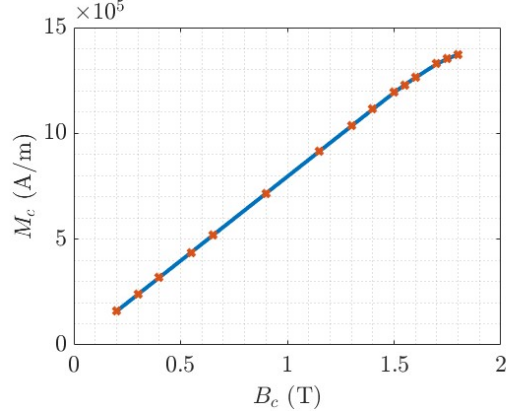


Figure 3.7: M-B curve of electrical steel

Applying Ampere's law over the Amperian loop as shown, we get:

$$Ni = H_c l_c + H_g l_g + H_{PM} l_{PM} \quad (3.7)$$

$$\begin{aligned} &= \left(\frac{B}{\mu_0} - M_c(B) \right) l_c + \frac{B}{\mu_0} l_g + \left(\frac{B - B_r(T_r)}{\mu_0} \right) l_{PM} \\ &= H_c(B) l_c + \frac{B}{\mu_0} l_g + \left(\frac{B - B_r(T_r)}{\mu_0} \right) l_{PM} \end{aligned} \quad (3.8)$$

Rearranging the above equation, we obtain:

$$i = \left(\frac{l_g + l_{PM}}{\mu_0 N^2 A_c} \right) \lambda - \frac{H \left(\frac{\lambda}{N A_c} \right)}{N} - \frac{B_r(T_r)}{N \mu_0} l_{PM}, \quad (3.9)$$

$$i = \left(\frac{l_c + l_g + l_{PM}}{\mu_0 N^2 A_c} \right) \lambda - M \left(\frac{\lambda}{N A_c} \right) - \frac{B_r(T_r)}{N \mu_0} l_{PM}, \quad (3.10)$$

$$i = h(\lambda) - I_{pm}(T_r). \quad (3.11)$$

where,

$$I_{pm}(T_r) = \frac{B_r(T_r)}{N \mu_0} l_{PM} \quad (3.12)$$

In order to illustrate the advantages of the proposed model, we consider a model consisting of a nonlinear inductance and the more commonly used flux-linkage source representing permanent magnet excitation.

$$\lambda = L(i)i + \Lambda_{pm}(T_r). \quad (3.13)$$

The key distinction between the two models is how the current/flux-linkage relationships react to changes in rotor temperature (and hence the magnet temperature). In the case of the model (3.13) with a PM flux linkage source, the result of a rotor temperature change would be a constant offset $\Delta\lambda$ in the $\lambda - i$ curve. In the case of the proposed model, a change in rotor temperature would result in a constant offset ΔI of the curve.

Figure 3.8 shows a representative flux linkage/current relationship at two different rotor temperatures. One figure plots flux linkage versus current, the other current versus flux linkage, both represent the same data and model the eq. 3.13-3.11 respectively. As can be

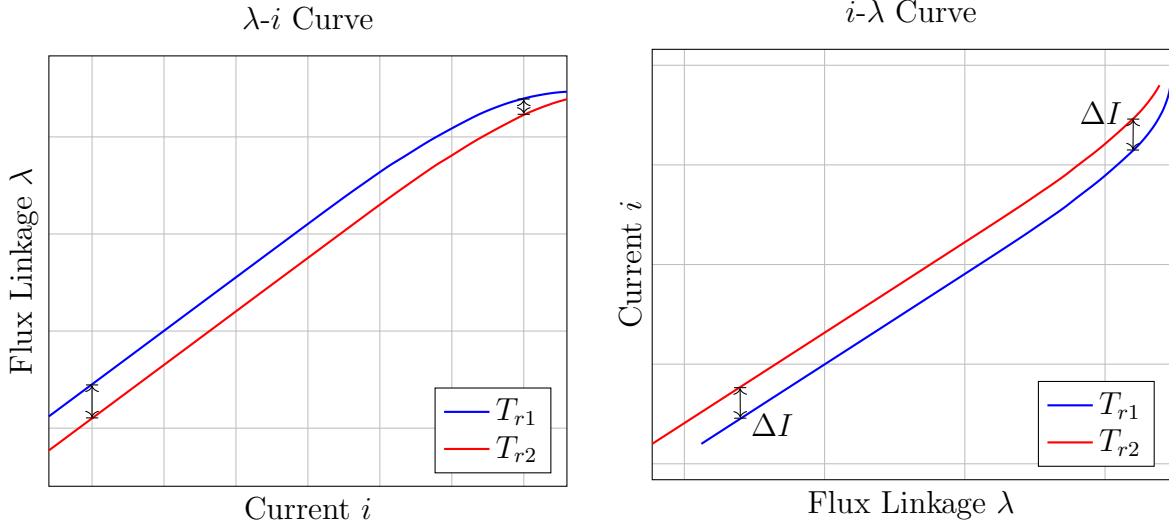


Figure 3.8: Representative flux-linkage/current relationship capturing magnetic saturation at two different rotor temperatures. Top: $\lambda - i$ curve. Bottom: $i - \lambda$ curve.

seen, the variation of the $\lambda - i$ curve with temperature is not captured well by a constant offset in the flux linkage as modeled in Eq. (3.13). This is because the saturation effect results in a convergence of the flux-linkage values at high current excitation. However, the variation of the $i - \lambda$ curve is well captured by a constant offset in current as seen in Fig. 3.8(b) and hence with the model in Eq. (3.3).

3.3 Proposed Model

The proposed equivalent two-phase model for a Permanent Magnet Synchronous Machine (PMSM) is developed considering the standard smooth airgap model (i.e., effects due to the slots in the stator core are neglected).

3.3.1 Current/flux-linkage relationship

The current/flux-linkage relationships of the proposed machine model are presented in the rotor reference frame (designated by a superscript 'r'). The current/flux-linkage relationships are an extension of the single-coil device discussed in the prequel to a vector representation:

$$\vec{i}^r = \vec{h}(\vec{\lambda}^r) - \vec{i}_{pm}^r(T_r) \quad (3.14)$$

or

$$\begin{bmatrix} i_d^r \\ i_q^r \end{bmatrix} = \begin{bmatrix} h_d(\lambda_d^r, \lambda_q^r) \\ h_q(\lambda_d^r, \lambda_q^r) \end{bmatrix} - \begin{bmatrix} I_{pm}(T_r) \\ 0 \end{bmatrix} \quad (3.15)$$

where I_{pm} represents the permanent magnet excitation.

3.3.2 Properties of NL current/flux linkage relationship in rotor reference frame

The mapping $\vec{h}(\vec{\lambda}^r)$ captures the nonlinear effects of the saturation of soft magnetic materials of the machine and so has the following property

$$\vec{h}(\vec{0}) = \vec{0}. \quad (3.16)$$

If the electric machine design has the standard symmetries, $\vec{h}(\vec{\lambda}^r)$ also has the following properties:

$$h_q(\lambda_d^r, -\lambda_q^r) = -h_q(\lambda_d^r, \lambda_q^r) \quad (3.17)$$

$$h_q(-\lambda_d^r, \lambda_q^r) = h_q(\lambda_d^r, \lambda_q^r) \quad (3.18)$$

$$h_d(\lambda_d^r, -\lambda_q^r) = h_d(\lambda_d^r, \lambda_q^r) \quad (3.19)$$

Assuming standard magnetic material properties (e.g., incremental permeabilities are always positive), it can be shown that $\vec{h}(\vec{\lambda}^r)$ is injective and surjective, and so an inverse flux-linkage/current relationship exists:

$$\vec{\lambda}^r = \vec{h}^{-1}(\vec{i}^r + \vec{i}_{pm}^r) = \vec{g}(\vec{i}^r + \vec{i}_{pm}^r). \quad (3.20)$$

It can also be shown that, for $\alpha_1, \beta_1 \geq 0$

$$0 \leq \alpha_1 \|\vec{x}\|^2 \leq \vec{x}^T \frac{d\vec{h}(\vec{x})}{d\vec{x}} \vec{x} \leq \beta_1 \|\vec{x}\|^2 \leq \infty \quad (3.21)$$

Example implementations of the proposed magnetics model and its inverse relationship using two-dimensional lookup tables are shown in Fig. 3.9. Techniques for determining the datapoints in the lookup tables and the PM current as a function of rotor temperature will be discussed in detail in the sequel. As can be seen in these implementations, the computational complexity of the proposed model is such that it is suitable for use in real-time control applications. More efficient models can be created by exploiting the symmetries of the magnetic behavior discussed earlier.

3.3.3 Stator voltage dynamic equations in rotor reference frame

It can be shown that the dynamic stator voltages can be presented in vector and scalar form as follows:

$$\vec{v}^r = R\vec{i}^r + \frac{N_p}{2}\omega_r \mathbf{J} \vec{\lambda}^r + \frac{d\vec{\lambda}^r}{dt} \quad (3.22)$$

$$= R \left[\vec{h}(\vec{\lambda}^r) - \vec{i}_{pm}^r \right] + \frac{N_p}{2}\omega_r \mathbf{J} \vec{\lambda}^r + \frac{d\vec{\lambda}^r}{dt}, \quad (3.23)$$

$$v_d^r = R \left[h_d(\vec{\lambda}^r) - I_{pm} \right] - \frac{N_p}{2}\omega_r \lambda_q^r + \frac{d\lambda_d^r}{dt}, \quad (3.24)$$

$$v_q^r = R h_q(\vec{\lambda}^r) + \frac{N_p}{2}\omega_r \lambda_d^r + \frac{d\lambda_q^r}{dt}$$

where ω_r is the angular velocity of the rotor and R is the winding resistance.

3.3.4 Electromagnetic Torque

To derive the torque expression, we begin with an expression for the stored magnetic field energy w_{fld} in a three-phase PMSM machine [15]. To simplify the analysis we leave out a (position-independent) energy component due to the magnets, which does not impact the

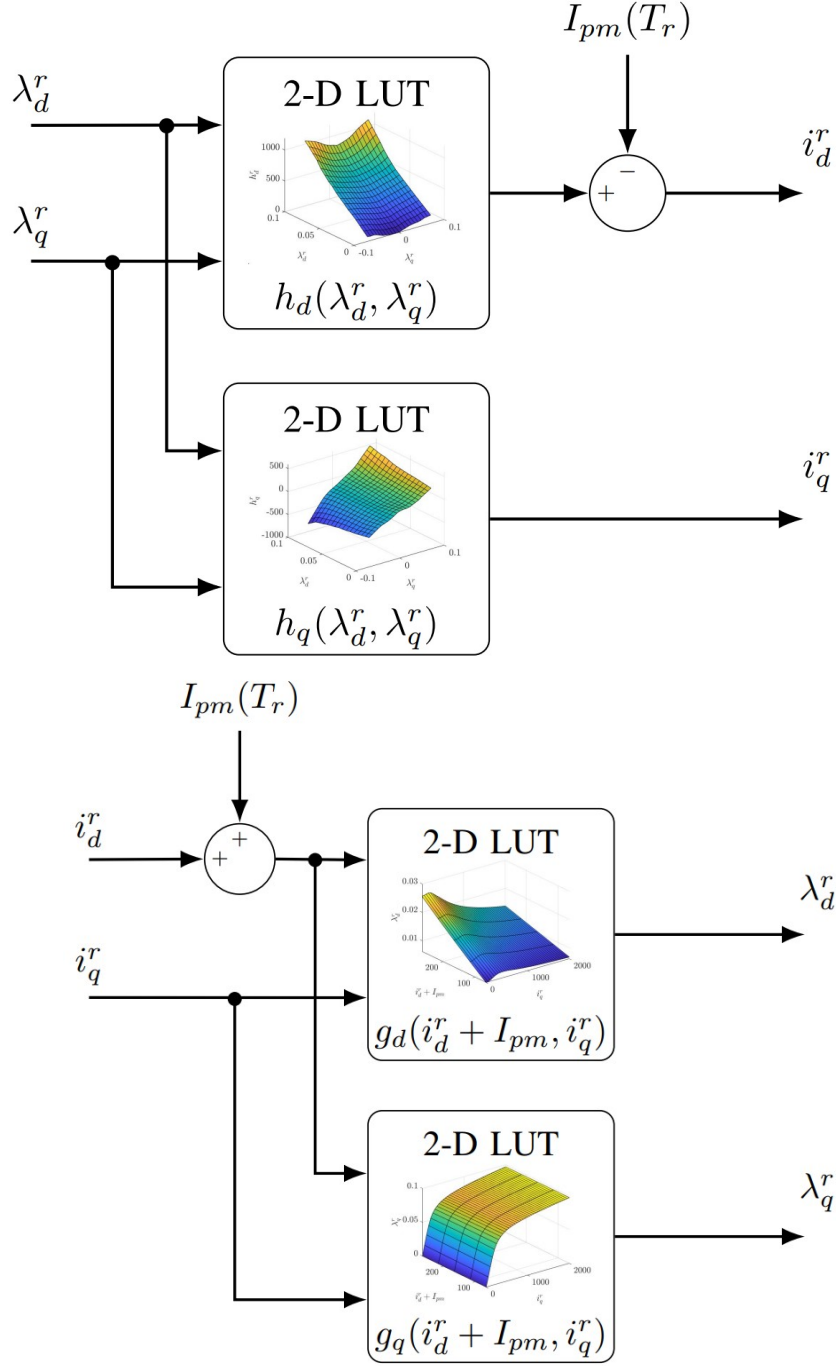


Figure 3.9: Example implementations of the proposed magnetics model using two-dimensional lookup tables. Top: flux linkage to current mapping. Bottom: current to flux linkage mapping

final result.

$$dw_{fld} = \vec{i}^T d\vec{\lambda} \quad (3.25)$$

$$\begin{aligned} w_{fld}(\vec{\lambda}, \theta_r) &= W_{pm\Phi} + \frac{3}{2} \int_0^{\vec{\lambda}} \vec{i}^T(\vec{x}) d\vec{x} \\ &= \frac{3}{2} \int_0^{\vec{\lambda}} \left\{ e^{\mathbf{J}\theta_{re}} \left[\vec{h}(e^{-\mathbf{J}\theta_{re}} \vec{x}) - \vec{i}_{pm}^r \right] \right\}^T d\vec{x}. \end{aligned} \quad (3.26)$$

The electromagnetic torque can therefore be shown to be:

$$\begin{aligned} \tau_{em} &= - \frac{\partial w_{fld}(\vec{\lambda}, \theta_r)}{\partial \theta_r} \\ &= - \frac{3N_p}{4} \int_0^{\vec{\lambda}} \frac{\partial}{\partial \theta_{re}} \left(\vec{i}^T(\vec{\lambda}') \right) d\vec{\lambda}' \end{aligned} \quad (3.27)$$

$$= \frac{3N_p}{4} \left\{ \left(e^{-\mathbf{J}\theta_{re}} \vec{\lambda} \right)^T \mathbf{J} \vec{h}(e^{-\mathbf{J}\theta_{re}} \vec{\lambda}) + \left(e^{-\mathbf{J}\theta_{re}} \vec{\lambda} \right)^T \mathbf{J} \vec{i}_{pm}^r \right\} \quad (3.28)$$

$$\begin{aligned} &= \frac{3N_p}{4} \vec{\lambda}^r \otimes \vec{i}^r = \frac{3N_p}{4} \left[- \left(\vec{\lambda}^r \right)^T \mathbf{J} \vec{h}(\vec{\lambda}^r) + \vec{\lambda}^r \mathbf{J} \vec{i}_{pm}^r \right] \\ &= \frac{3N_p}{4} \left[h_q(\vec{\lambda}^r) \lambda_d^r - h_d(\vec{\lambda}^r) \lambda_q^r + I_{pm}(T_r) \lambda_q^r \right], \end{aligned} \quad (3.29)$$

where $W_{pm\Phi}$ is the field energy due to PM that is independent of rotor angle, the first term in (3.29) corresponds to a (nonlinear) reluctance torque and the second term is the PM torque.

3.4 Validation and Comparison Results

In this section we will illustrate the accuracy of the proposed model through comparison with a "baseline" PM flux-linkage model. We will first introduce the baseline model. We will then compare how changes in the rotor temperature affect the accuracy of the proposed and baseline models. The first comparisons will be conducted using FEA results for three different machine designs: an interior permanent machine, a distributed-winding surface mount permanent magnet (SMPM) machine, and a concentrated-winding SMPM machine. Finally, a comparison will be conducted using experimental results.

3.4.1 Baseline model

We will use the following nonlinear PM flux linkage model as a baseline for comparison with the proposed model:

$$\vec{\lambda}^r = \vec{f}(\vec{i}^r) + \vec{\lambda}_{pm}^r(T_r), \quad (3.30)$$

where

$$\vec{f}(\vec{i}^r) = \begin{bmatrix} f_d(i_d^r, i_q^r) \\ f_q(i_q^r, i_d^r) \end{bmatrix}, \quad \vec{\lambda}_{pm}^r(T_r) = \begin{bmatrix} \Lambda_{PM}(T_r) \\ 0 \end{bmatrix}$$

and the torque is given by

$$\begin{aligned} \tau_{em} &= \frac{3N_p}{4} [\lambda_d^r i_q^r - \lambda_q^r i_d^r], \\ &= \frac{3N_p}{4} [f_d(i_d^r, i_q^r) i_q^r + \Lambda_{pm}(T_r) i_q^r - f_q(i_d^r, i_q^r) i_d^r]. \end{aligned} \quad (3.31)$$

This particular model was chosen for comparison as it uses the conventional permanent magnet flux linkage source to capture the permanent magnet excitation, yet separates the effects of non-linear saturation behavior and rotor temperature variation in a similar way as the proposed nonlinear PM current model.

3.4.2 Model comparison using FEA-based data

We first use two-dimensional finite element analysis (FEA) results from ANSYS as the basis for comparison. The process used to compare the two models is as follows:

1. Using FEA, calculate the flux linkages λ_d^r and λ_q^r and the electromagnetic torque τ_{em} over a range of currents i_d^r and i_q^r with the rotor temperature set at $T_1 = 20^\circ C$.
2. Use the flux-linkage/current data to determine the values of $I_{pm}(T_1)$ (used in the proposed model) and $\Lambda_{pm}(T_1)$ (used in the baseline model). The value of Λ_{pm} is determined from λ_d^r by setting the currents to zero, and the value of I_{pm} is determined by finding the value of i_d^r that sets $\lambda_d^r = 0$.
3. Generate a look-up table for the $\vec{\lambda}^r = \vec{g}(\vec{i}^r + \vec{i}_{pm}^r)$ mapping of the proposed model using $20^\circ C$ FEA data. Note that we determine the $\vec{g}(\cdot)$ mapping rather than the $\vec{h}(\cdot)$ mapping to better enable the comparison.

4. Generate a look-up table for the $\vec{f}(\vec{i}^r)$ in (3.30) for the baseline model using the $20^\circ C$ FEA data.
5. Repeat steps 1) and 2) for a different rotor temperature (T_2).
6. Compute the flux linkages and torques with the proposed model using the nonlinear look-up table for $\vec{g}(\cdot)$ determined from the T_1 data and $I_{pm}(T_2)$ for the proposed model.
7. Compute the flux linkages and torques with the comparison model using the nonlinear look-up table $\vec{f}(\vec{i}^r)$ from the T_1 data and $\Lambda_{pm}(T_2)$ for the baseline model.
8. Compare the results from the previous steps 6) and 7) with the FEA results at the new rotor temperature (T_2).

3.4.2.1 Interior permanent magnet machine

We first consider the interior permanent magnet machine design presented in [16] and shown in Fig. 3.10a. The permanent magnet material used is NdFeB N35UH from Arnold Magnetics [3]. The electrical steel material used in the machine model is M-19 29 Ga. The mechanical dimensions and winding parameters are kept the same as in [16]. With the recent advancements in cooling mechanisms as shared in [1], peak current densities of $48 \text{ A}_{rms}/\text{mm}^2$ can be achieved. We use this value to determine the maximum current of the mapping, though we note that this machine was not specifically designed for these current densities. An example magnetic flux density distribution for this machine design is shown in Fig. 3.10b.

The NL $\vec{\lambda}^r - \vec{i}^r$ mapping, determined using FEA at temperatures $T_1=20^\circ C$ and $T_2=100^\circ C$, is shown in Fig. 3.11.

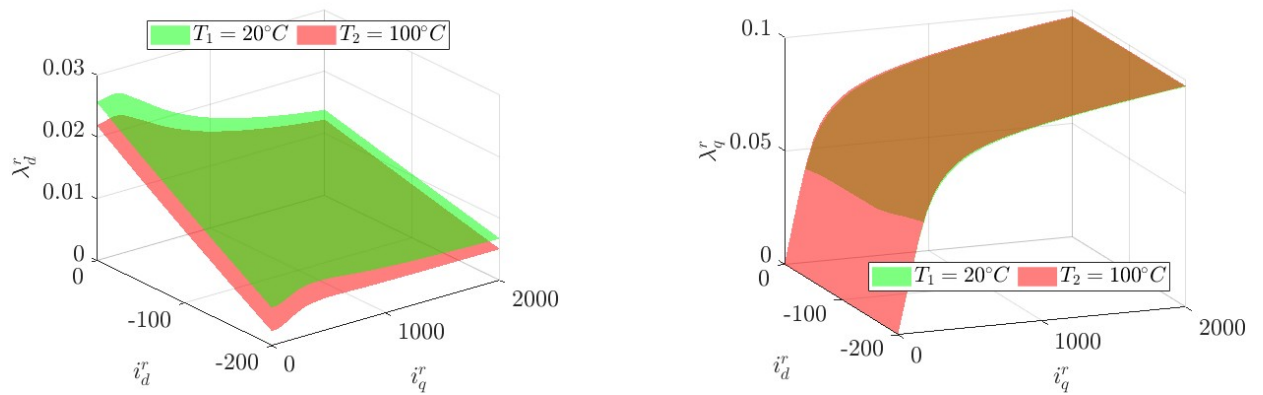
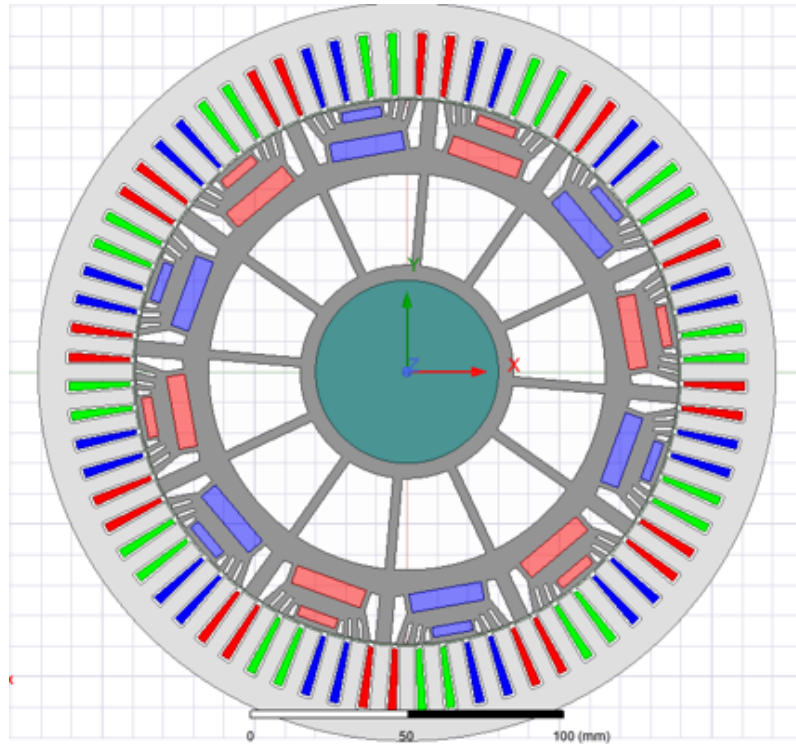
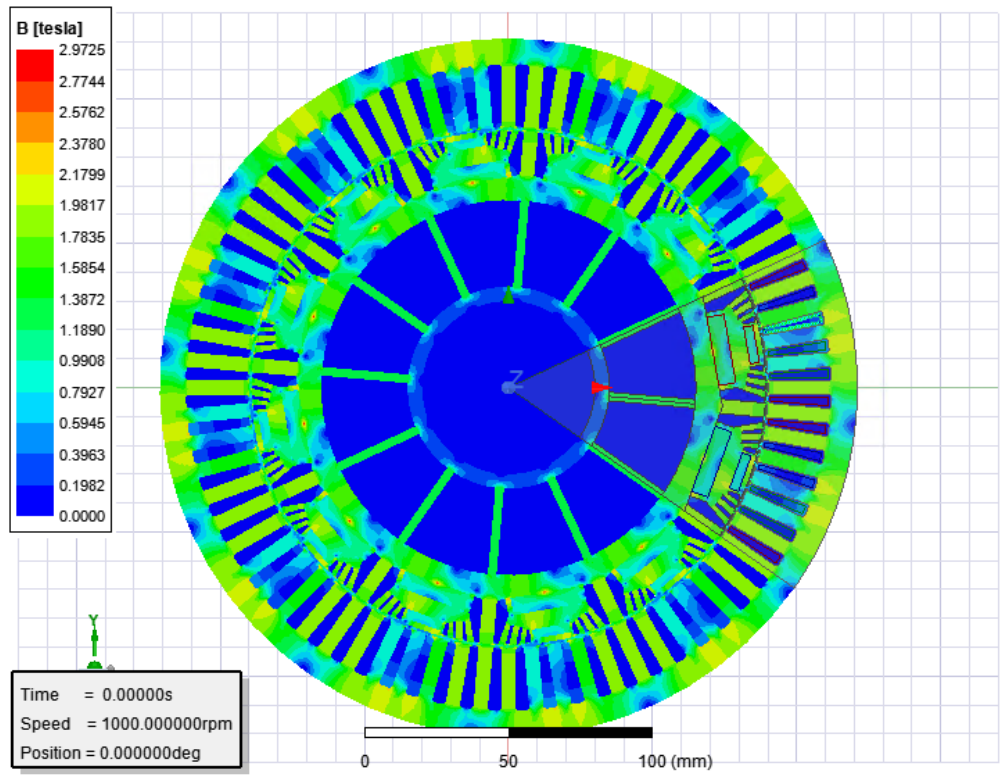


Figure 3.11: $\vec{\lambda}^r - \vec{i}^r$ mapping of interior permanent magnet machine at different rotor temperatures (FEA results). Left: λ_d^r mapping. Right: λ_q^r mapping



(a)



(b)

Figure 3.10: (a) Cross-section of Interior Permanent Magnet machine [16], (b) Flux density plot calculated using FEA

As is to be expected, we see a significant change in λ_d^r whereas λ_q^r is not substantially different. The value of I_{pm} and Λ_{pm} was determined for multiple temperatures and is shown in Table. 3.1.

Table 3.1: I_{pm} and Λ_{pm} values for Interior Permanent Magnet machine

Rotor Temperature ($^{\circ}C$)	I_{pm} (A)	$\Lambda_{pm}(V - s)$
20	264.8	0.0256
60	245.6	0.0238
80	235.1	0.0227
100	225.4	0.0219

The absolute errors in λ_d^r and λ_q^r for both the proposed PM current model and the baseline PM flux linkage model are shown in Figs. 3.12 - 3.13.

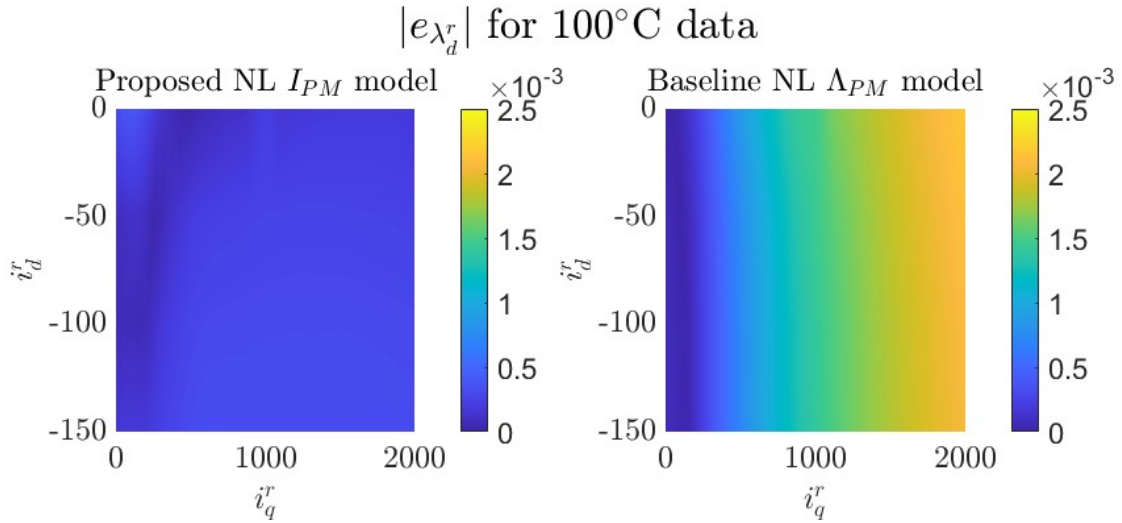


Figure 3.12: Absolute errors in λ_d^r for Interior Permanent Magnet machine at $T = 100^{\circ}C$ for the proposed (left) and baseline (right) models, FEA results.

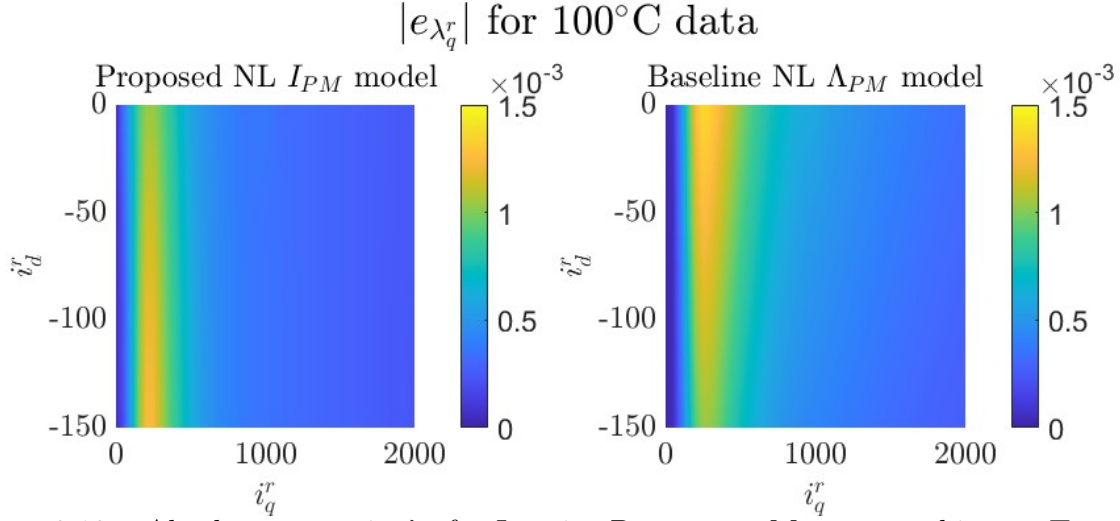


Figure 3.13: Absolute errors in λ_q^r for Interior Permanent Magnet machine at $T = 100^\circ\text{C}$ for the proposed (left) and baseline (right) models, FEA results.

It can be seen that the baseline PM flux linkage model experiences significantly larger error in λ_d^r at high values of i_q^r when compared to the proposed PM current model. In the case of λ_q^r , the model errors are similar, and occur at the onset of magnetic saturation on the quadrature axis.

The relative error in torque for the two models is highlighted in Fig. 3.14.

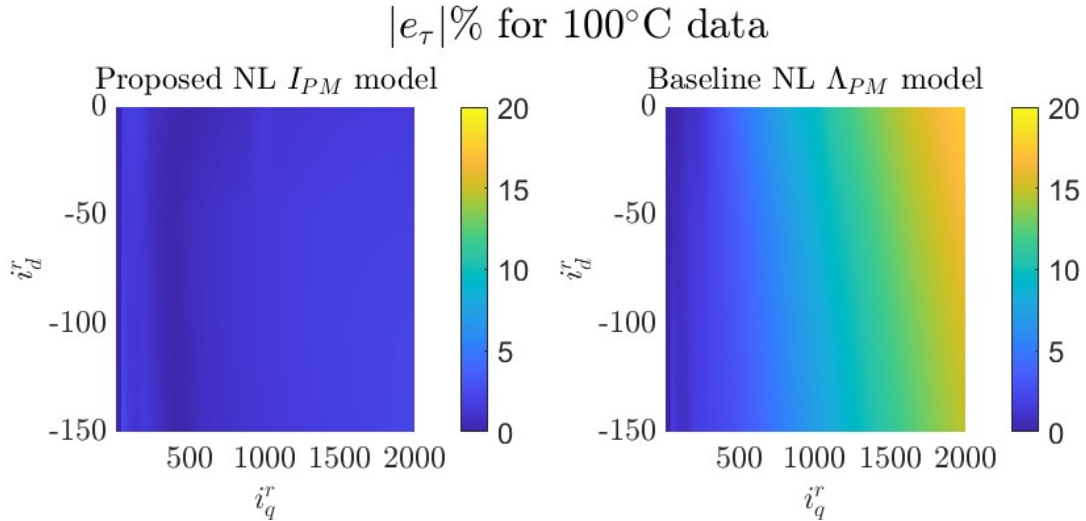


Figure 3.14: Relative torque errors (in %) for Interior Permanent Magnet machine at $T = 100^\circ\text{C}$ for proposed (left) and baseline (right) models, FEA results.

It can be clearly seen that the proposed PM current model shows a significant improvement in torque accuracy when compared to the PM flux linkage model, especially at high values of i_q^r when the machine iron is heavily saturated.

3.4.2.2 Distributed-winding SMPM machine

A 4 pole, distributed winding SMPM (Fig. 3.15a) with machine parameters as shown in Table. 3.2 is simulated in ANSYS FEA for rotor temperatures $T_1 = 20^\circ C$ and $T_2 = 120^\circ C$. The magnetic flux density plot is shown in Fig. 3.15a. The machine is simulated with current density of $13 \text{ A}_{rms}/\text{mm}^2$, which corresponds to forced-air cooling conditions. The NL $\vec{\lambda}^r - \vec{i}^r$ mapping, determined using FEA at temperatures $T_1=20^\circ C$ and $T_2=100^\circ C$, is shown in Fig. 3.16. It can be seen that the proposed PM current model outperforms the PM flux linkage model.

Table 3.2: Distributed winding SMPM parameters

Parameter	Value
Number of poles	4
Number of slots	24
Conductors per slot	100
Parallel branches	1
Magnet material	NdFe35 (N35UH)
Steel type	M19 24G
Stator outer diameter (mm)	160
Stator inner diameter (mm)	80
Rotor outer diameter (mm)	78
Magnet thickness (mm)	5
Machine length (mm)	80

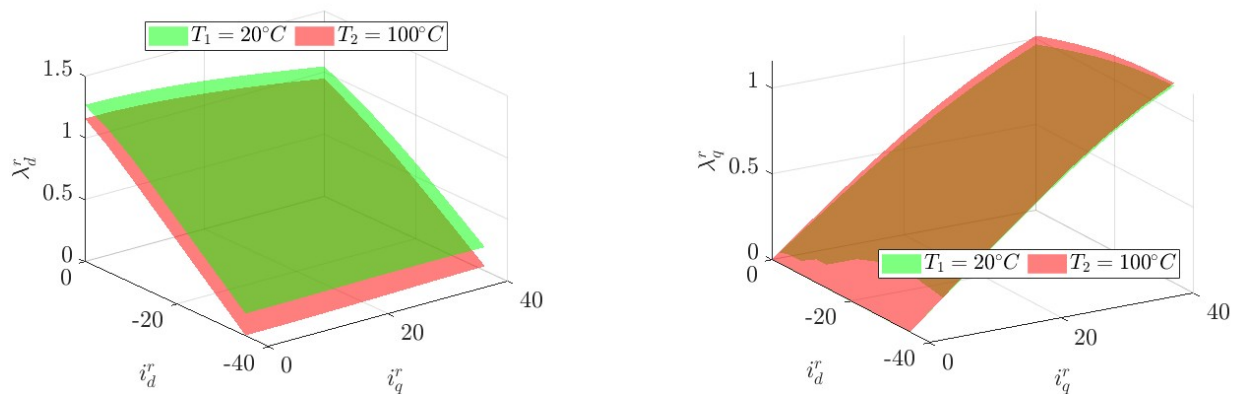
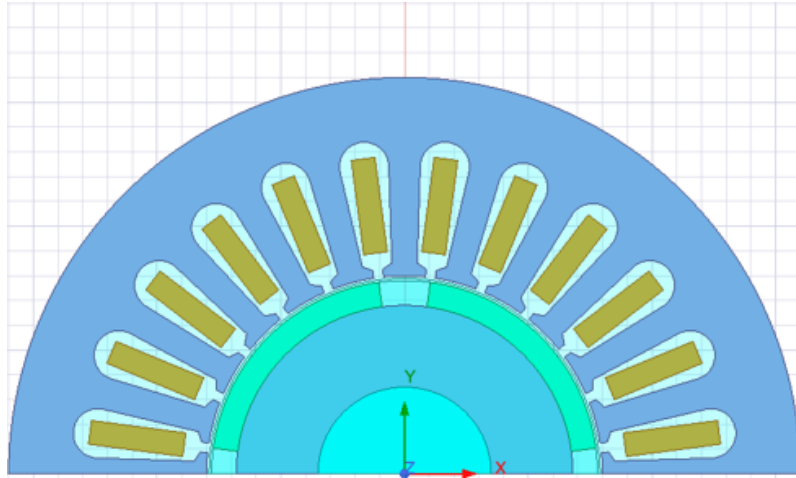
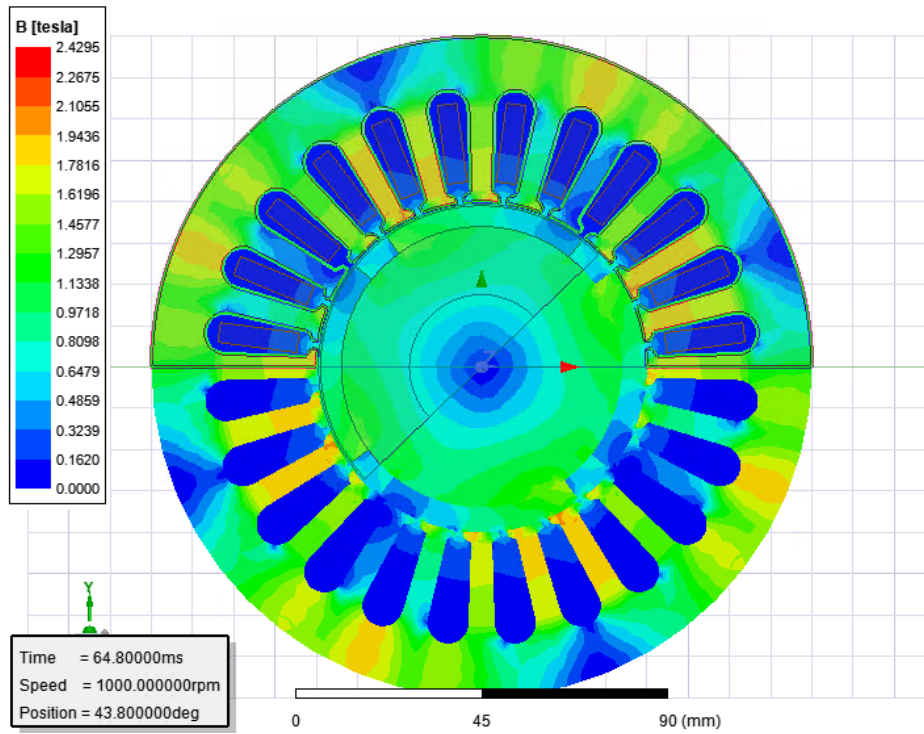


Figure 3.16: $\vec{\lambda}^r - \vec{i}^r$ mapping of distributed winding machine design at different rotor temperatures (FEA results). Top: λ_d^r mapping. Bottom: λ_q^r mapping

Values of I_{pm} and Λ_{pm} for different rotor temperatures are provided in Table. 3.3 for this machine design. Figs 3.17 through 3.19 show absolute errors in λ_d^r and λ_q^r and the relative



(a)



(b)

Figure 3.15: (a) Cross-section of Distributed winding SMPM machine in ANSYS Maxwell, (b) Flux density plot calculated using FEA

Table 3.3: I_{pm} and Λ_{pm} values for Distributed winding SMPM machine

Rotor Temperature ($^{\circ}C$)	I_{pm} (A)	$\Lambda_{pm}(V - s)$
20	40.38	1.2643
60	38.4	1.2281
80	37.3	1.2048
100	36.3	1.1842
120	35.1	1.156

torque error for the two models at temperature T_2 . It can be seen that, for λ_d^r and hence torque, significant errors occur for negative values of i_d^r . This is due to the fact that the machine iron is saturated by the permanent magnets at zero current excitation. The field-weakening effect of negative i_d^r drives the iron out of saturation and so causes the effective PM flux linkage to increase, resulting in the error.

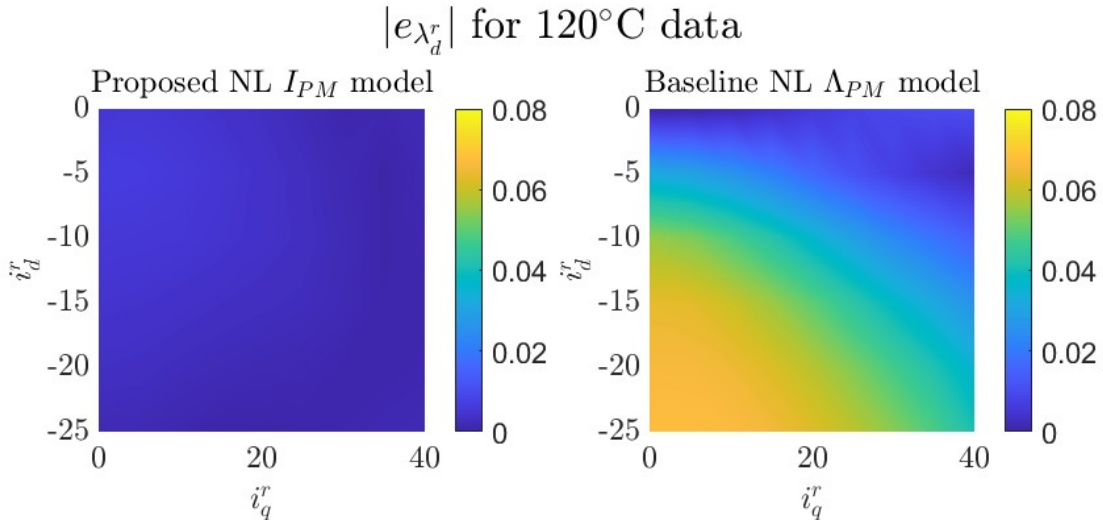


Figure 3.17: Absolute errors in λ_d^r for distributed winding SMPM at $T = 120^{\circ}C$ for proposed (left) and baseline (right) models, FEA results.

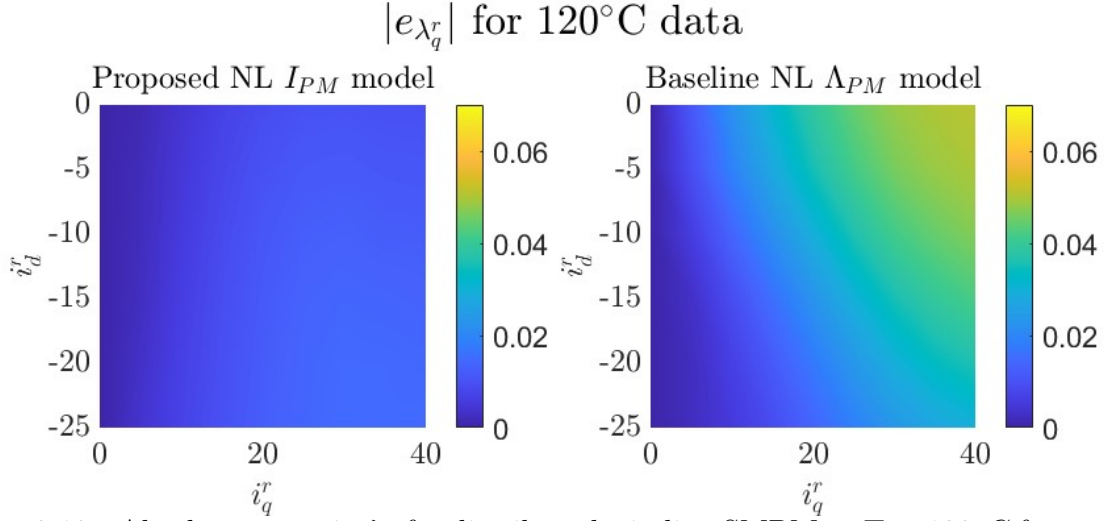


Figure 3.18: Absolute errors in λ_q^r for distributed winding SMPM at $T = 120^\circ C$ for proposed (left) and baseline (right) models, FEA results.

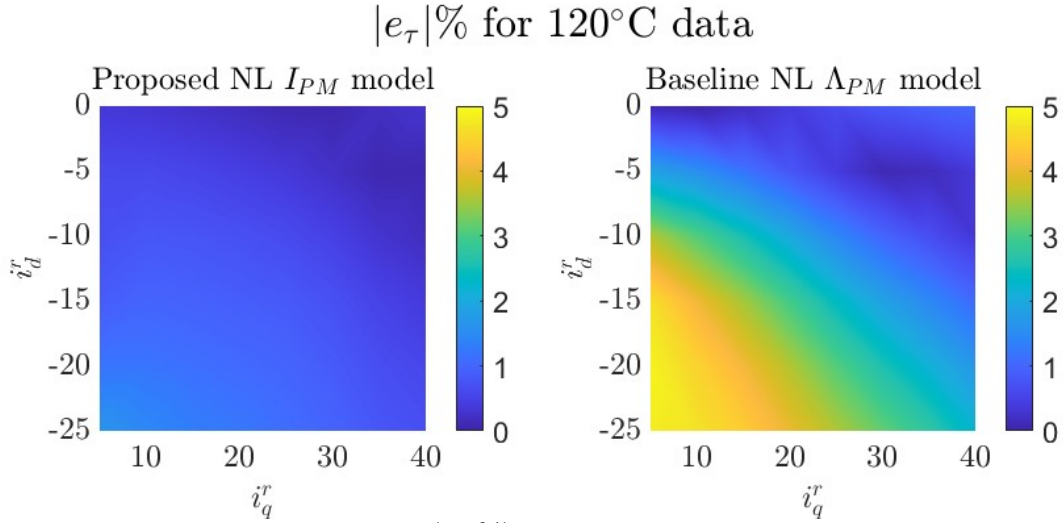


Figure 3.19: Relative torque errors (in %) for distributed winding SMPM at $T = 120^\circ C$ for proposed (left) and baseline (right) models, FEA results.

3.4.2.3 Concentrated-winding SMPM machine

A simple 4-pole concentrated winding SMPM with machine parameters as shown in Table. 3.4 is simulated in ANSYS FEA for rotor temperatures $T_1 = 20^\circ C$ and $T_2 = 100^\circ C$. The machine is simulated with a peak current density of $35.5 \text{ A}_{rms}/\text{mm}^2$, corresponding to liquid cooling conditions. The machine cross section is shown in Fig. 3.20a and the flux density plot for this design is shown in Fig. 3.20b. The NL $\vec{\lambda}^r - \vec{i}^r$ mapping, determined using FEA at temperatures $T_1=20^\circ C$ and $T_2=100^\circ C$, is shown in Fig. 3.21.

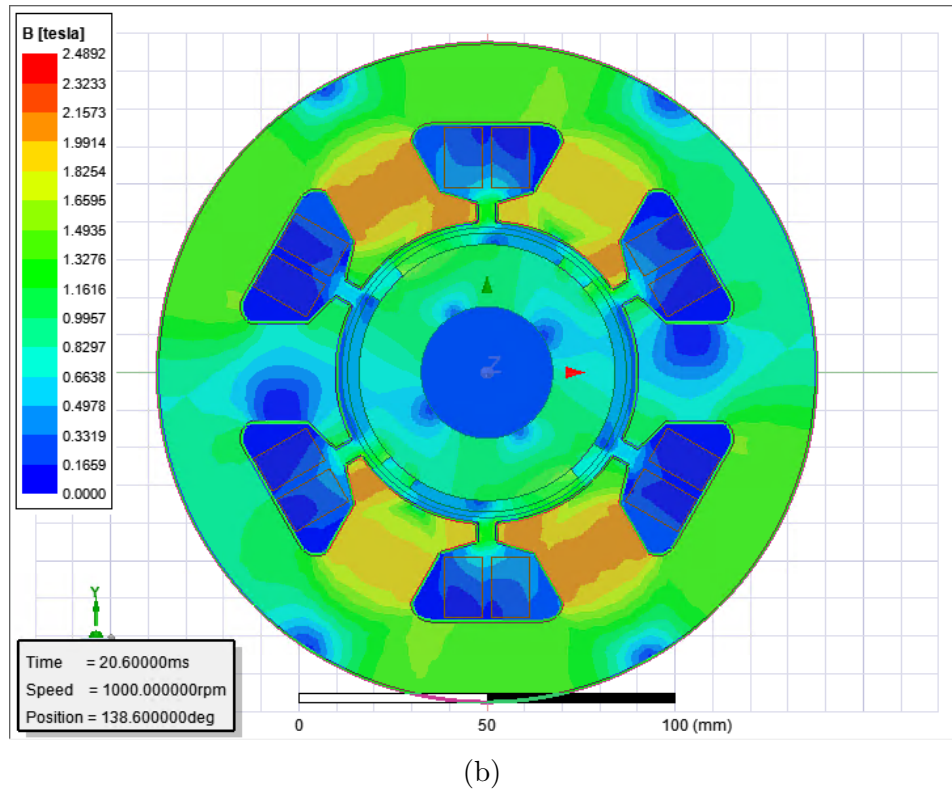
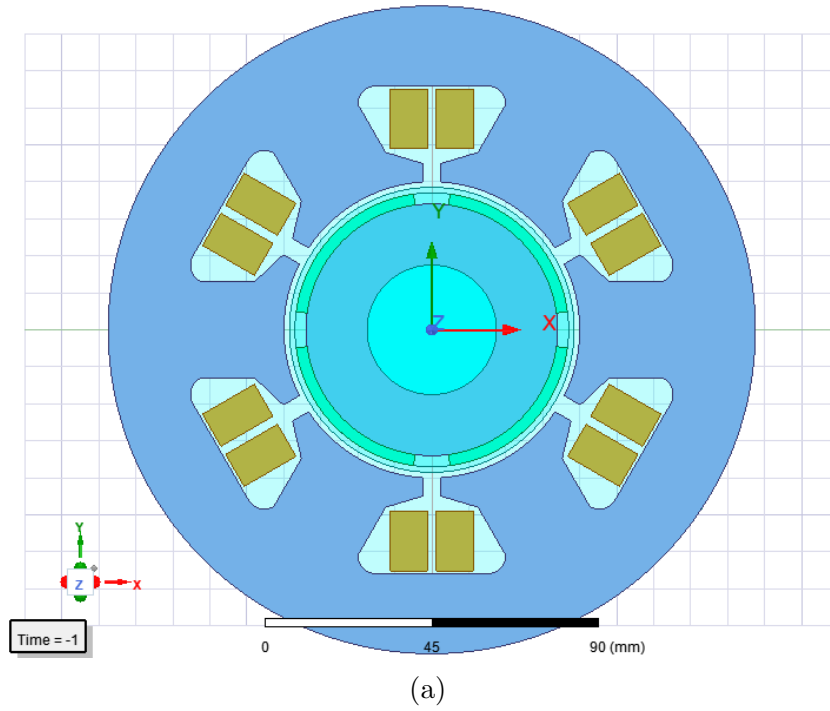


Figure 3.20: (a) Cross-section of concentrated winding SMPM machine in ANSYS Maxwell, (b) Flux density plot calculated using FEA

Table 3.4: Concentrated winding SMPM parameters

Parameter	Value
Number of poles	4
Number of slots	6
Conductors per slot	220
Parallel branches	1
Magnet material	NdFe35 (N35UH)
Steel type	M19 24G
Stator outer diameter (mm)	175
Stator inner diameter (mm)	80
Rotor outer diameter (mm)	74
Magnet thickness (mm)	3
Machine length (mm)	80

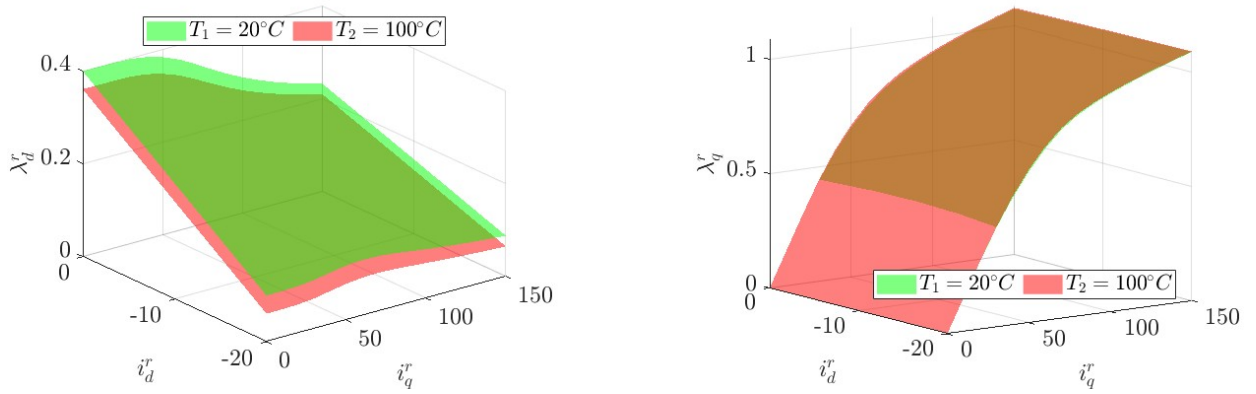


Figure 3.21: $\vec{\lambda}^r$ - $\vec{i}^r$ mapping of concentrated winding machine design at different rotor temperatures (FEA results). Top: λ_d^r mapping. Bottom: λ_q^r mapping

The I_{pm} and Λ_{pm} values for this machine as a function of rotor temperature are provided in Table 3.5. The absolute error in λ_d^r and λ_q^r for both models is compared as shown in

Table 3.5: I_{pm} and Λ_{pm} values for concentrated winding SMPM machine

Rotor Temperature ($^{\circ}C$)	I_{pm} (A)	$\Lambda_{pm}(V - s)$
20	26.6167	0.399
60	25.3747	0.3809
80	24.6345	0.3698
100	24.0188	0.3605

Figs. 3.22-3.23. The relative error in torque by both models is compared in Fig. 3.24. We

can clearly observe the improvement in accuracy in the proposed model from Figs 3.22 and 3.24, especially in the region of high torque where magnetic saturation is most prominent.

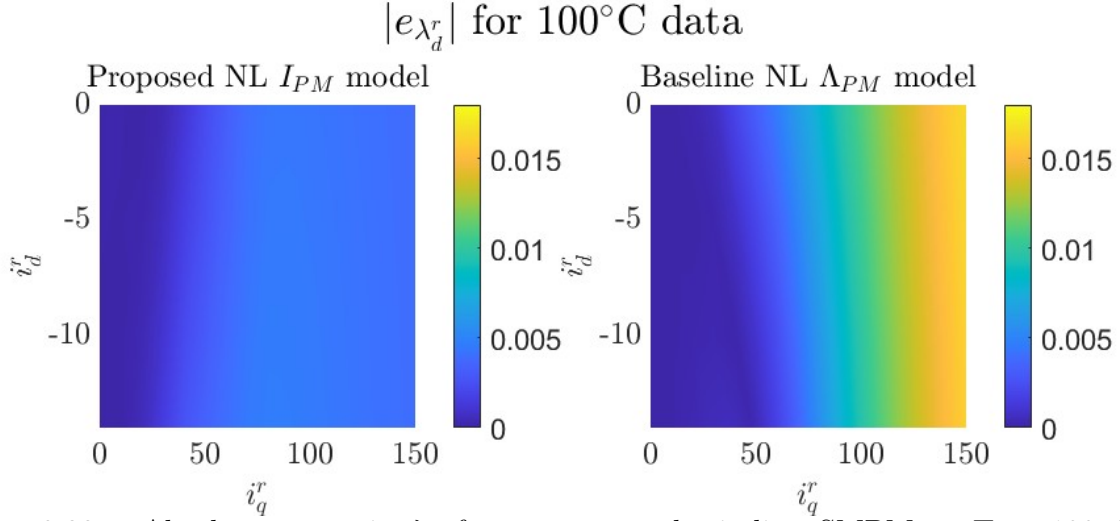


Figure 3.22: Absolute errors in λ_d^r for concentrated winding SMPM at $T = 100^\circ\text{C}$ for proposed (left) and baseline (right) models, FEA results.

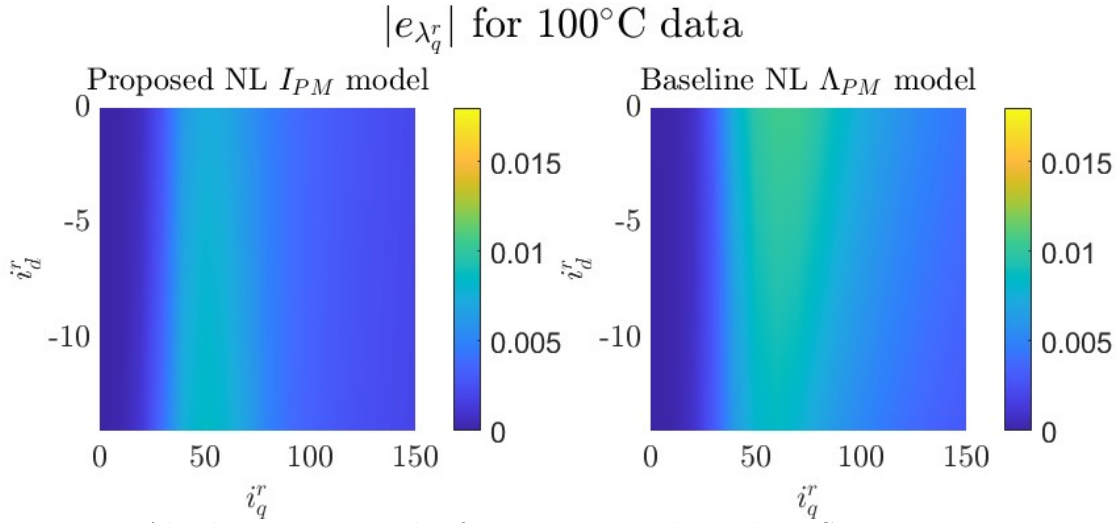


Figure 3.23: Absolute errors in λ_q^r for concentrated winding SMPM at $T = 100^\circ\text{C}$ for proposed (left) and baseline (right) models, FEA results.

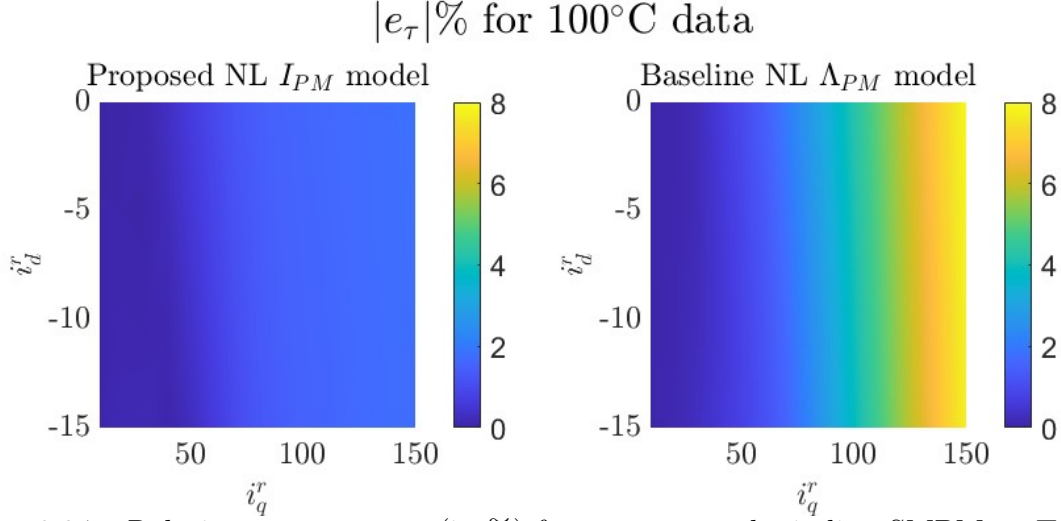


Figure 3.24: Relative torque errors (in %) for concentrated winding SMPM at $T = 100^\circ\text{C}$ for proposed (left) and baseline (right) models, FEA results.

3.4.3 Experimental Validation

3.4.3.1 Setup

A three-phase, eighteen-pole, 145kW SMPM from Unique Mobility (UQM) with specifications as shown in Table 3.6 was characterized using a dSpace Scalexio real-time target controller and a three-phase voltage source inverter consisting of SKiiP 1814 GB17E4-3DUW V2 IGBTs, made by Semikron, is used with a switching frequency of 10 kHz, bus voltage of 175 V, and dead time of 2 μs . Space-Vector Modulation (SVM) is used as the modulation scheme to calculate the desired duty cycles. The rotor speed is kept constant at 500rpm by an electric-machine-based dynamometer. This speed was chosen as the voltage/current relationships of the machine are relatively insensitive to both winding and core losses at the corresponding electrical frequency (75Hz). An infrared temperature sensor from Texense (IRN2) is used to measure the temperature of the magnets on the rotor. This sensor is pointed at the permanent magnets on the rotor of the machine, which effectively form a solid ring in this machine design. While this approach captures a surface temperature of the magnet rather than an internal or average temperature, the purpose of the IR temperature measurement is to establish different rotor temperature conditions in our experiments for the purposes of comparison of the magnetic models, not necessarily to determine an accurate temperature. The measurement, modeling, and estimation of rotor temperature is outside the scope of this manuscript.

Since it is difficult to keep the rotor temperature constant, efforts were made to collect data quickly to keep the temperature variation within 5°C of the stated value. The experimental

setup of the hardware is shown in Fig. 3.25.

Table 3.6: Test machine rating

Description	Value
Number of poles	18
Manufacturer	UQM Technologies
Type	PM Brushless
Maximum speed	8000rpm
Maximum Current	500A
Maximum Voltage	450V
Rated Power	145KW peak/ 85 KW continuous

3.4.4 Characterization Procedure

The goal of the proposed procedure is to characterize the steady-state flux linkages in the rotor reference frame as a function of direct and quadrature currents and rotor temperature. We would like to acknowledge and thank Fanny for sharing her initial work on characterization that is elaborated below.

A characterization process similar to that provided in [2, 34] is used where the flux linkage can be estimated by integrating the electromotive force (EMF) of the windings. Flux linkage can be estimated by integrating the back emf which is the difference of stator voltage in stationary frame and the stator resistive voltage drop, as shown below,

$$\frac{d\vec{\lambda}}{dt} = \vec{e} = \vec{v} - R\vec{i}, \quad (3.32)$$

where R is the stator winding resistance, and $\vec{\lambda} = [\lambda_\alpha \ \lambda_\beta]^T$, $\vec{e} = [e_\alpha \ e_\beta]^T$, $\vec{i} = [i_\alpha \ i_\beta]^T$ and $\vec{v} = [v_\alpha \ v_\beta]^T$ are the equivalent two-phase flux linkage, electromotive force, stator current, and stator voltage vectors in the stationary reference frame, respectively.

The absence of voltage sensor and, moreover, the voltage applied to the stator winding are Pulse Width Modulated (PWM) signals, makes it complex to measure the applied voltage. Control command voltages are used instead, but in order to have an accurate flux linkage estimation, inverter characteristics should also be subtracted along with the stator resistance voltage drop. Inverter characteristics typically consist of transistor voltage drop and dead-time voltage drop. In a half-bridge configuration, the complementary PWM signals to the two transistors in one phase have dead-time where PWM to both transistors is held at zero

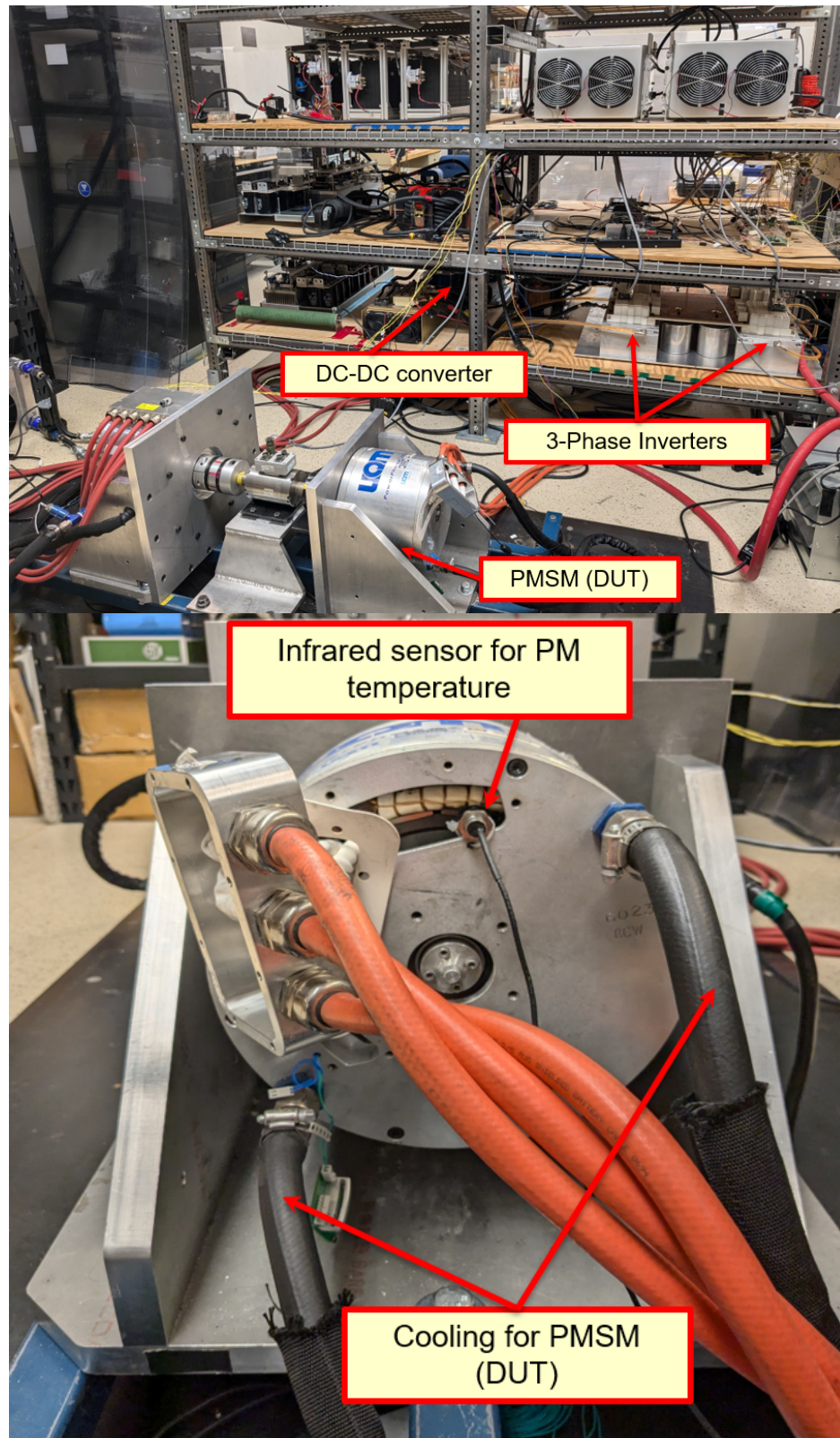


Figure 3.25: Top: Hardware setup for experiment, Bottom: Infrared sensor for rotor temperature measurement on the SMPM test machine

to ensure there is no dead-short circuit. Thus, the flux linkage can be estimated as follows:

$$\vec{\lambda} = \int (\tilde{v} - \vec{v}_{RDT}) dt, \quad (3.33)$$

where $\tilde{v}$ is the command voltage from the controller and

$$\vec{v}_{RDT} = R\vec{i} + \vec{v}_t + \vec{v}_{dead}, \quad (3.34)$$

where $\vec{v}_t(t)$ is the transistor voltage drop and $\vec{v}_{dead}(t)$ is the effective voltage drop due to the dead-time effect in the inverter. In the following we will approximate the RDT voltage with its fundamental component $\vec{v}_{1RDT}$, which can be represented as a nonlinear function of the current vector $\vec{i}$ as follows:

$$\vec{v}_{RDT} \approx \vec{v}_{1RDT} = V_{1RDT} \left(\|\vec{i}\| \right) \frac{\vec{i}}{\|\vec{i}\|}, \quad (3.35)$$

i.e., the magnitude of $\vec{v}_{RDT}$ is a nonlinear function of the current magnitude, and its angle is the same as that of $\vec{i}$. This approximation will not affect the steady-state results.

The integral of the command voltage $\tilde{v}(t)$ can be shown to be

$$\vec{s}(t) = \int \tilde{v}(t) dt = \vec{w}_{RDT}(t) + \vec{\lambda}(t), \quad (3.36)$$

where

$$\vec{w}_{RDT}(t) = \int \vec{v}_{RDT}(t) dt. \quad (3.37)$$

In steady-state operation, this can be represented as follows in the rotor reference frame:

$$\vec{W}_{RDT}^r = \frac{-\mathbf{J}\vec{V}_{RDT}^r}{\Omega_{re}} = -\frac{\mathbf{J}}{\Omega_{re}} V_{RDT}^r(\|\vec{I}\|) \frac{\vec{I}}{\|\vec{I}\|} \quad (3.38)$$

Fig. 3.26 shows a block diagram of the control system used for characterization. The characterization approach assumes that the flux linkage/current relationships of the machine possess the following (standard) symmetries:

$$\lambda_d^r(i_d^r, -i_q^r, T_r) = \lambda_d^r(i_d^r, i_q^r, T_r), \quad (3.39)$$

$$\lambda_q^r(i_d^r, -i_q^r, T_r) = -\lambda_q^r(i_d^r, i_q^r, T_r). \quad (3.40)$$

These symmetries are exploited to separate the steady-state flux linkage $\vec{\lambda}$ from $\vec{W}_{RDT}$. This is achieved by using the current regulator to command the following two steady-state current

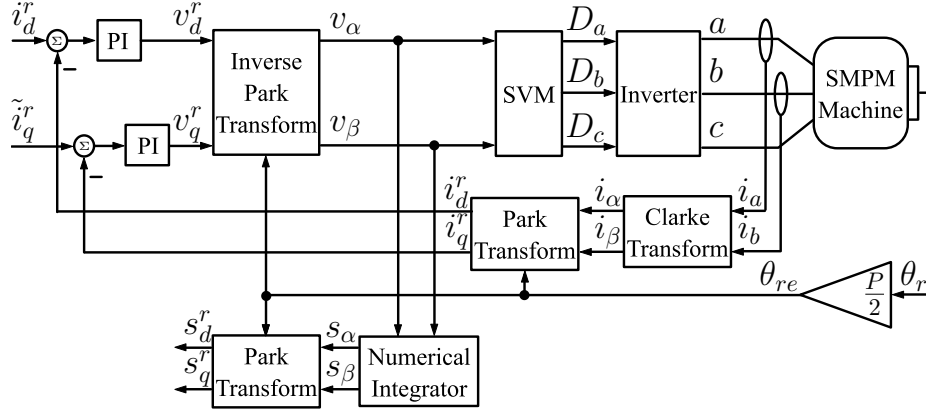


Figure 3.26: Data acquisition and controller block diagram for recording the integrated voltages at different current operating conditions.

vectors, where the direct component is the same and the sign of the quadrature component is flipped.

$$\vec{I}_1^r = \begin{bmatrix} I_d^r \\ I_q^r \end{bmatrix}, \vec{I}_2^r = \begin{bmatrix} I_d^r \\ -I_q^r \end{bmatrix} \quad (3.41)$$

and determining the corresponding integrals $\vec{S}_1^r$ and $\vec{S}_2^r$. These operating points are commanded one immediately after the other so that the rotor temperature remains relatively unchanged.

It can then be shown that, using Eqs. (3.36) and (3.38) and the symmetries in (3.39) and (3.40), the steady-state flux linkages corresponding to I_d^r and I_q^r are given by

$$\Lambda_d^r = \frac{S_{1d}^r + S_{2d}^r}{2}, \Lambda_q^r = \frac{S_{1q}^r - S_{2q}^r}{2} \quad (3.42)$$

3.4.4.1 Numerical Integration

The integration of $\vec{s}$ is done numerically using the following discrete-time transfer function:

$$\vec{S}(z) = \frac{T_s(z-1)}{(z-a)^2} \tilde{\vec{V}}(z), \quad (3.43)$$

where $a = 0.9994$ was chosen so that the transfer function has the same magnitude and phase shift as ideal integration at the frequency of characterization ($75Hz$), as shown in Fig. 3.27. The discrete-time transfer function also has zero DC gain to eliminate the effects of DC offsets on the integration.

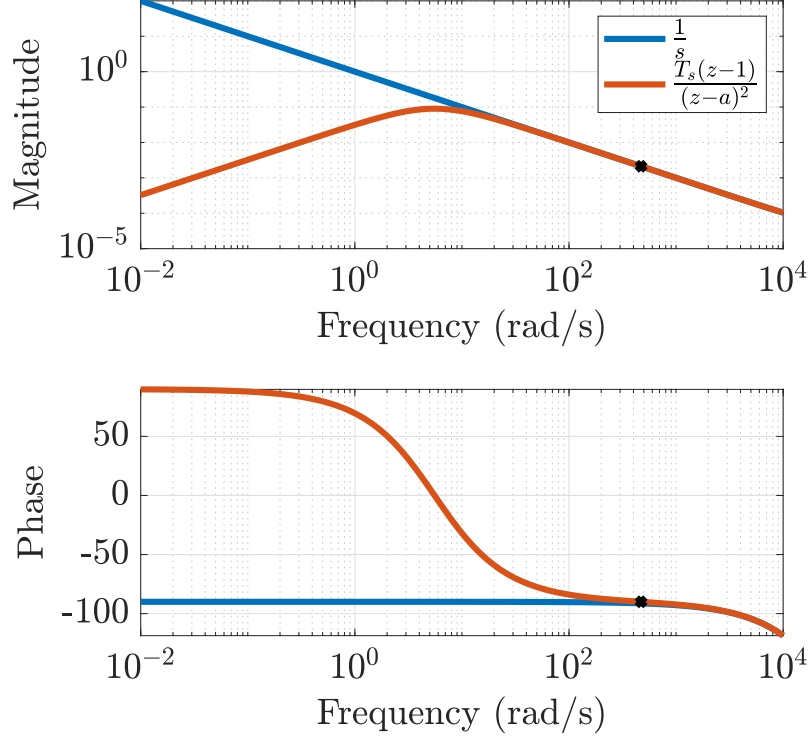


Figure 3.27: Bode plots comparing numerical integrator to ideal integration, $a = 0.9994$. Crosses represent magnitude and phase at the electrical frequency used for characterization (75 Hz)

3.4.4.2 Operating Points

The machine under test does not experience appreciable saturation as shown in Figs.3.29-3.30. Therefore, to evaluate the model over a broad range and illustrate the benefits of the proposed model, the following data set was collected and analyzed:

- Test 1 Data:

Current i_q^r was set to zero and i_d^r was varied from -300A to 500A.

The resulting characterization results for Test 1 is shown in Figs. 3.28 for rotor temperatures $T_1 = 60^\circ C$ and $T_2 = 90^\circ C$.

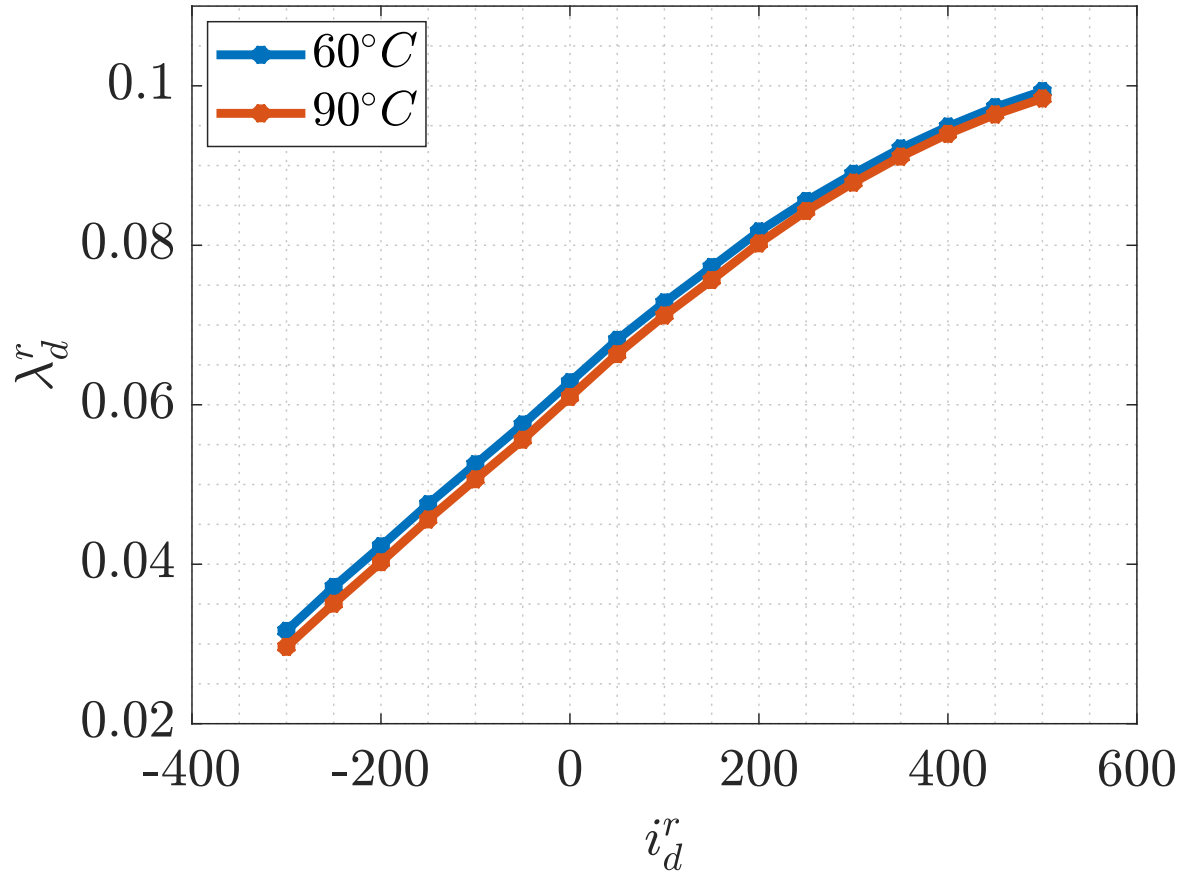


Figure 3.28: Experimental $\lambda_d^r - i_d^r$ characterization of UQM 145 machine for rotor temperatures $T_1 = 60^\circ C$ and $T_2 = 90^\circ C$, $i_q^r = 0$.

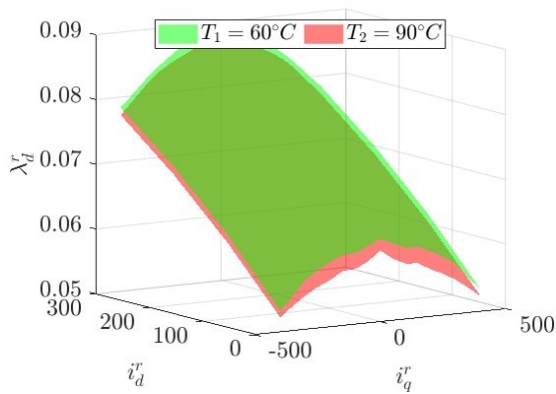


Figure 3.29: Experimental $\lambda_d^r - i_d^r$ characterization of UQM 145 machine for rotor temperatures $T_1 = 60^\circ C$ and $T_2 = 90^\circ C$.

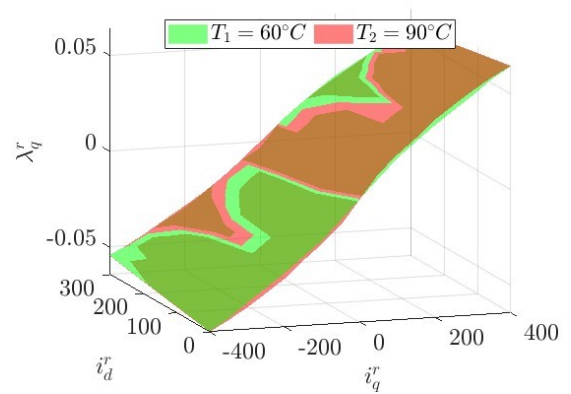


Figure 3.30: Experimental $\lambda_q^r - i_d^r$ characterization of UQM 145 machine for rotor temperatures $T_1 = 60^\circ C$ and $T_2 = 90^\circ C$

3.4.4.3 Calculation of I_{pm} and Λ_{pm}

The flux-linkage estimation process used does not provide accurate flux-linkage estimates when there are high levels of field weakening (i.e., large negative values of i_d^r .) This is due to the fact that, in these circumstances, the emf of the windings becomes a small fraction of the overall voltage. As a result, the estimated flux linkages become highly sensitive to errors in the estimated winding resistance, voltage drops of the inverter transistors, and deadtime, which are used to calculate the emf. Hence we cannot accurately determine the zero-flux-linkage condition as we did with the FEA study, and so another method must be used to determine the value of I_{pm} . This is achieved through the following process:

- Using the experimentally determined $\lambda_d^r - i_d^r$ data at temperature T_1 , create a function which calculates $\lambda_d^r = g_{dT_1} (i_d^r + I_{pmT_1})$ for a given value of I_{pmT_1} through the use of lookup tables and linear interpolation.
- Repeat the above step for data taken at temperature T_2 .
- Determine the values of I_{pm} at temperatures T_1 and T_2 simultaneously by solving the following minimization problem

$$\min_{I_{pmT_1}, I_{pmT_2}} \left\| g_{dT_1} (i_d^r + I_{pmT_1}) - g_{dT_2} (i_d^r + I_{pmT_2}) \right\|^2, \quad (3.44)$$

where $\|\cdot\|$ represents the L^2 norm.

For the baseline NL PM flux linkage model, Λ_{pm} is computed from the experimental through two different methods as described below.

Method 1: This is same method described earlier in 3.4.2 where Λ_{pm} is determined from the estimated λ_d^r when the currents to zero.

Method 2: This method is analogous to the one used to compute I_{pm} . Functions $f_{dT_1} (\lambda_d^r - \Lambda_{pmT_1})$ and $f_{dT_2} (\lambda_d^r - \Lambda_{pmT_2})$ are constructed for given values of Λ_{pmT_1} and Λ_{pmT_2} using the experimental data taken at temperatures T_1 and T_2 , respectively. The values of the PM flux linkage at temperatures T_1 and T_2 are then determined simultaneously by solving the following minimization problem

$$\min_{\Lambda_{pmT_1}, \Lambda_{pmT_2}} \left\| f_{dT_1} (\lambda_d^r - \Lambda_{pmT_1}) - f_{dT_2} (\lambda_d^r - \Lambda_{pmT_2}) \right\|^2. \quad (3.45)$$

The solver fmincon in MATLAB was used to solve the optimization problems to compute I_{pm} and Λ_{pm} . The values of I_{pm} and Λ_{pm} for both rotor temperatures ($T_1 = 60^\circ C$ and $T_2 = 90^\circ C$) from experimental results is shown in Table. 3.7.

Table 3.7: Experimental I_{pm} and Λ_{pm} values

T_r ($^{\circ}C$)	I_{pm} (A)	$\Lambda_{pm}(V-s)$	
		Method 1	Method 2
60	603.6	0.0628	0.0625
90	585.4	0.0610	0.0609

3.4.4.4 Comparison results

The models were compared using the same technique used in 3.4.2 for the FEA results. Fig. 3.31 shows the absolute error in λ_d^r for the PM current model and NL PM flux linkage model for Test 1 data, where the value of Λ_{pm} for the nonlinear PM flux linkage model at temperature T_2 was computed using Method 1. While the models show similar accuracy in the linear region (i.e., negative i_d^r excitation), one can distinctly observe the improved accuracy of the proposed model in the region of magnetic saturation.

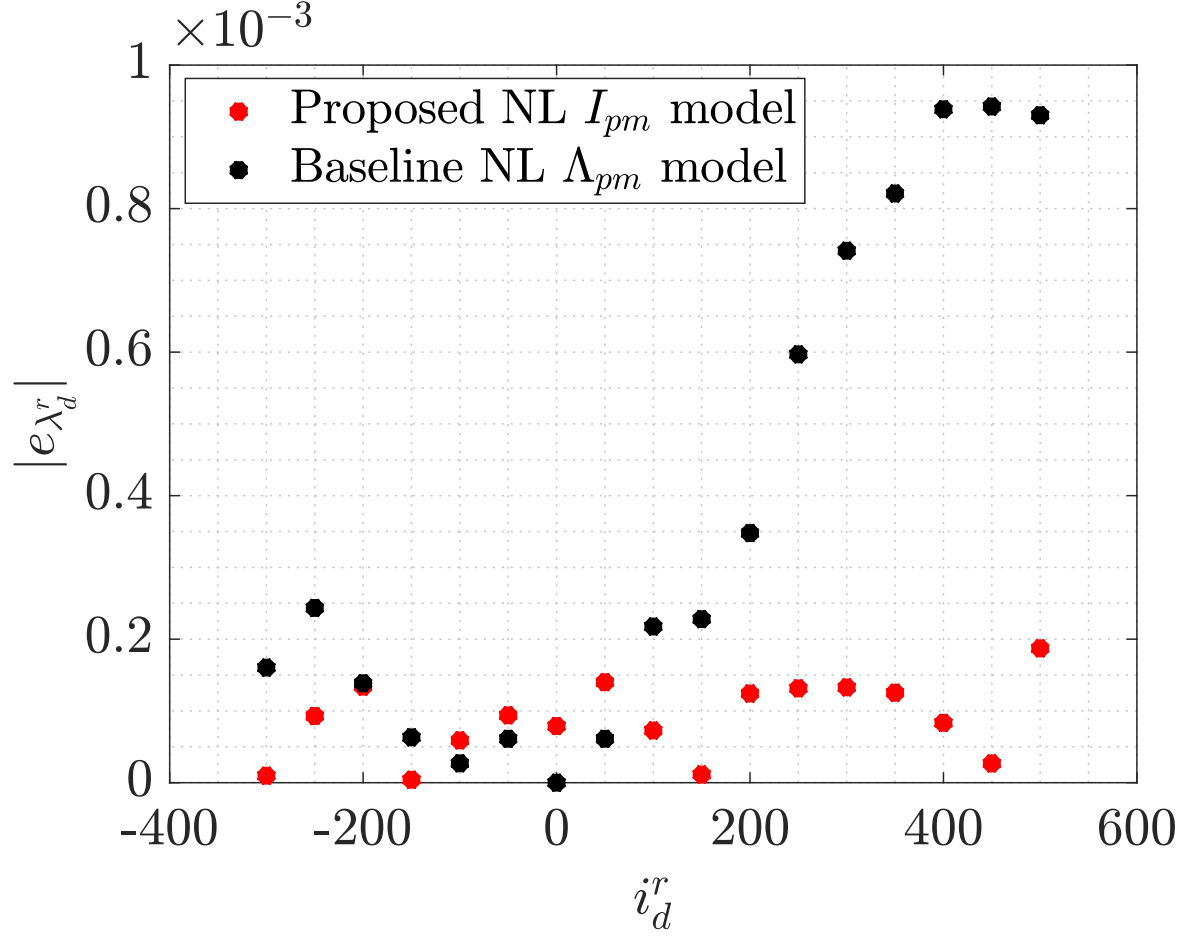


Figure 3.31: Absolute errors in λ_d^r for proposed NL PM current and baseline NL PM flux linkage models at $T_r = 90^\circ\text{C}$ using Method 1 for Λ_{pm} calculation, Test 1 data experimental results.

Fig. 3.32 compares the absolute error in λ_d^r when Λ_{pm} for the NL PM flux linkage model is computed using Method 2 for the Test 1 data. In this case, the proposed model shows better accuracy in both the nonlinear and linear regions.

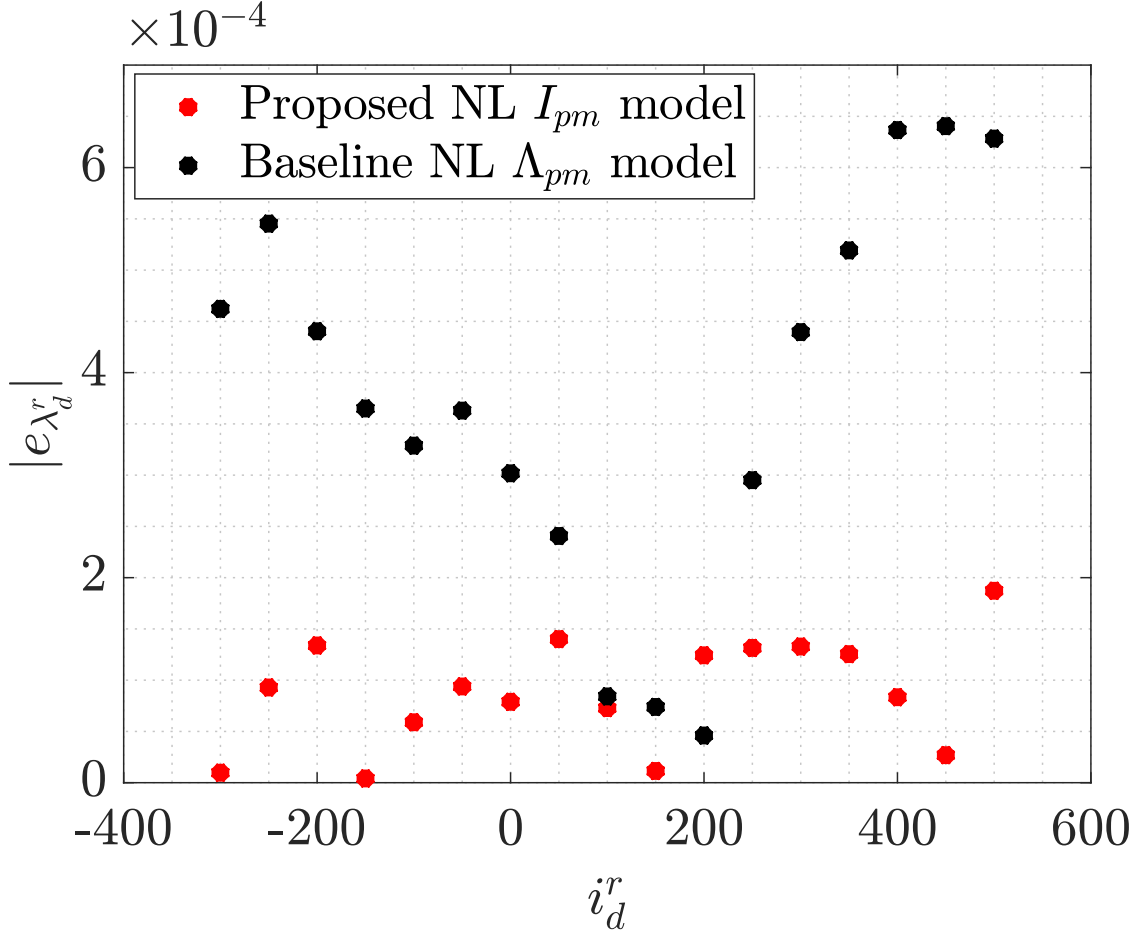


Figure 3.32: Absolute errors in λ_d^r for proposed NL PM current and baseline NL PM flux linkage models at $T_r = 90^\circ\text{C}$ using Method 2 for Λ_{pm} calculation, Test 1 data experimental results.

3.5 Conclusion

This chapter presented a machine model for PMSMs that captures the effects of magnetic saturation and rotor temperature variations, and its efficacy and ease of implementation are demonstrated. The main contribution of the proposed model is the decoupling of the effects of magnetic saturation and rotor temperature variation, the latter being captured in a PM current parameter I_{pm} . This parameter can be adaptively estimated, thereby avoiding the need for rotor temperature measurement. The proposed model is validated using FEA simulation for an interior permanent magnet machine design, a distributed winding SMPM design, and a concentrated winding SMPM design, and shows high accuracy at different rotor temperatures when compared with a nonlinear PM flux linkage model. Experimental results also confirm that the proposed model is more accurate than the nonlinear PM flux

linkage model.

CHAPTER 4

Adaptive Parameter Estimation Strategy for Permanent Magnet excitation (I_{pm})

4.1 Introduction

As described in the previous chapter, utilizing the decoupling of magnetic saturation from rotor temperature variation, the proposed model in summary consists of:

- $\vec{h}(\vec{\lambda}^r)$: NL flux linkage/current mapping that captures magnetic saturation and is relatively insensitive to temperature variation,
- $I_{pm}(T_r)$: PM current parameter that varies with respect to the rotor temperature

Since the proposed model captures the rotor temperature variation in a single parameter, by developing an accurate adaptive parameter estimation strategy for the PM current, we can eliminate the need for a rotor temperature sensor. An accurate estimator of the PM current parameter will therefore:

- Capture the effects of rotor temperature variation and magnet degradation for health/condition monitoring purposes,
- Achieve accurate torque regulation, and
- Allow estimation of rotor temperature for use in thermal/loss modeling.

4.2 Proposed Estimation Strategy

The proposed estimation strategy assumes that we can accurately estimate flux linkage using the measured current and command voltage, which hence will be referred to as the “actual” flux linkage observer. A Luenberger observer with output current injection that estimates

flux linkage considering uncertainty in I_{pm} parameter is shown in 4.2, where “ $\hat{\cdot}$ ” denotes an estimated value and R_o is the observer output injection gain. The idea is to use the difference in the flux linkage from the “Actual” and Luenberger observers to drive the I_{pm} estimator. The proposed I_{pm} estimator dynamic is highlighted in eq.(4.15). We know,

$$\vec{\lambda}^r = \vec{v}^r - R\vec{i}^r + \omega_{re}\mathbf{J}\vec{\lambda}^r, \quad (4.1)$$

$$\vec{\hat{\lambda}}^r = \vec{v}^r - R\vec{\hat{i}}^r + \omega_{re}\mathbf{J}\vec{\hat{\lambda}}^r + R_o \left(\vec{i}^r - \vec{\hat{i}}^r \right), \quad (4.2)$$

where eq. (4.1) is the actual flux linkage dynamics and eq. (4.2) is the Luenberger observer with output injection where,

$$\vec{i}^r = \left[\vec{h} \left(\vec{\lambda}^r \right) - \vec{i}_{pm}^r \right] \quad (4.3)$$

$$\vec{\hat{i}}^r = \left[\vec{h} \left(\vec{\hat{\lambda}}^r \right) - \vec{\hat{i}}_{pm}^r \right] \quad (4.4)$$

The flux linkage error dynamics can then be shown to be

$$\vec{e}_{\lambda^r} = -(R + R_o)\vec{e}_h + \omega_{re}\mathbf{J}\vec{e}_{\lambda^r} + R_o\vec{e}_{PM}, \quad (4.5)$$

where

$$\begin{aligned} \vec{e}_{\lambda^r} = \begin{bmatrix} e_{\lambda_d^r} \\ e_{\lambda_q^r} \end{bmatrix} &= \begin{bmatrix} \lambda_d^r - \hat{\lambda}_d^r \\ \lambda_q^r - \hat{\lambda}_q^r \end{bmatrix}, \vec{e}_h = \begin{bmatrix} e_{h_d} \\ e_{h_q} \end{bmatrix} = \begin{bmatrix} h_d \left(\vec{\lambda}^r \right) - h_d \left(\vec{\hat{\lambda}}^r \right) \\ h_q \left(\vec{\lambda}^r \right) - h_q \left(\vec{\hat{\lambda}}^r \right) \end{bmatrix}, \\ \vec{e}_{PM} = \begin{bmatrix} e_{I_{pm}} \\ 0 \end{bmatrix} &= \begin{bmatrix} I_{pm} - \hat{I}_{pm} \\ 0 \end{bmatrix} \end{aligned} \quad (4.6)$$

As shown in [47], in order to design a stable adaptive estimator with guaranteed convergence, a Lyapunov function is first formulated and then an adaptive law is chosen so that $\dot{V} \leq 0$ to ensure stability. We choose the following Lyapunov function,

$$V = \frac{1}{2} \|\vec{e}_{\lambda^r}\|^2 + \frac{1}{2} \frac{\|\vec{e}_{PM}\|^2}{K_1} \quad (4.7)$$

Taking the derivative of V yields,

$$\dot{V} = \dot{e}_{\lambda_d^r} e_{\lambda_d^r} + \dot{e}_{\lambda_q^r} e_{\lambda_q^r} + \frac{1}{K_1} \dot{e}_{I_{pm}} e_{I_{pm}} \quad (4.8)$$

Substituting from 4.5, we get

$$\dot{V} = -(R + R_o) e_{h_d} e_{\lambda_d^r} - (R + R_o) e_{h_q} e_{\lambda_q^r} + R_o e_{I_{pm}} e_{\lambda_d^r} + \frac{1}{K_1} \dot{e}_{I_{pm}} e_{I_{pm}} \quad (4.9)$$

$$(4.10)$$

Using the Mean Value Theorem we know that, given any $\vec{e}_{\lambda^r}$, there exists $\vec{\lambda}^r$ such that

$$\vec{e}_{\lambda^r}^T [\vec{h}(\vec{\lambda}^r) - \vec{h}(\vec{\lambda})] = \vec{e}_{\lambda^r}^T \frac{d\vec{h}(\vec{\lambda}^r)}{d\vec{\lambda}^r} (\vec{\lambda}^r - \vec{\lambda}) = \vec{e}_{\lambda^r}^T \frac{d\vec{h}(\vec{\lambda}^r)}{d\vec{\lambda}^r} \vec{e}_{\lambda^r} \geq \alpha \|\vec{e}_{\lambda^r}\|^2 \quad (4.11)$$

Thus to ensure $\dot{V} \leq 0$, we obtain the following adaptive law

$$\dot{e}_{I_{pm}} = -K_1 R_o e_{\lambda_d^r} = -K_a e_{\lambda_d^r} \quad (4.12)$$

Hence, by the choosing the above adaptive law, we obtain

$$\dot{V} \leq -\alpha (R + R_o) \|\vec{e}_{\lambda^r}\|^2 \quad (4.13)$$

i.e., the estimator with 4.12 as the adaptive law is stable in the sense of Lyapunov.

As I_{pm} changes with rotor temperature and thermal time scales are relatively long, we can assume that the actual permanent magnet current changes very slowly relative to the observer dynamics, and be considered effectively constant i.e.,

$$\frac{dI_{pm}}{dt} \approx 0 \quad (4.14)$$

Thus, the dynamics of the estimated permanent magnet current can be shown as,

$$\frac{d\hat{I}_{pm}}{dt} = K_a e_{\lambda_d^r} \quad (4.15)$$

where K_a is a control gain that determines the convergence rate of the parameter. Fig. 4.1 shows the block diagram of the proposed I_{pm} estimator.

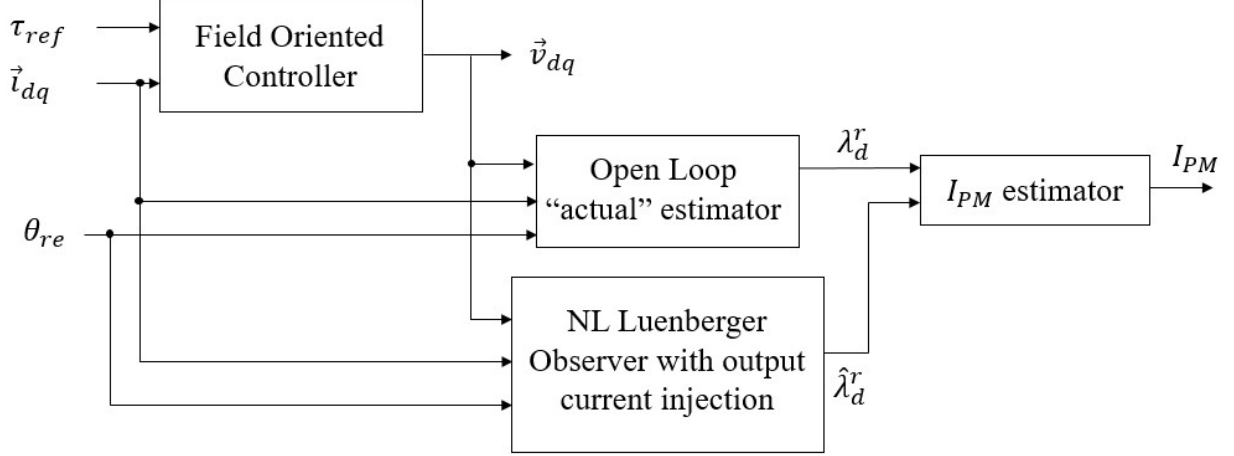


Figure 4.1: I_{pm} estimator block diagram

4.3 Open loop (“Actual”) Flux Linkage Observer

In practice we cannot measure the actual flux linkage, an open loop observer using the measured current and command voltage, which hence will be referred to as the “actual” flux linkage observer, is shown in Eq.4.16.

$$\vec{\lambda}^r = \vec{v}^r - R\vec{i}^r + \omega_{re}\mathbf{J}\vec{\lambda}^r \quad (4.16)$$

Since we use measured currents and command voltages, noise in the measured currents can cause stability issues. Further, a numerical integrator similar to the type described in section 3.4.4.1 is used here as the open loop estimator is implemented in the stationary reference frame. This numerical integrator isn’t accurate at low speeds and so the I_{pm} estimator should only be used at relatively high speeds. Another important thing to note is the accurate compensation of resistive drop is required in order to obtain a precise flux linkage estimation at low motor speeds. From eq.(4.16) we can observe that as the rotor speed increases, the back emf becomes dominant over the resistive voltage drop thereby making the open loop estimator less sensitive to uncertainty in resistance parameter. At high speed operation, the resistive voltage drop contribution becomes almost negligible, making it insensitive to variation in resistance parameter. Thus, we choose an operating speed that is high enough where back emf is dominant.

4.4 Validation of the proposed estimator

4.4.1 Simulation results

The performance of the proposed estimator is validated on MATLAB simulink. A PMSM machine model is created using the $\lambda - i$ relationship obtained from the experimental characterization of UQM machine as shown in chapter 3, section 3.4.4.2. A current step from -100A to -300A is given as shown in Fig. 4.2 for a PM temperature of $50^\circ C$. The initial value of I_{pm} is set to an incorrect value (500A). We can clearly observe from Fig. 4.3 that the proposed I_{pm} estimator converges to its actual value.

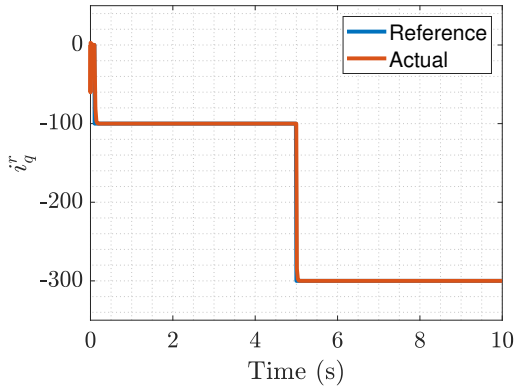


Figure 4.2: i_q^r current profile.

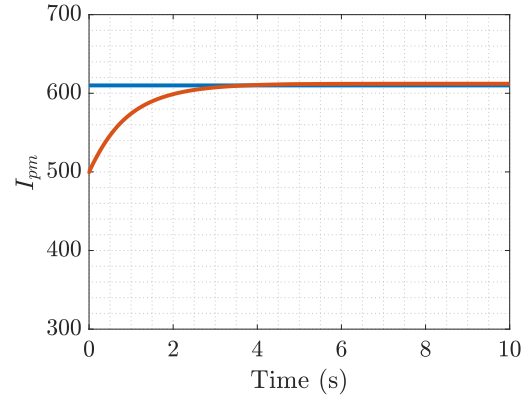


Figure 4.3: Estimated and actual I_{pm} parameter

4.4.2 Experimental results

4.4.2.1 I_{pm} estimator performance at low speed (500rpm)

The above I_{pm} estimator was tested on experimental setup described in section 3.4.3.1 in the Advanced Electric Powertrain Testbed in Michigan Power and Energy Lab. To test the performance of the estimator at low speeds, an operating point with speed of 500rpm, V_{bus} of 175V and varying i_q^r command was analyzed. Figs. 4.4-4.5 show the speed of operation and the varying i_q^r commanded over the complete test cycle respectively. The change in the PM temperature during the test cycle is shown in Fig. 4.6. The performance of the estimator can be seen in Figs. 4.7-4.8 that plots the estimated I_{pm} and "actual" I_{pm} and the relative error in the estimation. The "actual" I_{pm} is a linear extrapolation of the I_{pm} for different temperatures using I_{pm} values computed from the experimental characterization data of the machine at two different temperatures ($60^\circ C$ and $90^\circ C$) as explained in detail in section 3.4.4.3. We observe that, even though the estimator converges to a steady state value, the

estimator diverts from its value for each change in i_q^r current command thus resulting in a relative error as high as 3.5%.

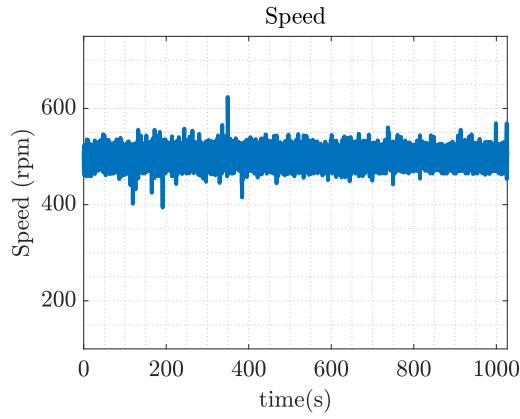


Figure 4.4: Rotor speed.

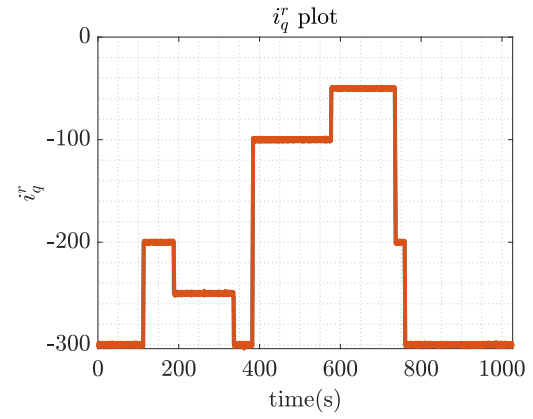


Figure 4.5: i_q^r current profile.

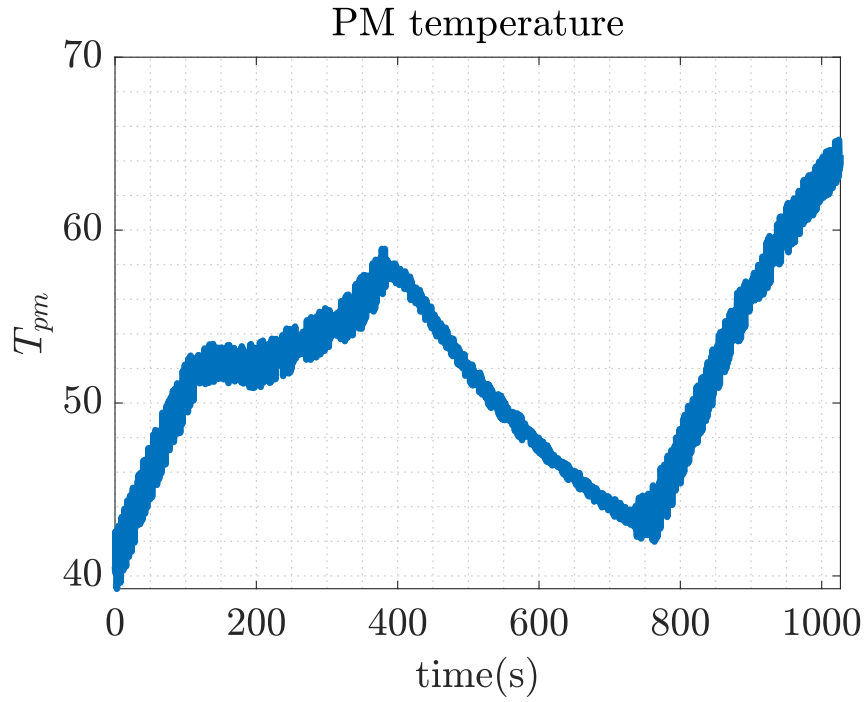


Figure 4.6: Measured PM temperature.

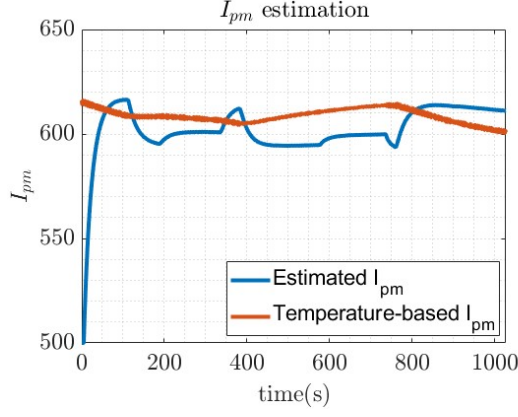


Figure 4.7: Estimated and temperature-based I_{pm} parameter.

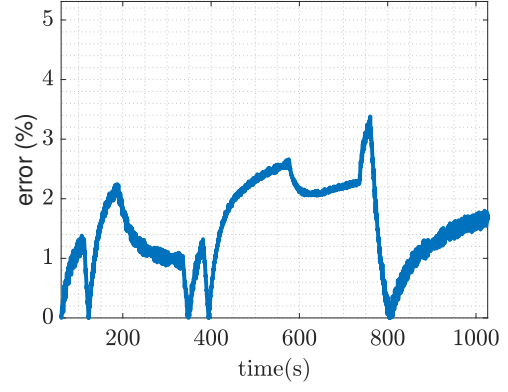


Figure 4.8: Relative error in I_{pm} estimation.

Upon looking more closely at the command voltage and the voltage drop due to resistance at this operating point, i.e. 500rpm and $i_q^r = -300\text{A}$ (Figs. 4.9-4.11), it can be observed that V_{RDT} is 14V and $\|V\|$ is 49V, i.e. the V_{RDT} is roughly 29% of the commanded voltage, so inaccurate compensation of the voltage drop has significant impact on the performance of the estimator as seen above.

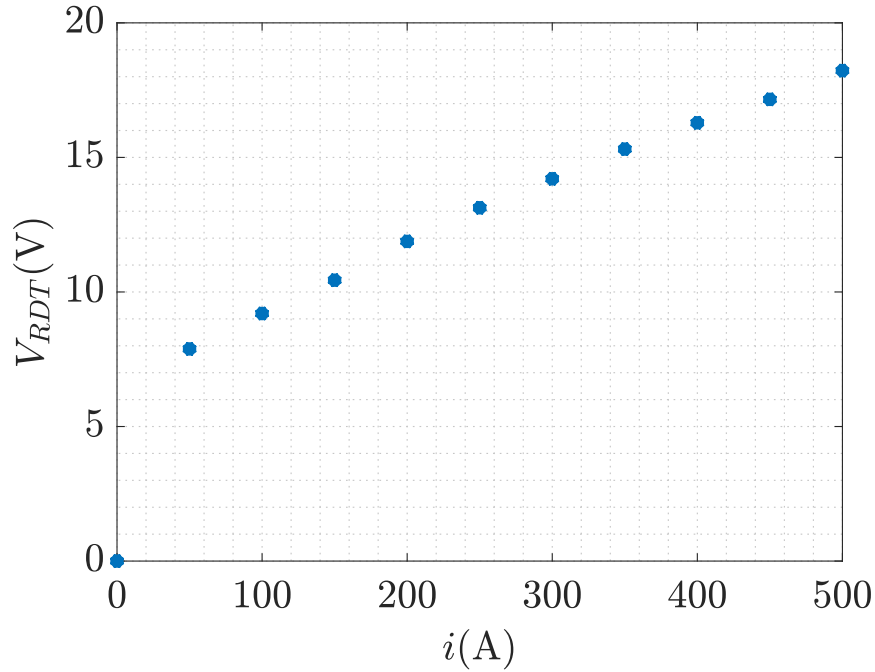


Figure 4.9: V_{RDT} at 175V V_{bus} , 500rpm and $T_r=60^\circ\text{C}$.

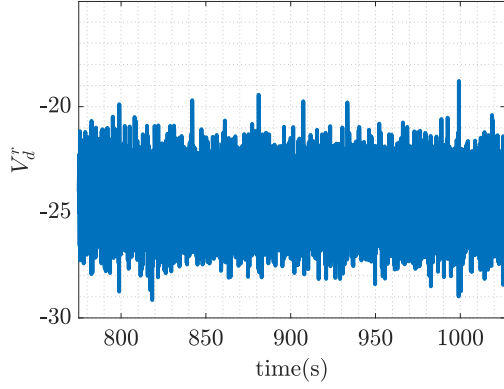


Figure 4.10: V_d^r for operating at 500rpm and -300A i_q^r .

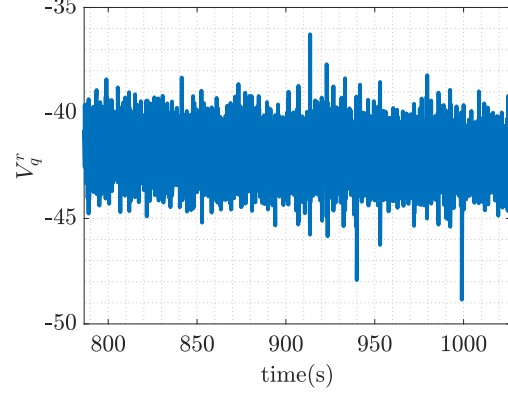


Figure 4.11: V_q^r for operating at 500rpm and -300A i_q^r .

4.4.2.2 I_{pm} estimator performance at 1500rpm and $T_r \approx 50^\circ\text{C}$

The estimator performance at higher speed (1500rpm) can be observed in Figs. 4.13-4.15 for a varying current command (Fig. 4.12). We can see that we are able to estimate the I_{pm} parameter within 0.4% steady state error when compared to the “actual” I_{pm} that is a function of rotor temperature. The initial condition for I_{pm} is again set at 500A.

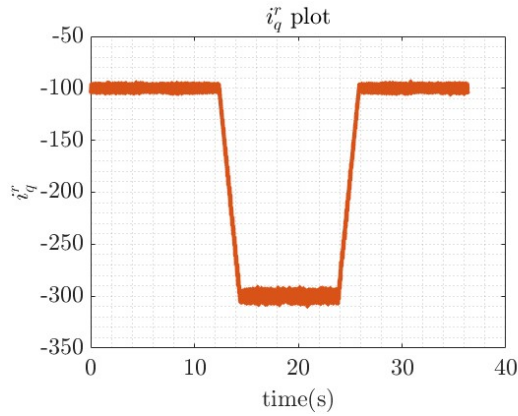


Figure 4.12: i_q^r current profile.

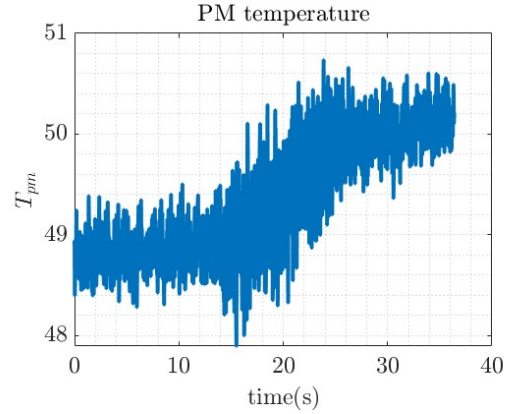


Figure 4.13: Measured PM temperature (T_r)

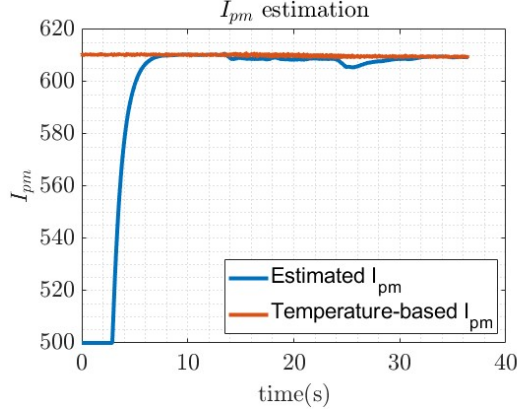


Figure 4.14: Estimated and temperature-based I_{pm} parameter

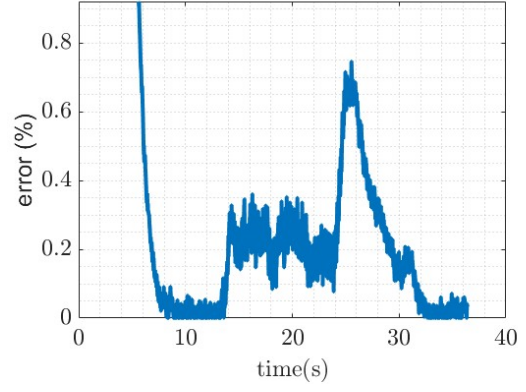


Figure 4.15: Relative error in I_{pm} estimation

4.4.2.3 I_{pm} estimator performance at 1500rpm and $T_r \leq 50^\circ\text{C}$

The estimator performance at 1500rpm and at $T_r = 40^\circ\text{C}$ (i.e. lower than 50°C) can be observed in Figs. 4.17-4.15 for a varying current command (Fig. 4.16) where the resistive voltage drop obtained from machine characterization at PM temperature 50°C is used in the flux estimators. We can see that we are able to estimate the I_{pm} parameter with about 0.6% steady state error when compared to the “actual” I_{pm} that is a function of rotor temperature. The initial condition for I_{pm} is again set at 500A.

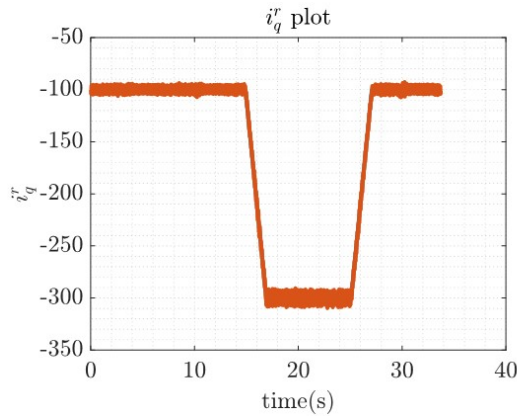


Figure 4.16: i_q^r current profile.

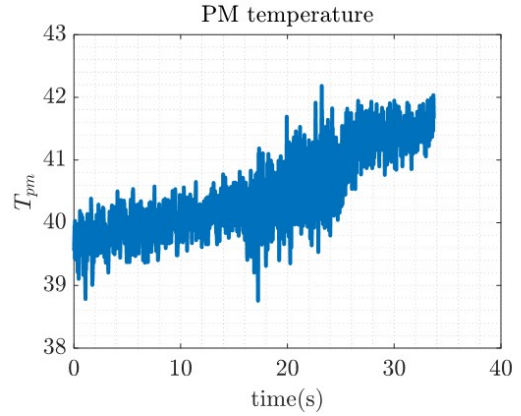


Figure 4.17: Measured PM temperature (T_r)

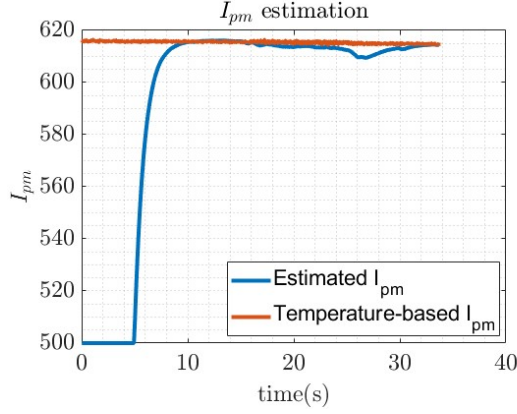


Figure 4.18: Estimated and temperature-based I_{pm} parameter

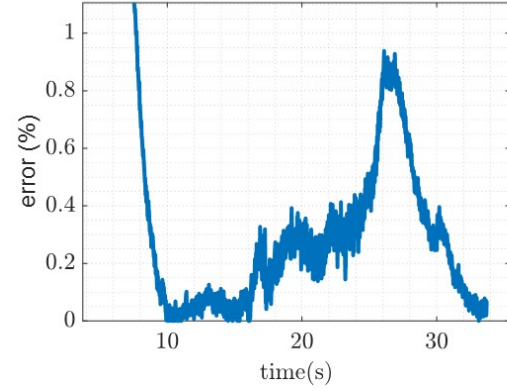


Figure 4.19: Relative error in I_{pm} estimation

4.4.2.4 I_{pm} estimator performance at 1500rpm and $T_r \geq 50^\circ\text{C}$

The estimator performance at 1500rpm and at T_r higher than 50°C can be observed in Figs. 4.21-4.15 for a varying current command (Fig. 4.20) where the resistive voltage drop obtained from machine characterization at PM temperature 50°C is used in the flux estimators. We can see that we are able to estimate the I_{pm} parameter with a slight increased error, i.e. about 0.8% steady state error when compared to the “actual” I_{pm} that is a function of rotor temperature. The initial condition for I_{pm} set at 500A.

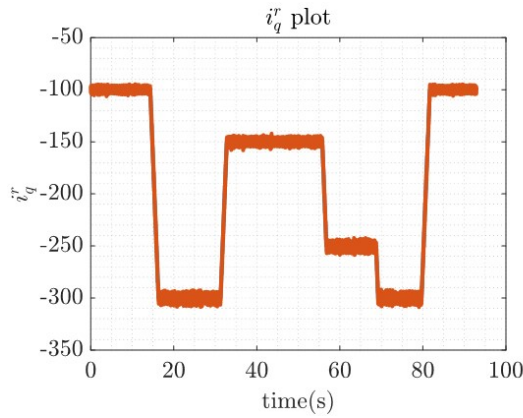


Figure 4.20: i_q^r current profile.

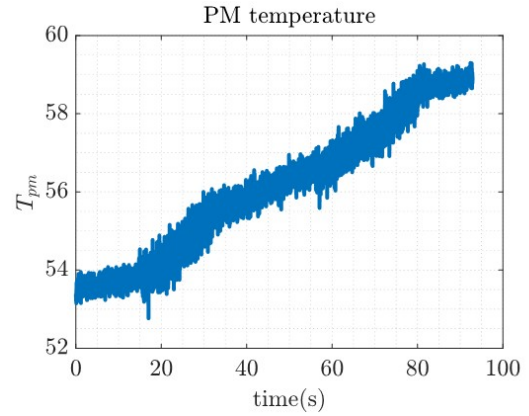


Figure 4.21: Measured PM temperature (T_r)

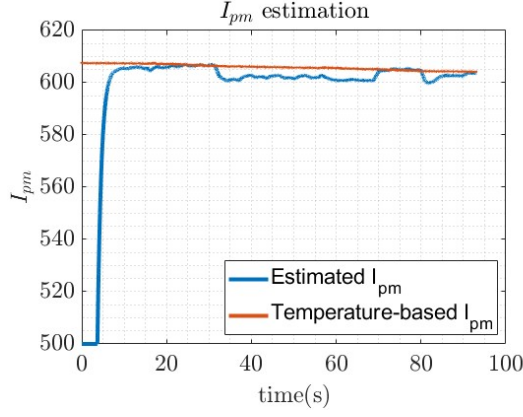


Figure 4.22: Estimated and temperature-based I_{pm} parameter

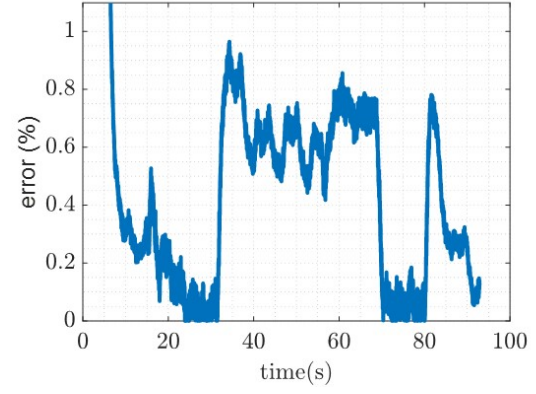


Figure 4.23: Relative error in I_{pm} estimation

4.4.2.5 I_{pm} estimator performance at 1500rpm and over wider temperature range

For test conditions where the PM temperature is varying over a wider range during the test cycle, the performance of the proposed estimator is shown in Figs. 4.24-4.27. From Figs. 4.28-4.30 we can clearly see that at the chosen operating point (i.e., rotor speed = 1500rpm), the resistive drop constitutes roughly 20% of the total applied voltage and therefore is not negligible. Since we have inaccurate compensation of the resistive voltage drop, we observe steady state error in I_{pm} estimation to be slightly increased but less than 1%.

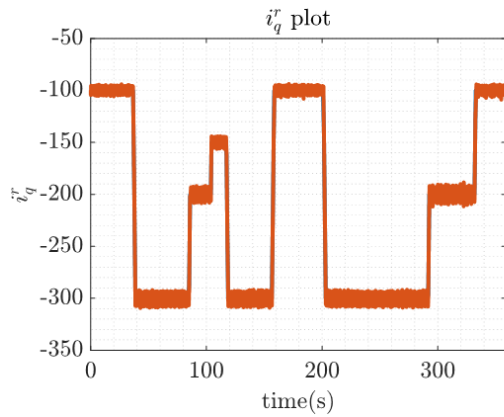


Figure 4.24: i_q^r current profile.

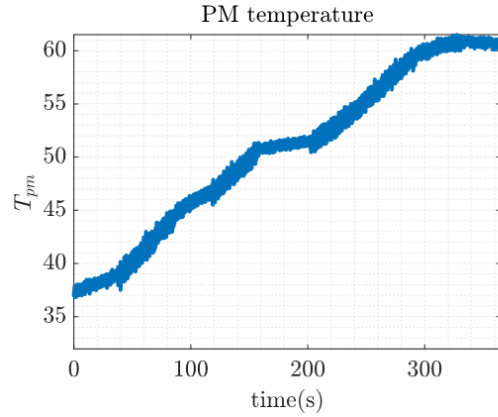


Figure 4.25: Measured PM temperature (T_r)

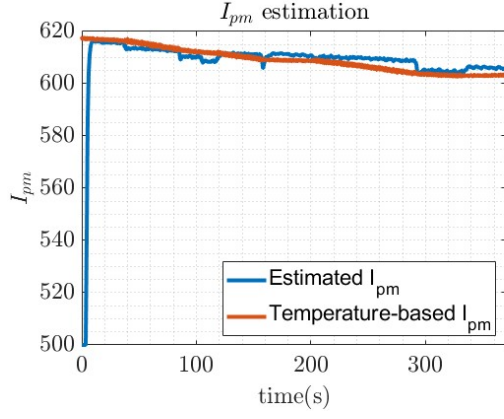


Figure 4.26: Estimated and temperature-based I_{pm} parameter over wide rotor temperature range at 1500rpm

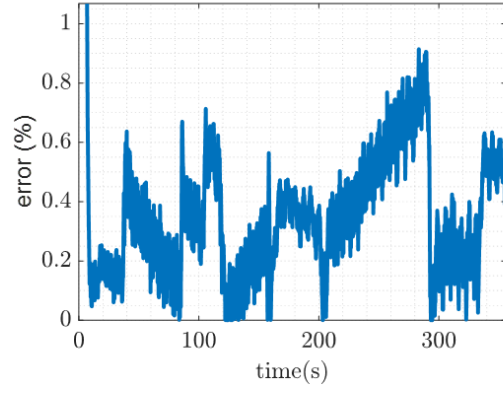


Figure 4.27: Relative error in I_{pm} estimation over wide rotor temperature range at 1500rpm

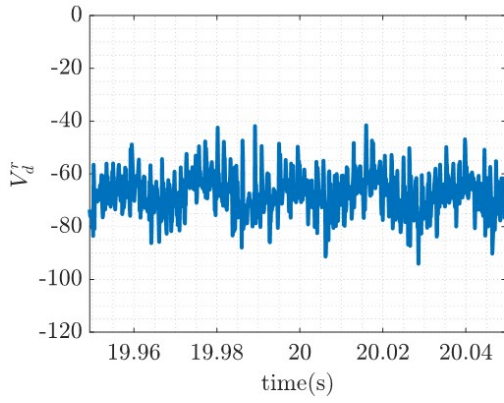


Figure 4.28: V_d^r when operating at 1500rpm and $i_q^r = -300A$.

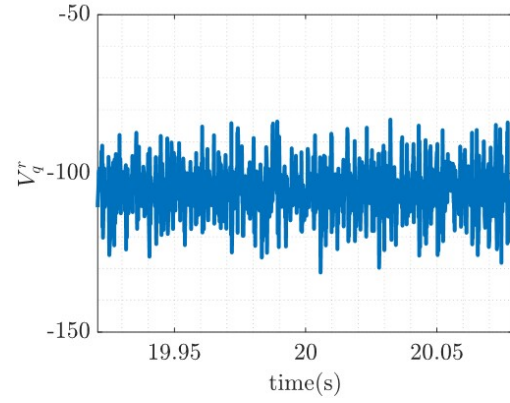


Figure 4.29: V_q^r when operating at 1500rpm and $i_q^r = -300A$.

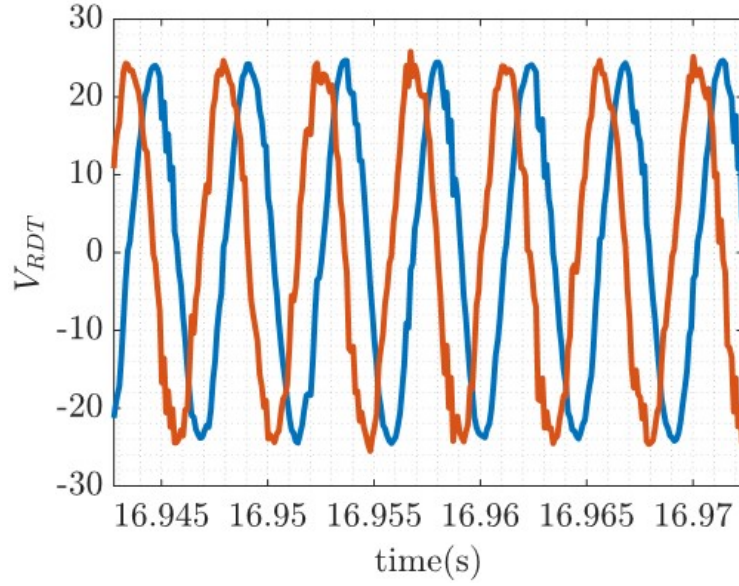


Figure 4.30: V_{RDT} in stationary reference frame when operating at 1500rpm and $i_q^r = -300A$.

4.5 Conclusion

This chapter presented an adaptive estimation strategy for I_{pm} parameter that is a function of the rotor temperature, and its efficacy and ease of implementation are demonstrated. It is shown that above an operating speed where V_{RDT} drop becomes negligible when compared to the back emf, the proposed estimation strategy accurately converges to the actual I_{pm} parameter, thereby avoiding the need for rotor temperature measurement. Demagnetization or some fault conditions lead to reduced level of PM excitation thus, an accurate estimation of I_{pm} is essential for high performance torque control and in fault or health monitoring applications to capture the amount of demagnetization. The proposed estimator is validated in simulation where the PMSM machine model uses the experimental characterization $\lambda - i$ relationship from the previous chapter. Experimental results also confirm the effectiveness of the proposed estimator strategy. It is also noted that the proposed estimator has decreased performance at lower speeds due to sensitivity to the resistive voltage drop. Higher accuracy is demonstrated at higher speeds. The implementation strategy of the proposed I_{pm} estimator above a certain speed can be performed by having the adaptation gain as a step/switching function where the adaptation gain at lower speed is 0 and it is set to be the designed value above the speed of interest, which essentially turns "on" the adaptation mechanism above a certain speed.

A robust implementation of the proposed adaptive law can be integrated that includes a

”switching σ -modification” that essentially bounds the parameter and ensures that it does not exceed defined limits (i.e. $|\hat{\theta}_i(t)| \leq M_o$)[41]. In this case, we know that considering the normal operating temperature of the machines the uncertainty in PM excitation does not exceed 4%, we observe no change in the estimator performance as the estimation error is less than 1% that is much lower than the extreme bounds of the I_{pm} parameter.

CHAPTER 5

Accurate and High Performance Torque Regulation

5.1 Introduction

Accurate torque regulation is critical in high-performance electric drives. However, the torque in electric drives is indirectly regulated, typically through current regulation. The accuracy of torque regulation therefore depends directly upon the modeled relationship between torque and current/flux, which in turn depends upon the accuracy of the magnetics model used. As noted previously, accurate magnetic models are challenging due to the nonlinear (NL) saturation phenomena in electrical steel as well as the variation of the permanent magnet (PM) excitation with rotor temperature. Leveraging the proposed NL magnetic model in the previous chapter, the goal is to develop a comprehensive control architecture for accurate torque regulation of PMSMs. The proposed controller leverages the proposed NL machine magnetics model that decouples NL magnetic saturation and temperature variation of PM and demonstrates its use in a real time controller. The proposed controller ensures optimal performance across the full range of operation and can be implemented for different types of PMSMs.

5.2 Sensitivity analysis

In Chapter 2 there are ample works that estimate the PM flux linkage for the purposes of accurate torque regulation, rotor temperature estimation and other applications like sensorless control. It was established in chapter 3 that modeling the PM excitation as a flux linkage source is inaccurate, especially in the region of deep saturation and rotor temperature variation.

The benefit of the proposed model with PM excitation represented as I_{pm} for torque regu-

lation that helps in determining accurate current/flux references from a torque reference can also be demonstrated by analyzing the sensitivity of torque with respect to PM excitation. For the baseline model, the torque expression is given by

$$\tau_{em} = \frac{3N_p}{4} \left[f_d(\vec{i}^r) i_q^r - f_q(\vec{i}^r) i_d^r + \Lambda_{pm}(T_r) i_q^r \right].$$

while the torque expression based on the proposed model is

$$\tau_{em} = \frac{3N_p}{4} \left[h_q \left(\vec{\lambda}^r \right) \lambda_d^r - h_d \left(\vec{\lambda}^r \right) \lambda_q^r + I_{pm}(T_r) \lambda_q^r \right]$$

5.2.1 Procedure for analyzing torque error with uncertainty in PM excitation

We use the two-dimensional finite element analysis (FEA) results from ANSYS for BMW i3 IPM machine as the basis for comparison. The process used to analyze the torque error with uncertainty in PM excitation for the two models is as follows:

1. Using FEA, calculate the flux linkages λ_d^r and λ_q^r and the electromagnetic torque τ_{em} over a range of currents i_d^r and i_q^r with the rotor temperature set at $T_1 = 20^\circ C$.
2. Use the flux-linkage/current data to determine the values of $I_{pm}(T_1)$ (used in the proposed model) and $\Lambda_{pm}(T_1)$ (used in the baseline model). The value of Λ_{pm} is determined from λ_d^r by setting the currents to zero, and the value of I_{pm} is determined by finding the value of i_d^r that sets $\lambda_d^r = 0$.
3. Generate a look-up table for the $\vec{\lambda}^r = \vec{g} \left(\vec{i}^r + \vec{i}_{pm}^r \right)$ mapping of the proposed model using $20^\circ C$ FEA data. Note that we determine the $\vec{g}(\cdot)$ mapping rather than the $\vec{h}(\cdot)$ mapping to better enable the comparison.
4. Generate a look-up table for the $\vec{f}(\vec{i}^r)$ in (3.30) for the baseline model using the $20^\circ C$ FEA data.
5. Repeat steps 1) and 2) for a different rotor temperature (T_2).
6. Compute the torque with the proposed model using the nonlinear look-up table for $\vec{g}(\cdot)$ determined from the T_1 data and with ± 4 % uncertainty in $I_{pm}(T_2)$ for the proposed model.
7. Compute the torque with the comparison model using the nonlinear look-up table $\vec{f}(\vec{i}^r)$ from the T_1 data and with ± 4 % uncertainty in $\Lambda_{pm}(T_2)$ for the baseline model.

8. Compare the results from the previous steps 6) and 7) with the torque obtained from FEA results at the new rotor temperature (T_2).

Fig. 5.1 plots the absolute error in torque for -4% uncertainty in PM excitation. In the region of heavy saturation and significant rotor temperature variation, the baseline model has torque error as high as 60Nm, whereas in the proposed model it is about 16Nm, i.e. 4 times lower. Similarly, for +4% uncertainty in PM excitation, the baseline model has a torque error roughly above 21Nm in the region of heavy saturation while the proposed model shows about 3Nm error in the same region and with rotor temperature variation.

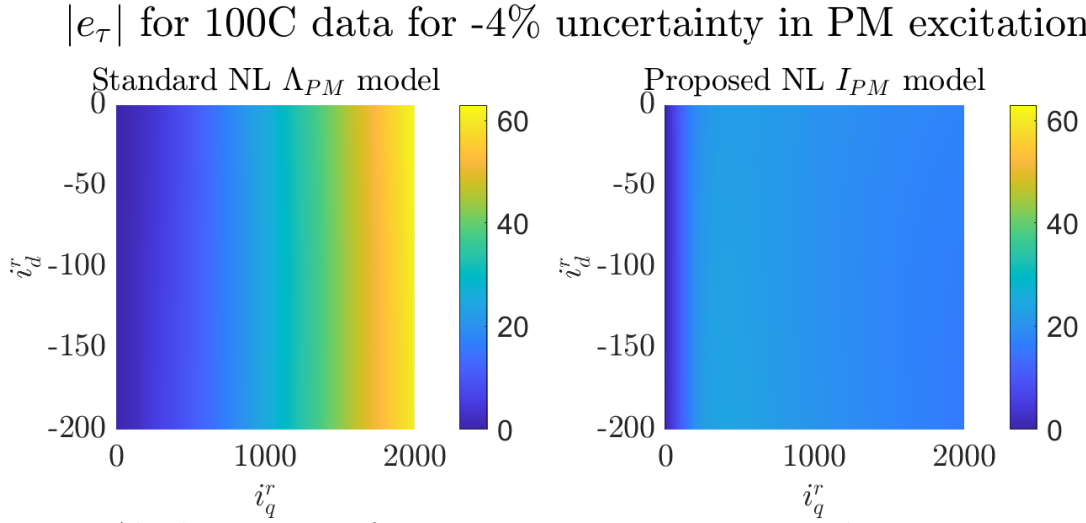


Figure 5.1: Absolute τ_{em} error for Interior Permanent Magnet machine at rotor temperatures $T_2 = 100^\circ C$ and -4% uncertainty in PM excitation.

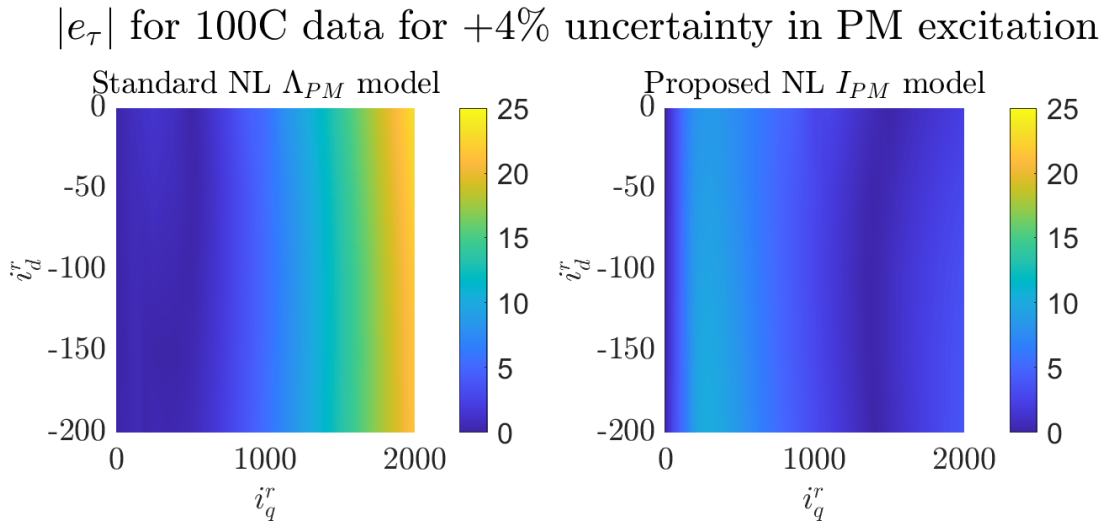


Figure 5.2: Absolute τ_{em} error for Interior Permanent Magnet machine at rotor temperatures $T_2 = 100^\circ C$ and +4% uncertainty in PM excitation.

5.3 Proposed Controller Architecture

The proposed model's torque expression as shown in eq.(??) is a function of flux linkage thereby naturally leading to a flux regulator as a preference for the regulation scheme as opposed to current regulator. Using the proposed non-linear magnetics model, a controller (Fig. 5.3) comprised of the following is presented:

- A non-linear Luenberger observer using the non-linear magnetics model to estimate flux linkage,
- An adaptive parameter estimation strategy for permanent magnet current (discussed in the previous chapter),
- A flux regulator that tracks the reference flux linkage commands, and
- A method for generating reference flux linkage references for a given torque reference.

A high level block diagram of the proposed controller scheme is shown in Fig. 5.3.

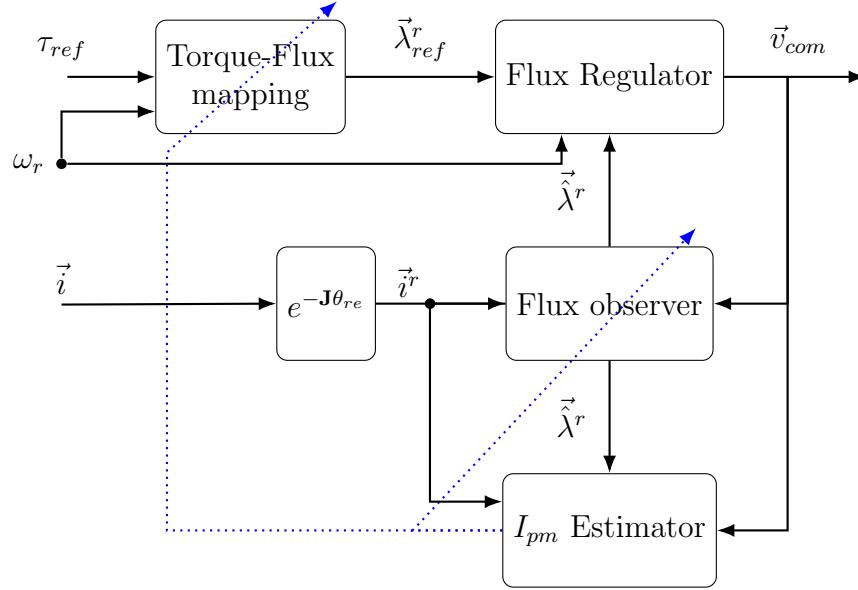


Figure 5.3: Example torque regulator based on proposed magnetics model

5.3.1 Method to generate flux linkage reference commands for a given torque command using the proposed model

In the previous chapter, the ease of implementation of the proposed model using LUT for a real-time controller application is presented. The most straight-forward and easy way to

compute flux linkage reference from a given torque command is by constructing 3D LUTs with reference torque, speed and I_{pm} parameter as the inputs. The inputs will be used to determine flux linkage references based on the operating condition from the pre-computed LUT data. The I_{pm} parameter captures the temperature effects that are essential for accurate torque regulation. This I_{pm} parameter can be obtained by implementing the online adaptive estimator that is discussed in the previous chapter. An example implementation of with τ_{ref} , ω_r (speed) and I_{pm} as inputs and flux linkage references (λ_{dcomm}^r and λ_{qcomm}^r) as outputs is shown in Fig. 5.4 below.

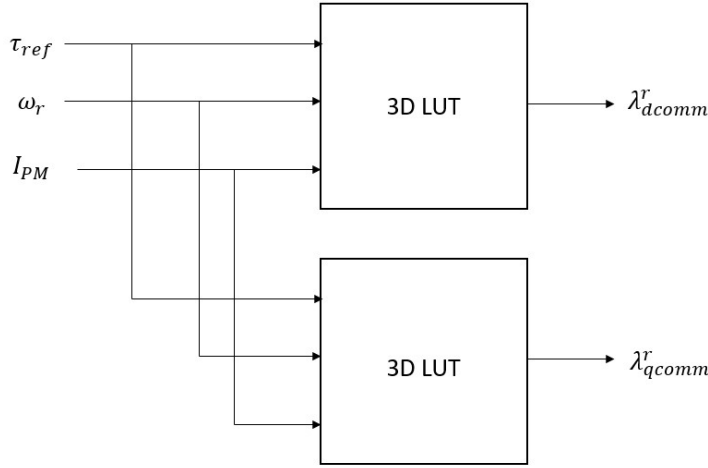


Figure 5.4: Example implementation for generating flux linkage commands.

5.3.2 Non-Linear Luenberger Observer

A NL Luenberger observer with output current injection that estimates flux linkage is shown in Eq. 5.1, where “ $\hat{\cdot}$ ” denotes an estimated value and $\mathbf{Ro}$ is the observer output injection gain. The output injection gain acts as a resistance and helps in tuning the convergence of the Luenberger observer. Higher observer gain reduces the sensitivity to uncertainty in the stator resistance parameter and helps in faster convergence of the observer, but also introduces noise from the measured currents into the estimated flux linkage. Based on a linear magnetics model, the observer gain can be set as shown in 5.2.

$$\vec{\hat{\lambda}}^r = \vec{v}^r - R\vec{\hat{i}}^r + \omega_{re}\mathbf{J}\vec{\hat{\lambda}}^r + \mathbf{Ro}\left(\vec{i}^r - \vec{\hat{i}}^r\right), \quad (5.1)$$

$$\mathbf{Ro} = \omega_p \mathbf{L} - R\mathbf{I}, \quad (5.2)$$

where $\mathbf{L}$ is the inductance matrix and ω_p is the observer corner frequency. The implementation of the NL Luenberger observer is shown in Fig. 5.5, where LUTs are used to map the computed flux linkage to currents in the observer.

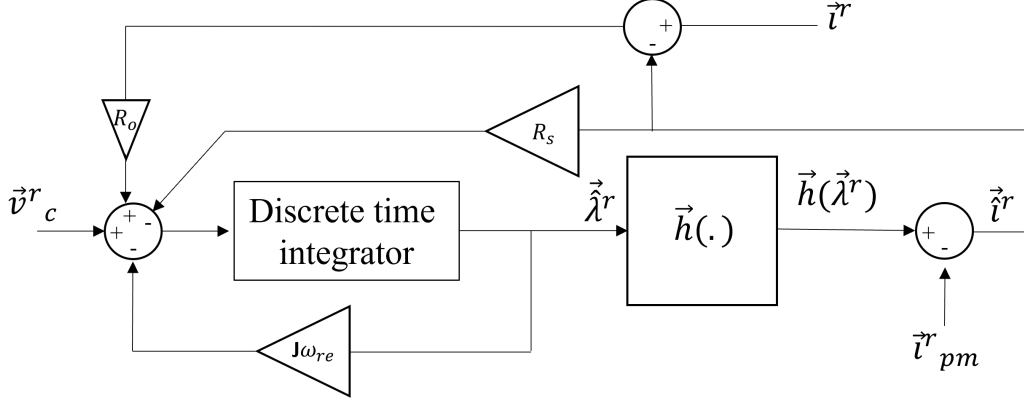


Figure 5.5: Discrete-time NL Luenberger observer implementation in Simulink.

5.3.3 Flux Regulator

The flux regulator used in the control scheme is a synchronous regulator implemented in the rotor reference frame where the variables regulated are flux linkages as opposed to currents. The voltage command from the standard current regulator is shown in eq.(5.6) and the voltage command from the flux regulator is given by eq.(5.7). The standard current regulator block diagram is shown in Fig. 5.6, where the cross-coupling compensation is typically incorporated by considering the linear inductance parameters and PM flux linkage. The block diagram of the flux regulator is shown in Fig. 5.7 which incorporates the speed dependent emf compensation by using the estimated flux linkages.

$$\begin{aligned} \vec{v}^r &= \omega_{re} \mathbf{J} \mathbf{L} \vec{i}^r + \omega_{re} \mathbf{J} \vec{\lambda}_{pm} + K_{pi} \left(\vec{i}^r - \vec{i}^r \right) + K_{ii} \vec{s}_i^r, \\ \vec{s}_i^r &= \vec{i}^r - \vec{i}^r, \end{aligned} \quad (5.3)$$

$$\begin{aligned} \vec{v}^r &= \omega_{re} \mathbf{J} \vec{\lambda}^r + K_{p\lambda} \left(\vec{\lambda}^r - \vec{\lambda}^r \right) + K_{i\lambda} \vec{s}_\lambda^r, \\ \vec{s}_\lambda^r &= \vec{\lambda}^r - \vec{\lambda}^r, \end{aligned} \quad (5.4)$$

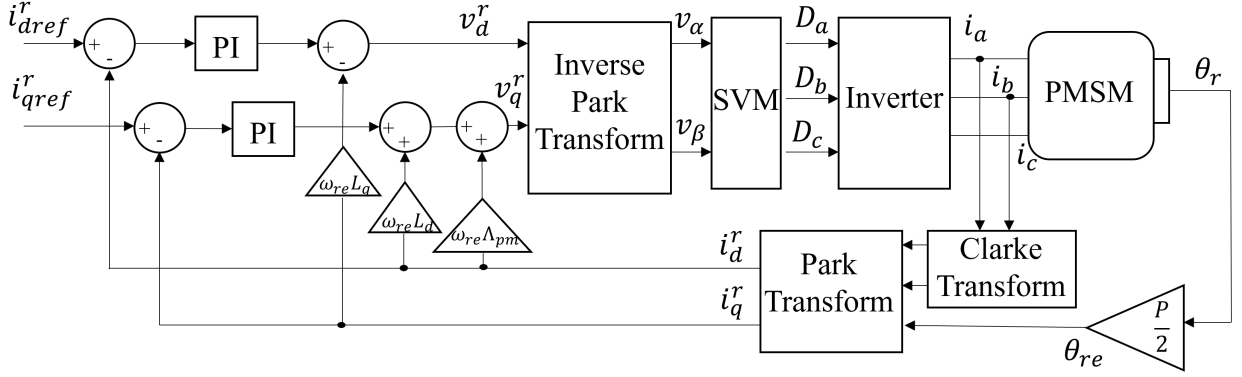


Figure 5.6: Standard current regulator block diagram.

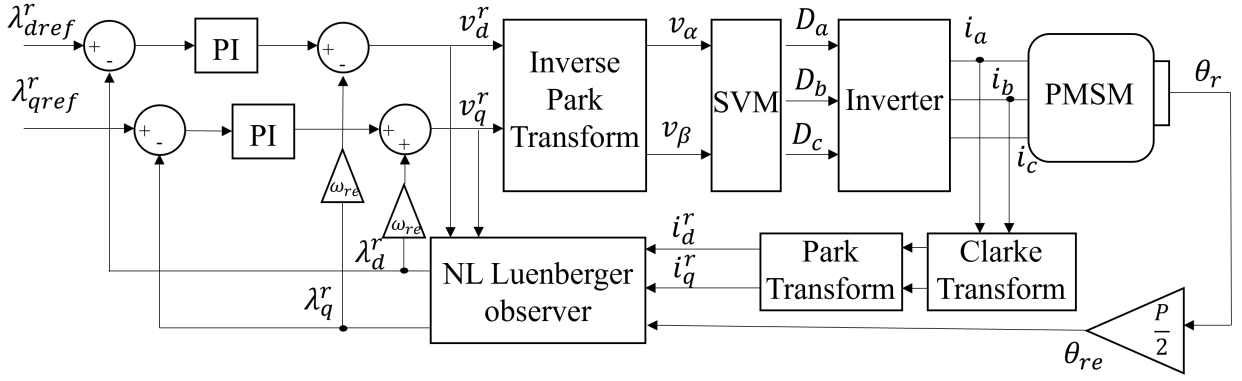


Figure 5.7: Proposed flux linkage regulator block diagram.

Since the UQM machine under test does not magnetically saturate under normal operating points, the gains of the controller using linear machine model parameters can be set as shown below [32]:

$$\begin{aligned} K_{pi} &= \Omega_{bw} L, \\ K_{ii} &= \Omega_{bw} R, \end{aligned} \quad (5.5)$$

$$\begin{aligned} K_{p\lambda} &= \Omega_{bw}, \\ K_{i\lambda} &= \Omega_{bw} \frac{R}{L}, \end{aligned} \quad (5.6)$$

where Ω_{bw} represents the bandwidth of the regulators.

5.4 Controller Performance

5.4.1 Simulation results

A flux regulator and current regulator are implemented and compared in the MATLAB Simulink simulation environment. The flux linkage/current relationship obtained for the UQM SMPM machine from the characterization in chapter 3, section 3.4.4.3 is used in the Simulink machine model. A discrete-time controller with a switching frequency of 10kHz is simulated to closely match the experimental controller setup. The sample rate of the discrete-time controller is set to be the same as the switching frequency, allowing synchronous sampling. The controller gains for both regulators are designed to achieve the same bandwidth (300Hz) and chosen as described in the previous section. The results shown here compare the flux regulator with the standard current regulator with cross coupling compensation using the inductance and PM flux linkage parameter.

For an operating condition with speed set at a constant 3000rpm, a torque step from -79Nm to -222Nm for both current regulator and flux regulator is shown in Fig. 5.8, which highlights the improved transient response of the flux regulator when compared to current regulator. The improvement in the transient performance for flux regulation over current regulation observed in Fig.5.8 is due to accurate cross coupling compensation in flux regulation. As can be seen in section 3.4.4.2, λ_d^r is not constant (i.e. $\lambda_d^r \neq \Lambda_{pm}$) for the operating regions with increased torque reference. In the current regulator, as can be seen from eq.(5.6) the cross coupling compensation consists of $\omega_{re}\Lambda_{pm}$ for V_q^r command which is inaccurate at higher torque commands, whereas in flux regulation, the compensation directly takes into account the flux linkage computed from the NL Lunberger observer, thereby having accurate cross coupling compensation and improved transient performance.

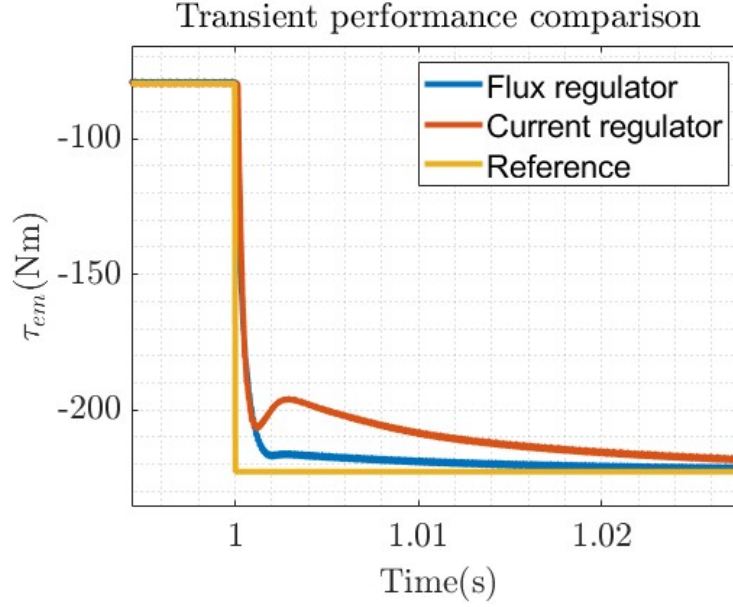


Figure 5.8: Comparison of transient performance for a torque step between current and flux regulator at 3000rpm.

5.4.1.1 Simulation results at voltage limits of inverter

Fig. 5.9 shows the improved performance of the flux regulator when the drive is operating at the voltage limits where the regular current regulator has overshoot and sluggish performance. In order to hit the voltage limits, the operating speed of the simulation is increased to 3600rpm. The cross-coupling compensation is kept the same as from the previous simulation test case for both the regulators. The improvement in the transient performance due to inaccurate cross-coupling compensation in the case of current regulation can be seen when operating at the voltage limits, as seen in Fig. 5.9.

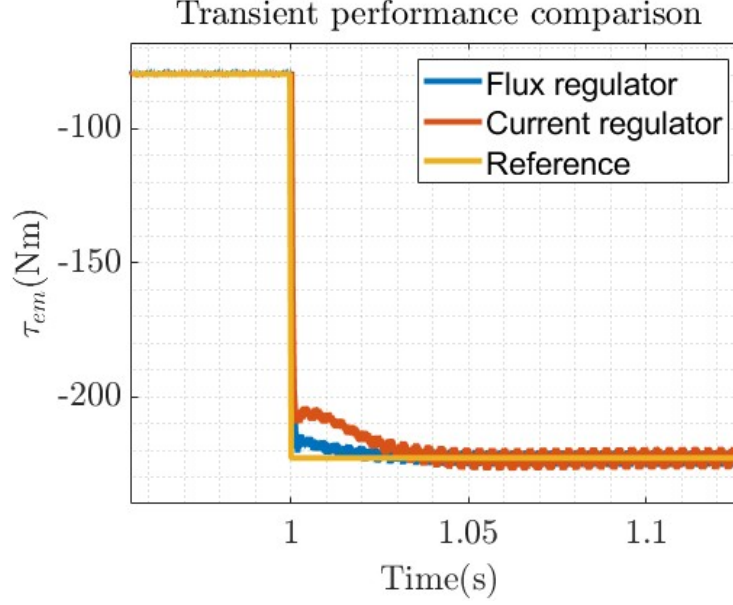


Figure 5.9: Comparison of transient performance for a torque step between current and flux regulator when operating at voltage limits (3600rpm).

5.4.1.2 Simulation results with noise

Finally, to demonstrate the benefits of the inherent filtering associated with flux estimation, normally (Gaussian) distributed random noise with magnitude 40A is injected into the three phase current measurements for both controllers in the simulation. Fig.5.10 shows the noise injected into the Phase A current measurement. The flux regulator incorporates a NL Luenberger observer that inherently filters the noise in current measurements, thus leading to a cleaner voltage command and duty cycle waveform. We can clearly observe that the flux regulator has lower noise in the torque computed from the cross product of flux linkage and measured current when compared to the current regulator in Fig. 5.11. Fig. 5.12 compares the noise in the phase A duty cycle waveforms in both controllers, highlighting the improvement in duty cycle waveform in flux regulator, which helps with better voltage utilization without triggering protections in the hardware implementation.

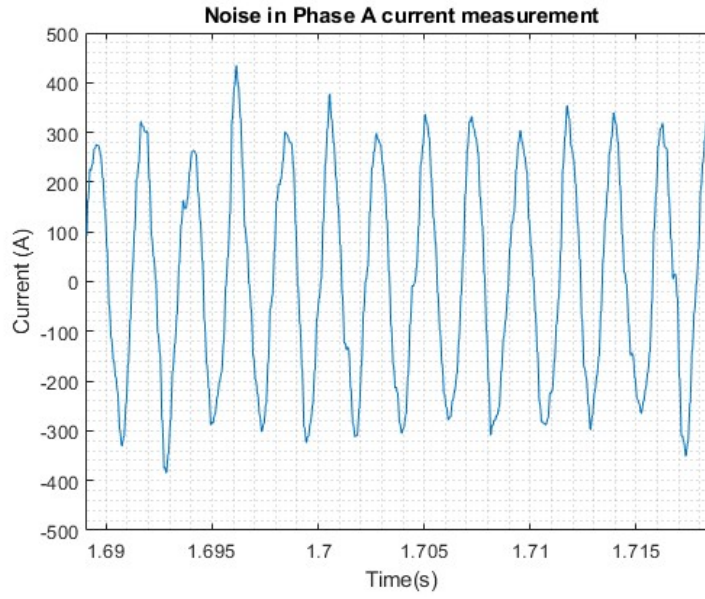


Figure 5.10: Noise in Phase A current.

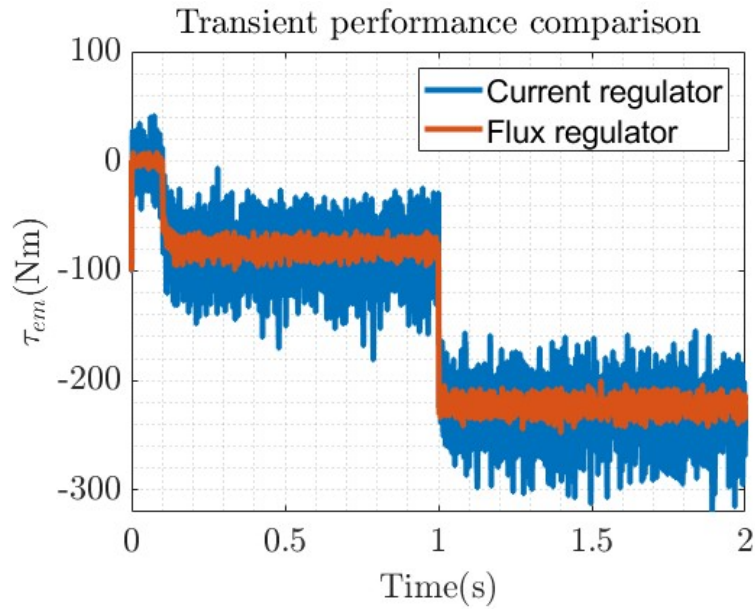


Figure 5.11: Comparison of transient performance for a torque step between current and flux regulator with noise in the three phase current measurements.

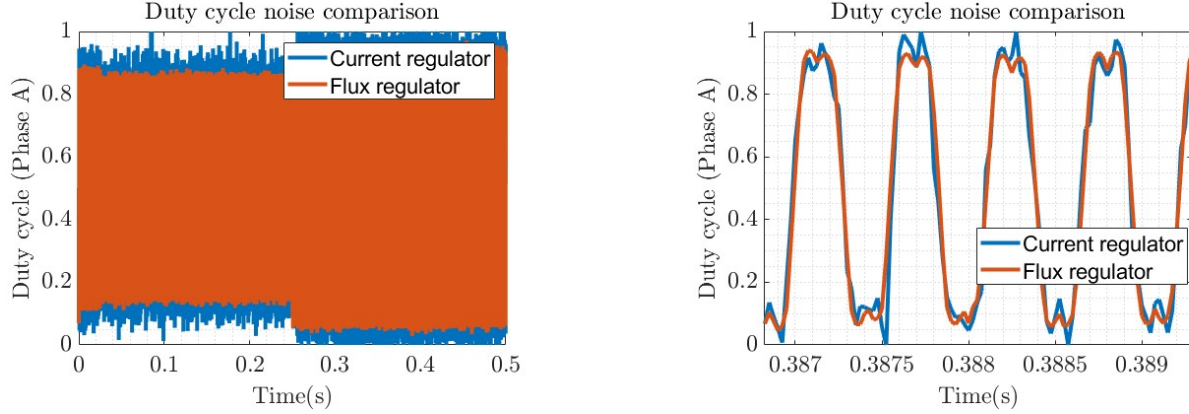


Figure 5.12: Phase A duty cycle comparison in presence of noise in current measurements. Right: Zoomed in portion of the duty cycle waveform

5.4.2 Experimental results

Using experimental setup in Michigan Power and Energy Lab, the improvement in controller performance due to actual noise present in the system is demonstrated herein. Both, the proposed above flux regulator and the standard current regulator are implemented and tested at 2000rpm and torque steps as shown in Fig. 5.13, i.e the current levels are the same in both regulators. The controller gains for current regulator were adjusted to account for the V_{RDT} , and to keep the comparison consistent, the flux regulator gains were chosen by dividing the current regulator gains with inductance term as described earlier. Moreover, the test results analyzed in this section are under steady state conditions and not for transient performance. The voltage commands in the stationary reference frame for both the regulators is shown in Fig. 5.14 that clearly highlights lower influence of noise in the controller for flux regulator as opposed to the current regulator. Fig. 5.16 illustrates the noise in the three phase duty cycle in the current regulator and flux regulators respectively. It can be seen that, at higher speed and high current when the noise in the system is significant, flux regulation ensures the best voltage utilization due to less impact on the controller due to noise.

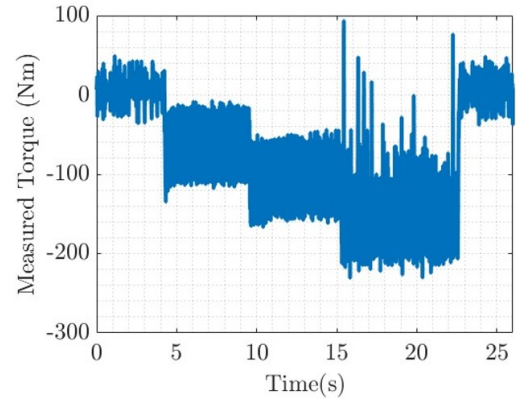
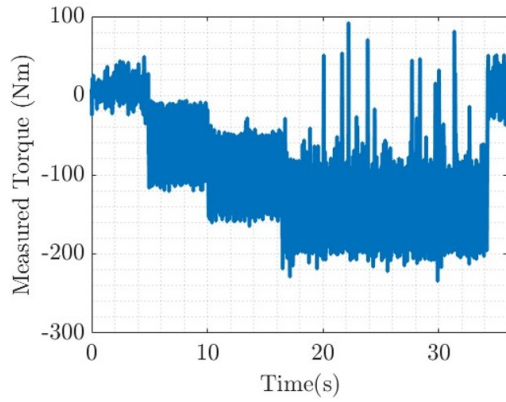


Figure 5.13: Three phase measured torque (2000rpm) Left: Current regulator Right: Flux regulator

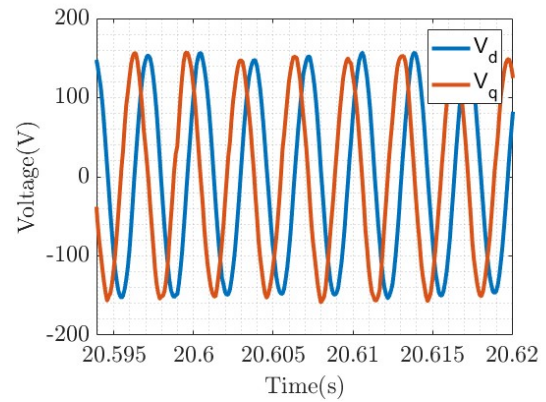
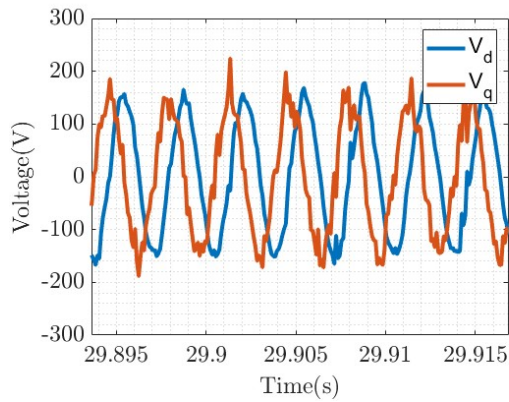


Figure 5.14: Voltage commands in the stationary reference frame (2000rpm) Left: Current regulator Right: Flux regulator

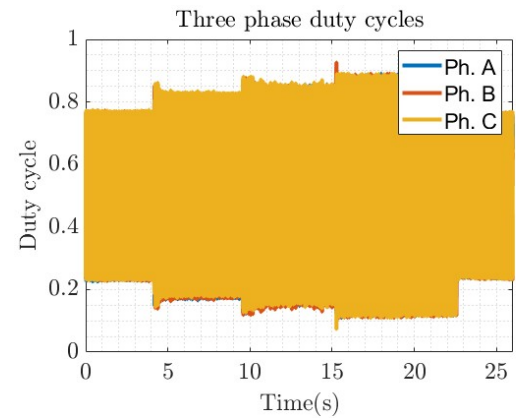
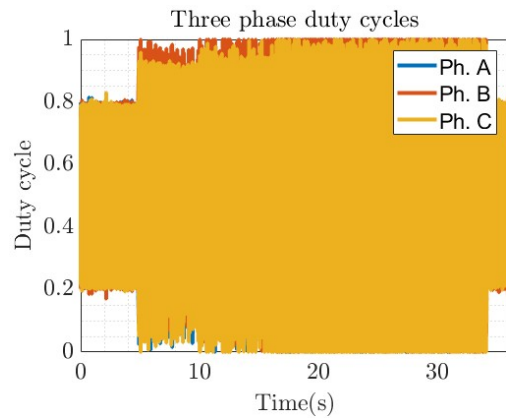


Figure 5.15: Three phase duty cycle (2000rpm) Left: Current regulator Right: Flux regulator

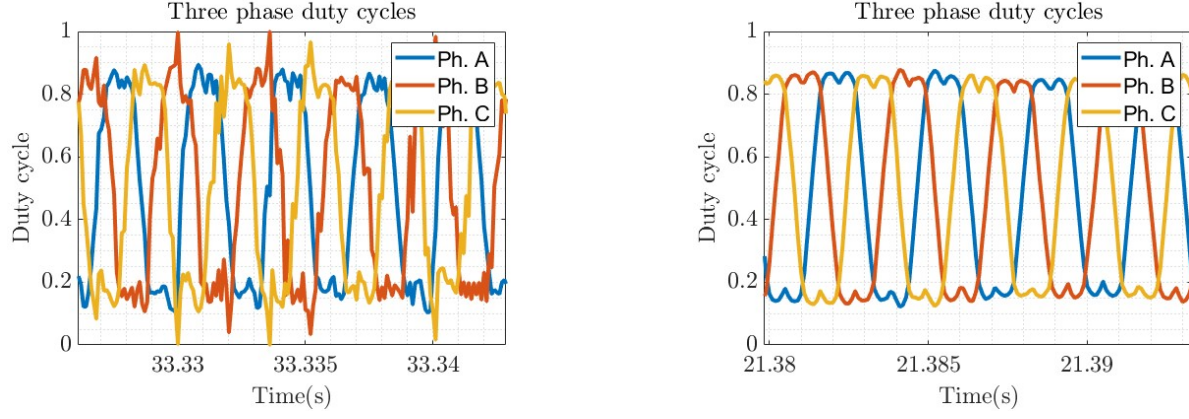


Figure 5.16: Zoomed in portion of three phase duty cycle (2000rpm) showing the noise Left: Current regulator Right: Flux regulator

5.4.2.1 Operation of Flux Regulator at Different Speeds

The proposed flux regulator was tested at different speeds and is shown to be stable at all operating conditions. Fig. 5.17 shown the stable operation of the flux regulator at zero speed and Fig. 5.18 demonstrate the stable operation at 500rpm.

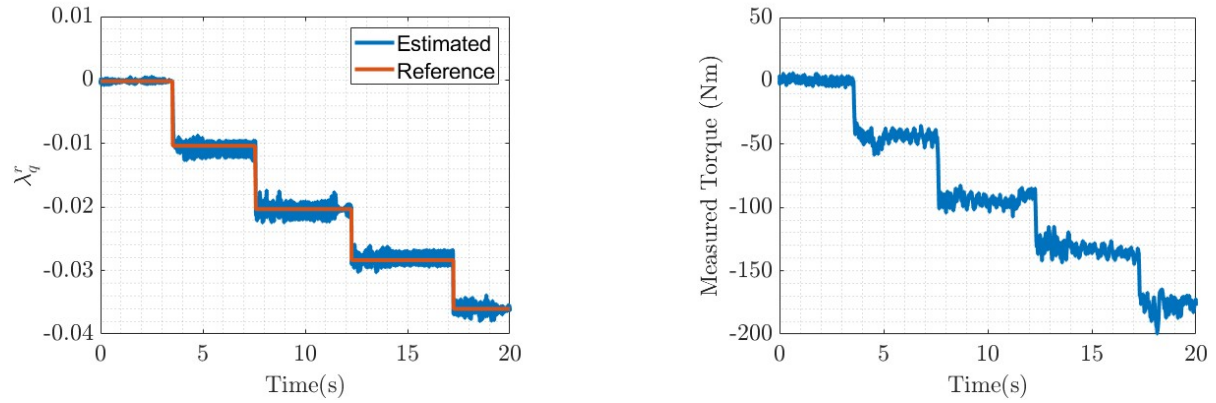


Figure 5.17: Zero speed operation Left: λ_q^r tracking Right: Measured torque

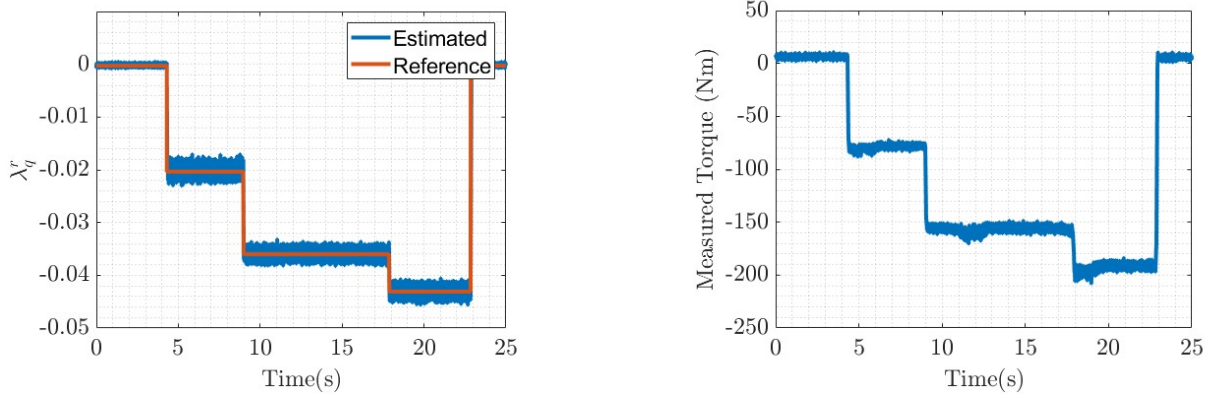


Figure 5.18: Mid-speed (500rpm) operation Left: λ_q^r tracking Right: Measured torque

5.5 Conclusion

Building upon the proposed non-linear model naturally leads to flux regulation as opposed to current regulation for indirect torque control. Flux regulation is shown to be robust and has better transient performance than current regulation under dynamic conditions. One of the main challenges, however, is the accurate estimation of flux linkage considering the non-linear dynamics. We implement a non-linear Luenberger observer with output current injection enabling in accurate flux linkage estimation for flux regulation. Thus, the major benefits of the proposed torque regulator structure can be summarised as follows:

- Accurate torque regulation considering non-linearity due to saturation and variation in rotor temperature
- Improved transient performance compared to standard current regulator due to superior cross-coupling compensation
- Accurate flux estimation using the proposed non-linear magnetics model
- Increased voltage utilization and reduced torque ripple due to inherent filtering of noise by the flux linkage estimator

CHAPTER 6

Conclusion and Future Work

6.1 Conclusion

This dissertation proposed a new non-linear magnetics model for Permanent Magnet Synchronous Machines (PMSMs) that decouples magnetic saturation from the effects of rotor temperature variations. Leveraging the proposed model, a high-performance controller is developed and an accurate adaptive estimation strategy for the permanent magnet excitation parameter is proposed. Chapter 3 introduces the proposed model for the PMSM, followed by simulation validation for three different machine designs using ANSYS FEA. The later portion of the chapter focuses on hardware validation of a SMPM in the Michigan Power and Energy Lab demonstrating the improved accuracy of the proposed model. It is shown that the proposed model accurately predicts torque and is relatively insensitive to the PM excitation when compared to the baseline model from literature, especially in the region of heavy magnetic saturation and significant rotor temperature variation.

A high-performance torque controller is explored demonstrating the effectiveness of the proposed model in a real-time controller implementation. The controller architecture is presented that consists of an FOC implementation of a flux regulator. A NL luenberger observer used in the controller to estimate flux linkage is also discussed. Both simulation and experimental results are shared that highlights the improved transient performance due to better compensation of the speed dependent cross coupling terms and reduced noise in the controller waveforms.

Finally, an adaptive estimation algorithm for I_{pm} is proposed in Chapter 5 that eliminates the need of a rotor temperature sensor. The importance of the I_{pm} parameter as part of the proposed model is discussed in Chapters 2 and 3. The proposed algorithm ensures accurate torque regulation, captures the rotor temperature-dependent aspects of the machine and can also be used for condition monitoring/fault monitoring purposes. The proposed estimation strategy is validated both on simulation and on experimental hardware.

6.2 Future Work

This section contains potential future work based on the contributions of the thesis work. The work can be extended in the following ways:

6.2.1 Accurate and optimal torque controller formulation

Another approach as part of the future work on controller synthesis is a formulation to determine flux linkage commands as a real-time optimization problem taking into account the current and dynamic voltage constraints while minimizing the steady state machine losses.

A possible formulation of the optimization problem is given below:

$$\begin{aligned}
& \min_{u_d, u_q} \sum_{k=1}^{N-1} (Qe^2[k] + \vec{u}[k]^T \mathbf{R} \vec{u}[k]) + \alpha P_{ssloss} \left(\vec{\lambda}^r[N] \right) \\
& \text{s.t.} \\
& \vec{\lambda}^r[k+1] = \vec{\lambda}^r[k] + \vec{u}[k], \\
& g(\vec{\lambda}^r[\cdot]) \leq V_{max}, \\
& \text{where} \\
& g(\vec{\lambda}^r[\cdot]) = \left\| w_{re} \mathbf{J} \vec{\lambda}^r[k] + \frac{(\vec{\lambda}^r[k] - \vec{\lambda}^r[k-1])}{\Delta t} \right\| \\
& y[k] = \frac{3N_p}{4} (\lambda_q^r[k] \hat{I}_{pm} - \lambda_q^r[k] h_d(\lambda_d^r[k], \lambda_q^r[k]) + \lambda_d^r[k] h_q(\lambda_d^r[k], \lambda_q^r[k])) \\
& e[k] = r[k] - y[k]
\end{aligned}$$

In the above problem formulation, the last term in the objective function ensures that the final operating point is the one that minimizes steady state machine losses. The optimization problem can be updated at a lower rate than the flux regulator, reducing the computational effort required for solving it in real time. This formulation will ensure that the final operating point settles to the steady-state minimum loss operating point, and that the voltage constraint is not violated during the transient condition. This implementation will be investigated further as part of future work. A computationally efficient real-time solver for the above formulation will also be looked into.

Preliminary simulation results were generated for a test condition of $\tau_{ref} = 80Nm$, $w_{re} = 250rpm$ and initial operating condition $\tau=10Nm$, Fig. 6.1 shows the trajectory of the flux linkage reference commands generated as result of the constrained optimization problem.

It can be noted from Fig. 6.2 that the final operating point settles to the steady-state minimum loss operating point, and Fig. 6.3 highlights that the voltage constraint is not violated during the transient. Fig.6.4 shows the step response for the given torque reference.

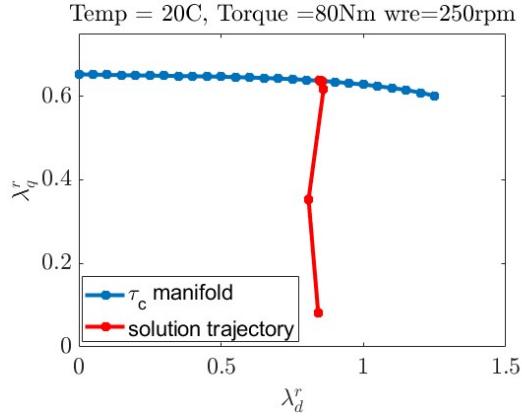


Figure 6.1: $\vec{\lambda}_{ref}^r$ trajectory

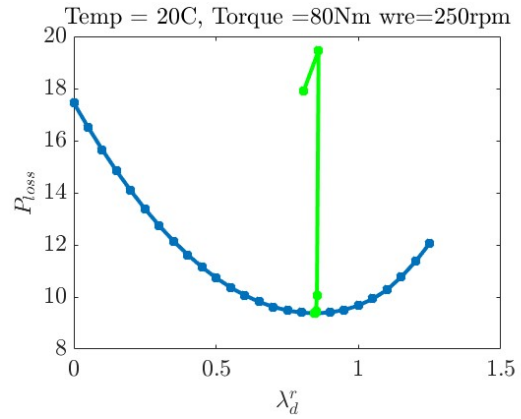


Figure 6.2: Steady state P_{loss}

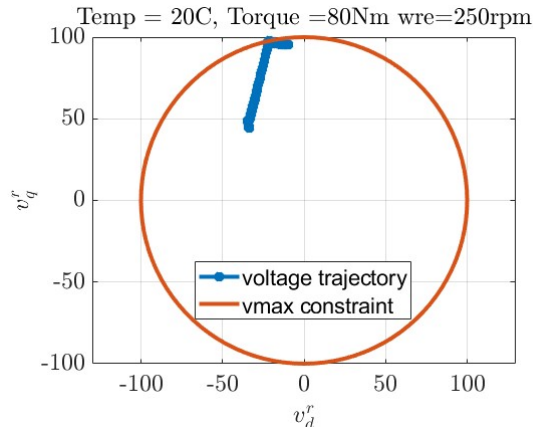


Figure 6.3: Voltage trajectory

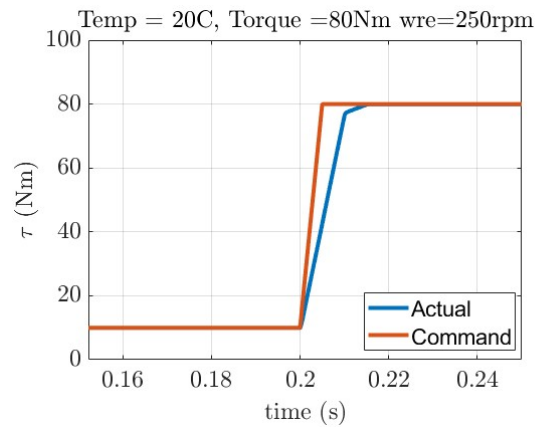


Figure 6.4: Torque step-response

Another potential area of future work is the exploration of the Model Predictive Control implementation of the flux regulator instead of FOC. An MPC formulation can include the torque-flux linkage mapping problem along with switching state selection based on the desired objective.

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