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NOVEL MODELING, TESTING AND CONTROL APPROACHES TOWARDS ENERGY EFFICIENCY IMPROVEMENT IN PERMANENT MAGNET SYNCHRONOUS MOTOR AND DRIVE SYSTEMS

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NOVEL MODELING, TESTING AND CONTROL APPROACHES TOWARDS ENERGY EFFICIENCY IMPROVEMENT IN PERMANENT MAGNET SYNCHRONOUS MOTOR AND DRIVE SYSTEMS

By

Aiswarya Balamurali

A Dissertation
Submitted to the Faculty of Graduate Studies through the Department of Electrical & Computer Engineering in Partial Fulfillment of the Requirements for the Degree of Doctor of Philosophy at the University of Windsor

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NOVEL MODELING, TESTING AND CONTROL APPROACHES TOWARDS
ENERGY EFFICIENCY IMPROVEMENT IN PERMANENT MAGNET
SYNCHRONOUS MOTOR AND DRIVE SYSTEMS

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DECLARATION OF CO-AUTHORSHIP / PREVIOUS PUBLICATION

I hereby declare that this dissertation incorporates material that is result of joint research, as follows: This dissertation includes the outcome of publications which also have co-authors who are / were graduate students or post-doctoral fellows supervised by Dr. Narayan Kar. The co-author list also includes Dr. Voiko Loukanov, from our industrial partner, D&V Electronics. I am aware of the University of Windsor Senate Policy on Authorship and I certify that I have properly acknowledged the contribution of other researchers to my thesis, and have obtained written permission from each of the co-authors to include the above materials in my thesis. In all cases, only primary contributions of the author towards these publications are included in this dissertation. I certify that, with the above qualification, this thesis, and the research to which it refers, is the product of my own work. The contribution of co-authors was primarily the guidance and assistance in experimentation, data analysis, and manuscript review and improvement.

This dissertation includes materials and extended work of research from nine original papers that have been published/accepted/to be submitted for publication in peer reviewed IEEE Transactions and international conferences, as follows:

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ABSTRACT

This thesis investigates energy efficiency improvement in permanent magnet synchronous motor (PMSM) and drive system to achieve high-performance drive for practical industrial and primarily, traction applications. In achieving improved energy efficiency from a system level, this thesis proposes: (1) Accurate modeling and testing of loss components in PMSM considering inverter harmonics; (2) Easy-to-implement, accurate parameter determination techniques to understand variations in motor parameters due to saturation, cross-saturation and temperature; and (3) Control methodologies to improve system level efficiency considering improved loss models and parameter variations.

An improved loss model to incorporate the influence of motor-drive interaction on the motor losses is developed by taking time and space harmonics into account. An improved winding function theory incorporating armature reaction fields due to fundamental and harmonic stator magnetic fields is proposed to calculate the additional harmonic losses in the PMSM. Once all contributing losses in the motor are modelled accurately, an investigation into control variables that affect the losses in the motor and inverter is performed. Three major control variables such as DC link voltage, switching frequency and current angle are chosen and the individual losses in the motor and inverter as well as the system losses are studied under varying control variables and wide operating conditions. Since the proposed loss as well as efficiency modeling involves machine operation dependent parameters, the effects of parameter variation on PMSM due to saturation and temperature variation are investigated. A recursive least square (RLS) based multi-parameter estimation is proposed to identify all the varying parameters of the PMSM.
to improve the accuracy and validity of the proposed model. The impact of losses on these parameters as well as the correct output torque considering the losses are studied. Based on the proposed loss models, parameter variations and the investigation into control variables, an off-line loss minimization procedure is developed to take into account the effects of parameter variations. The search-based procedure generates optimal current angles at varying operating conditions by considering maximization of system efficiency as the objective.

In order to further simplify the consideration of parameter variations in real-time conditions, an on-line loss minimization procedure using DC power measurement and loss models solved on-line using terminal measurements in a PMSM drive is proposed. A gradient descent search-based algorithm is used to calculate the optimal current angle corresponding to maximum system efficiency from the input DC power measurement and output power based on the loss models. During the thesis investigations, the proposed models and control techniques are extensively evaluated on a laboratory PMSM drive system under different speeds, load conditions, and temperatures.
This dissertation is dedicated to

my parents,

Mrs. Rajapriya Balamurali and Mr. N. V. Balamurali

and my best friend and husband,

Anindya Banerji,

for their unconditional love and support!

Paatti, I wish you were here for me to read this to...
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# TABLE OF CONTENTS

DECLARATION OF CO-AUTHORSHIP / PREVIOUS PUBLICATION ........ iii

ABSTRACT ....................................................................................................... vii

DEDICATION ................................................................................................. ix

ACKNOWLEDGEMENTS .............................................................................. x

LIST OF TABLES ........................................................................................... xx

LIST OF FIGURES .......................................................................................... xxi

NOMENCLATURE ........................................................................................... xxviii

LIST OF ABBREVIATIONS ............................................................................. xxx

CHAPTER 1 INTRODUCTION ...................................................................... 1

1.1 Overview of Electric Motors and Drives in Traction Applications .......... 1

1.2 Review of Factors Influencing System Level Efficiency and Performance
   Improvements in PM Based Electric Motor and Drive System ...................... 5

   1.2.1 Motor–Drive Interaction and Accuracy of Loss Models .................... 5

   1.2.2 Parameter Variations Under Varying Operating Conditions .......... 6

   1.2.3 Control Algorithm and Methodology of Implementation .............. 9

1.3 Motivation of the Thesis ........................................................................ 11

1.4 Dissertation Outline and Research Contributions towards Improved System
   Efficiency in PMSM Drives ....................................................................... 12

CHAPTER 2 COMPREHENSIVE ANALYTICAL MODELS TOWARDS STUDY OF
HARMONIC MOTOR LOSS BEHAVIOUR IN A PWM–FED PERMANENT MAGNET
MACHINE ................................................................................................... 16

2.1 Introduction ............................................................................................ 16
2.2 Sources of Harmonic Motor Losses .................................................................19

2.3 Analytical Magnetic Flux Density Derivation in PMSM .................................20

2.3.1 Stator MMF Computation Including Carrier Harmonic Ripple Current as a Function of PWM Parameters .................................................................21

2.3.2 Rotor MMF as a Function of Armature Reaction Field from Stator Excitation and Air–gap Magnetic Flux Density Derivation ...............................................24

2.4 Analytical Derivation of Harmonic Iron Losses ..............................................27

2.4.1 Procedure to Determine Harmonic Iron Losses ..........................................27

2.4.2 Simulation Results of Harmonic Iron Loss Determination at Various Operating Conditions ...................................................................................................31

2.5 Analytical Derivation of PWM Harmonic Factor Contributed by Magnet Eddy Current Losses .........................................................................................32

2.6 Experimental Investigations ...........................................................................35

2.7 Summary ......................................................................................................37

CHAPTER 3 INVESTIGATION INTO VARIATION OF PERMANENT MAGNET SYNCHRONOUS MOTOR AND DRIVE LOSSES FOR SYSTEM LEVEL EFFICIENCY IMPROVEMENT ............................................................................38

3.1 Introduction ..................................................................................................38

3.2 Mathematical Modeling of Inverter Losses ....................................................41

3.2.1 Conduction Losses of VSI Inverter .............................................................41

3.2.2 Switching Losses of VSI Inverter .................................................................41

3.3 Mathematical Modeling of Motor Losses .......................................................42
3.3.1 Copper Losses ..............................................................................................43
3.3.2 Fundamental Iron Losses .............................................................................43
3.3.3 Harmonic Iron Losses ..................................................................................46
3.4 Dependence of Inverter Losses on Control Variables .........................................46
3.5 Dependence of Motor Losses on Control Variables ...........................................49
  3.5.1 Simulation Results of Fundamental Motor Losses with Varying Current Angle ......................................................................................................................50
  3.5.2 Simulation Results of Fundamental Motor Losses with Varying DC Link Voltage ......................................................................................................................51
  3.5.3 Results of Fundamental Motor Losses with Varying Switching Frequency ..................................................................................................................53
  3.5.4 Simulation Results of Harmonic Motor Losses with Varying DC Link Voltage ..................................................................................................................................53
  3.5.5 Simulation Results of Harmonic Motor Losses with Varying Switching Frequency ..................................................................................................................54
  3.5.6 Simulation Results of Harmonic Motor Losses with Varying Current Angle ..................................................................................................................55
3.6 Study of Magnet Eddy Current Losses with Varying Control Variables Using 2–D FEA Co-Simulation ..............................................................................................................56
3.7 System Level Losses with Varying Control Variables ...........................................57
  3.7.1 Current Angle .............................................................................................57
4.3.3 On–line Multi–Parameter Determination through Two– Stage RLS Estimation Algorithm ........................................................................................................80

4.3.4 Mathematical Modeling and Simulation of Proposed Algorithm for Parameter Identification .....................................................................................................84

4.3.5 Results and Investigation of Identified Parameters from Mathematical Model ..................................................................................................................84

4.3.6 Impact of Iron Losses at Various Operating Points ..................................86

4.3.7 Experimental Validation of On–line Multi–Parameter Estimation and Results ..............................................................................................................................88

4.3.8 Conclusions on on–line parameter determination ........................................91

4.4 Conclusions .........................................................................................................92

CHAPTER 5 IMPROVED MAXIMUM EFFICIENCY CONTROL OF PERMANENT MAGNET SYNCHRONOUS MACHINES CONSIDERING EFFECTS OF CORE SATURATION AND TEMPERATURE VARIATION .................................................................93

5.1 Introduction .........................................................................................................93

5.2 Development of Non–linear Model Based Efficiency Improvement Method .......94

5.2.1 Voltage Equation Based Flux Linkage Mapping Considering Saturation and Cross–Saturation ........................................................................................................95

5.2.2 Flux Linkage Considering PM Flux Variation Due to Temperature .............99

5.3 Implementation of Optimal Current Angle Computation .................................101

5.3.1 Overview of Current Angle Derivation ..........................................................101

5.3.2 Loss Models for Inverter and Motor ............................................................102
CHAPTER 5 AN OFF-LINE METHOD USING CIRCUIT SIMULATION FOR ENERGY EFFICIENCY IMPROVEMENT IN PMSM MOTOR DRIVE SYSTEM

5.3.3 System Efficiency Computation for Optimal Current Angle Derivation

5.4 Analytical Results from the Developed Method

5.4.1 Results of Motor and System Efficiency Under Varying Operating Conditions

5.5 Experimental Validation using Laboratory PMSM

5.5.1 Experimental Setup

5.5.2 Tests at Varying Load-Speed Points Using $\gamma_{MEPA}$

5.5.3 Sweep tests for Optimal Current Angle Benchmarking

5.5.4 Effects of Developed Method Considering Saturation

5.5.5 Effects of Developed Method Considering Temperature Variation

5.5.6 Comparison with Conventional MTPA Control

5.6 Discussions and Conclusions

CHAPTER 6 ON-LINE METHOD USING DC POWER MEASUREMENT FOR ENERGY EFFICIENCY IMPROVEMENT IN PMSM MOTOR DRIVE SYSTEM

6.1 Introduction

6.2 Loss Models for On-line System Efficiency Improvement

6.2.1 Fundamental Losses in PMSM

6.2.2 Harmonic Copper and Core Losses

6.2.3 Inverter Losses

6.2.4 Total Losses
6.3 Proposed Gradient Descent Algorithm Based Maximum Efficiency Optimization Method .........................................................................................................................125
6.3.1 Search for Maximum Efficiency Angle Using Gradient Descent Algorithm ..................................................................................................................125
6.3.2 Implementation of Developed Method in Test Motor ...............................128
6.4 Experimental Investigations and Validations of the Proposed Maximum Efficiency Control Method ............................................................................................................131
6.4.1 Implementation of Developed Method in Test Motor ...............................132
6.4.2 Sweep Tests for Optimal Current Angle Benchmarking ...........................133
6.4.3 Experimental Results on Efficiency Improvement in IPMSM ....................134
6.5 Discussions on Adaptations of Developed Maximum Efficiency Control Method .........................................................................................................................141
6.5.1 Dynamic Response and Improved Adaptations ..................................141
6.5.2 Look–up Table Generation ...................................................................141
6.6 Conclusions ..............................................................................................141

CHAPTER 7 CONCLUSIONS AND FUTURE WORK ........................................143
7.1 Conclusions ..............................................................................................143
7.2 Future Work .............................................................................................145

BIBLIOGRAPHY .........................................................................................146
APPENDICES .........................................................................................155
Appendix A: Details of Side–Band Harmonic Current Derivation ....................155
A.1 Bessel Function used in Derivations.................................................................155

Appendix B: Mechanical Loss Determination............................................................155

Appendix C: Dynamic Equations for PMSM Modeling..............................................156

Appendix D: Permissions for Using Publication.......................................................158

VITA AUCTORIS ...........................................................................................................165
LIST OF TABLES

Table 2.1 Machine Design Parameters Used for Analytical Calculations .........................16
Table 2.2 Parameters of Test IPMSM Used for Validations ..............................16
Table 2.3 Magnet Losses for Various Carrier Frequencies with 200 VDC and 40 Hz Fundamental...........................................................................................34
Table 3.1 Parameters and Values of IGBT Inverter used for Analysis ..........42
Table 4.1 LSIPMSM Name Plate Data.................................................................72
Table 4.2 Results of Identified Parameters of LSIPMSM ...................................73
Table 4.3 Comparative Results of Identified Parameters for Model 1 and Model 2 ..........74
Table 4.4 Comparison of Simulated Inductance Estimation Results at $I_m=15.55$ A, $y=30.5$ Deg and 600 rpm ..............................................................84
Table 4.5 Comparison of Experimental Inductance Estimation Results at $I_m=10$ A, $y=20$ Deg and 200 rpm and 600 rpm .........................................................89
Table 5.1 Flux Linkage Coefficients at 25 °C .........................................................98
Table 5.2 Values of Current Angles Obtained from Sweep tests ......................113
Table 6.1 Parameters and values of IGBT Inverter Used for Analysis ...............124
Table 6.2 Comparison of Actual and Calculated MEA at 700 rpm, Varying $I_m$ ..........135
Table 6.3 Loss comparison at 2 A, 6 A and 10 A and 700 rpm .........................139
LIST OF FIGURES

Figure 1.1. A block diagram of Toyota Prius E–Motor and Drive ...........................................1

Figure 1.2. Efficiency maps for 2014 Honda Accord HEV at 300 V DC voltage. (a) Motor Efficiency. (b) Inverter Efficiency. (c) System Efficiency .................................................................4

Figure 2.1. Three–phase IPMSM designed and prototyped in–house with 48 slots, 8 poles distributed winding, and NdFeB35 magnets in the rotor. (a) Cross–section of the IPMSM. (b) Flux density distribution at rated condition. (c) Experimental setup with the laboratory prototyped motor ...................................................................................................................17

Figure 2.2. No–load back–EMF experimental validation of accuracy of 2–D electromagnetic analysis model used for validating developed model at 1,000 rpm ......18

Figure 2.3. Stator and rotor of the prototype IPMSM under test .............................................18

Figure 2.4. Magnitudes of frequency components in input voltage for frequency modulation=50 ..............................................................................................................................................20

Figure 2.5. Representation of the rotor and reference axes in the IPMSM analyzed ......21

Figure 2.6. Stator and rotor of the prototype IPMSM under test .............................................21

Figure 2.7. Magnetic circuit network showing one pole pair used for calculation of rotor MMF ........................................................................................................................................24

Figure 2.8. Results of the analytical model at 575 rpm, 10 kHz carrier frequency, 70 Nm and 200 V DC link voltage. (a) Stator MMF in rotor reference frame. (b) Rotor MMF considering fundamental and higher order carrier harmonics in input current .................29

Figure 2.9. Comparison of air–gap flux density calculated with respect to rotor position using only fundamental component of armature reaction field in air–gap flux density analytical model with 2–D FEA for considering only fundamental current component...29

Figure 2.10. Comparison of air–gap flux density calculated with respect to rotor position using analytical model with 2–D FEA for considering space harmonics and side–band time harmonics in current .......................................................................................................................30

Figure 2.11. Comparison of air–gap flux density calculated with respect to time using analytical model with 2–D FEA for considering space harmonics and side–band time harmonics in current .......................................................................................................................30

Figure 2.12. Calculation of iron and copper losses in IPMSM for 575 rpm and 700 rpm and comparison with 2–D FEA. (a) Results for 2A. (b) Results for 10 A ......................32
Figure 2.13. Current harmonic magnitudes used to study magnet eddy current losses.....34

Figure 2.14. Comparison of analytical and experimental results for total iron losses in stator under varying loading conditions for 200 rpm, 15 Nm and 7 kHz .........................35

Figure 2.15. Blocked rotor test for validation of developed model. (a) Test circuit representing input demagnetizing current. (b) Harmonic loss results including iron losses and magnet eddy current losses. .................................................................37

Figure 3.1. Schematic depiction of controllable losses in PMSM drive system.................38

Figure 3.2 Control scheme showing the variables for analyzing the behavior of system–level losses. ..............................................................................................................40

Figure 3.3. d– and q–axis equivalent circuit model of IPMSM incorporating iron loss resistance. (a) d–axis model with iron loss. (b) q–axis model with iron loss. ............44

Figure 3.4. Iron loss resistance values for varying speeds from conducted on the laboratory IPMSM....................................................................................................................45

Figure 3.5. Conduction losses as a function of varying inverter parameters. (a) Conduction losses varying with respect to $M_t$ and load current. (b) Conduction losses as a function of power factor angle and load current.........................................................47

Figure 3.6. Switching losses as a function of varying inverter parameters. (a) Switching losses varying with respect to $f_c$ and VDC. (b) Switching losses varying with respect to $f_c$ and load current.........................................................47

Figure 3.7. Phasor diagram of IPMSM in motoring mode .............................................48

Figure 3.8. Fundamental iron losses as a function of $\gamma$ and varying $I_m$ at switching frequency of 12 kHz and DC link voltage of 650 V. (a) Iron losses at 175 rpm. (b) Iron losses at 575 rpm.................................................................48

Figure 3.9. Fundamental copper and copper plus iron losses as a function of $\gamma$ and varying $I_m$ at switching frequency of 12 kHz and DC link voltage of 650 V. (a) Copper losses at 175 rpm. (b) Total fundamental losses at 575 rpm.................................50

Figure 3.10. Fundamental copper and iron losses as a function of DC link voltage at 575 rpm and 1000 rpm. (a) Constant torque region at 575 rpm and 70 Nm. (b) Flux –weakening region at 1,000 rpm and 40 Nm ..............................................................52

Figure 3.11. Comparison of harmonic iron losses at 450 V and 650 V and torques of 35 Nm and 70 Nm for varying speeds. (a) 275 rpm. (b) 575 rpm.................................54

Figure 3.12. Comparison of harmonic iron losses at varying switching frequencies at a loading condition of 575 rpm and 35 Nm.........................................................55
Figure 3.13. Comparison of harmonic iron losses for varying current angles at varying speed and load conditions. (a) \( I_m = 2\) A and varying speeds. (b) \( I_m = 14\) A and varying speeds. ........................................................................................................55

Figure 3.14. Comparison of magnet eddy current losses for varying switching frequencies and DC link voltage. (a) Varying switching frequency at 650 V. (b) Varying DC link voltage at 12 kHz. ........................................................................................................56

Figure 3.15. System losses calculated for varying current angles at 575 rpm and varying \( I_m \). .........................................................................................................................57

Figure 3.16. Input power, output power, losses and system efficiency calculated for varying current angles at 575 rpm and \( I_m = 2\) A. (a) Input and output power and total losses. (b) System efficiency. ........................................................................................................58

Figure 3.17. System level power variation at rated speed and torque. (a) Losses as a function of \( f_c \) at 650 V. (b) Losses as a function of VDC at 12 kHz. ...........................................................................59

Figure 3.18. System power loss variation as a function of VDC and carrier frequencies. (a) Power loss variation as a function of VDC for fixed switching frequencies. (b) Power loss variation as a function of \( f_c \) for fixed VDC. ........................................................................61

Figure 3.19. Input power, output power, losses and system efficiency calculated for varying current angles at 575 rpm and \( I_m = 15\) A. (a) Input and output power and total losses. (b) System efficiency. ........................................................................................................61

Figure 4.1. Equivalent circuit illustration of conventional LSIPMSM dynamic model. (a) Direct axis. (b) Quadrature axis. ........................................................................................................65

Figure 4.2. Experimental results for \( d \)– and \( q \)–axis DC current test (a) \( d \)–axis inductance as function of time. (b) \( q \)–axis inductance as a function of time. ..................................................68

Figure 4.3. Experimental results for \( d \)– and \( q \)–axis magnetization characteristics. (a) \( d \)–axis inductance as function of \( d \)–axis current. (b) \( q \)–axis inductance as function of \( q \)–axis current. ........................................................................................................68

Figure. 4.4. Flow chart of the IPSO algorithm developed for parameter determination. ..69

Figure 4.5. Particles updating in a circular behavior .................................................................69

Figure 4.6. Block diagram of experimental setup and optimization process.......................72

Figure 4.7 Experimental setup of the laboratory LSIPMSM .................................................72

Figure 4.8. Model validation for 20Hz, 85V supply. (a) Measured phase currents \( I_a, I_b \) and \( I_c \). (b) Calculated phase currents \( I_a, I_b \) and \( I_c \) using identified model .........................74
Figure 4.9. Model validation for 60 Hz, 240V supply. (a) Measured phase currents $I_a$, $I_b$ and $I_c$. (b) Calculated phase currents $I_a$, $I_b$ and $I_c$ using identified model without saturation (conventional) (c) Calculated phase currents $I_a$, $I_b$ and $I_c$ using identified model with improved model accounting for saturation (developed model). ........................................75

Figure 4.10. Calculated results for identified model for 60Hz, 240 V supply. (a) Torque and speed characteristics. (b) Calculated phase current $I_a$. Model 1: Conventional model without saturation, Model 2: Improved model incorporating saturation ...........................76

Figure 4.11. $d-$ and $q-$axis equivalent circuit model of IPMSM incorporating iron loss resistance. (a) $d-$axis model with iron loss. (b) $q-$axis model with iron loss. ...............78

Fig. 4.12. Iron loss resistance values for varying speeds from method 2 tests conducted on test IPMSM. ........................................................................................................79

Figure. 4.13. Structure of proposed two–step RLS estimation for multi–parameter estimation incorporating iron losses .................................................................83

Figure 4.14. $d-$axis inductance vs current estimation with and without iron losses through mathematical model for IPMSM and comparison with 2–D FEA. ......................85

Figure 4.15. $q-$axis inductance vs current estimation with and without iron losses through mathematical model for IPMSM and comparison with 2–D FEA. ......................85

Figure 4.16. Relative error from actual value in estimation of $d-$ and $q-$axis inductances with and without iron losses using mathematical model. ........................................86

Figure 4.17. $d-$axis inductance vs current estimation from mathematical model for IPMSM with and without iron losses and comparison with actual value. ......................87

Figure 4.18. $q-$axis inductance vs current estimation from mathematical model for IPMSM with and without iron losses and comparison with actual value. ......................87

Figure 4.19. Relative error of $L_{ds}$ from mathematically obtained values of $d-$axis inductance with and without iron losses at varying load torque conditions. ................87

Figure 4.20. Relative error of $L_{qs}$ from mathematically obtained values of $q-$axis inductance with and without iron losses at varying load torque conditions. ................88

Figure 4.21 Three–phase currents obtained from experimental identification process for $I_m$ =10 A and $\gamma= 20^\circ$ at a speed of 600 rpm. .................................................................89

Figure 4.22. Measured $I_{ds}$ and $I_{qs}$ currents for identification process at $I_m =10$ A and $\gamma= 20^\circ$ at a speed of 600 rpm. .................................................................89

Figure 4.23. Comparison of $L_{ds}$ and $L_{qs}$ obtained experimentally from stage 1 of identification at 600 rpm with and without iron losses for $I_m=10$ A and $\gamma= 20^\circ$. ..........90
Figure 4.24. Comparison of $\lambda_{PM}$ obtained from stage 2 of identification at 600 rpm with and without iron losses for $I_m=10$ A and $\gamma=20^\circ$.

Figure 4.25. Measurement of winding temperature using thermocouples attached in the windings at $I_m=10$ A and $\gamma=20^\circ$ at a speed of 600 rpm.

Figure 5.1. Flux linkage representation with respect to inductances and PM flux linkage depicting saturation and temperature variations effects.

Figure 5.2. Block diagram representing test method for determination of flux maps experimentally.

Figure 5.3. Procedure and measurements of current control used for flux linkage fitting.

Figure 5.4. Flux linkage maps obtained in the $d-$ and $q-$ axis for room temperature. (a) $d-$axis flux linkage. (b) $q-$axis flux linkage.

Figure 5.5. Permanent magnet temperature measured using thermal image camera at various operating conditions. (a) Room temperature. (b) Magnet temperature measured at 47$^\circ$C. (b) Magnet temperature measured at 60$^\circ$C.

Figure 5.6. Permanent magnet flux variation with temperature.

Figure 5.7. Stator temperature measurement and resistance variation with temperature (a) Stator temperature measurement using RTDs. (b) Stator resistance variation with temperature.

Figure 5.8. Procedure of deriving optimal current angle for a specific speed and peak current at room temperature, $T_1$.

Figure 5.9. $d-$ and $q-$axis equivalent circuit model of IPMSM incorporating iron loss resistance with respect to flux linkage. (a) $d-$axis model with iron loss. (b) $q-$axis model with iron loss.

Figure 5.10. Simulated values of current angle, $\gamma_{MEPA}$ with respect to peak current and speed.

Figure 5.11. Simulated values of efficiency variation with current angle showing maximum efficiency angles for varying speeds and $I_m=2$ A. (a) System efficiency. (b) Motor efficiency.

Figure 5.12. System efficiency corresponding to simulated values of current angle, $\gamma_{MEPA}$ in 3–D and 2–D forms. (a) Surface map of system efficiency. (b) System efficiency with respect to $I_m$ for various speeds.

Figure 5.13. Motor efficiency corresponding to simulated values of current angle, $\gamma_{MEPA,m}$ in 3–D and 2–D forms. (a) Surface map of motor efficiency. (b) Motor efficiency with respect to $I_m$ for various speeds.
Figure 5.14. Control diagram for implementation of the developed maximum efficiency method in laboratory IPMSM drive.................................................................110

Figure 5.15. Comparison of experimental efficiency and simulated efficiency for various $I_m$ and speeds at room temperature.................................................................112

Figure 5.16. Results from sweep test at 500 rpm and $I_m=10$ A. (a) Efficiency vs current angle. (b) Output torque vs current angle.................................................................113

Figure 5.17. Comparison of $\gamma_{MEPA, actual}$, $\gamma_{MEPA}$ and $\gamma_{MEPA, unsat}$ for varying $I_m$ at 575 rpm.114

Figure 5.18. Efficiencies at 25°C and 85°C with temperature compensation with respect to $\gamma$ and $I_m$. (a) 25°C. (b) 85°C.................................................................115

Figure 5.19. Efficiencies obtained experimentally at 85°C with and without temperature compensation as functions of speed at 14 A and 2 A.................................116

Figure 5.20. Efficiency comparison the developed method with MTPA for varying loading and speeds.................................................................117

Figure 5.21. Motor efficiency comparison of the developed method with MTPA for varying loads and speeds.................................................................117

Figure 6.1. Iron losses calculated from no–load tests at varying speeds and flux linkages to determine average hysteresis and eddy current loss coefficients..................122

Figure 6.2. Flowchart representing implementation of gradient descent optimization for maximizing efficiency in PMSM drives considering system losses................................................130

Figure 6.3. Implementation of the developed maximum efficiency angle detection...........131

Figure 6.4. Results from sweep test at 700 rpm and $I_m=10$ A. (a) Efficiency vs current angle. (b) Output torque vs current angle. (c) DC voltage measurement. (d) DC current measurement.................................................................134

Figure 6.5. Current angle iteration to determine maximum efficiency angle at 575 rpm and current change from 6 A to 10 A; initial $\gamma$ was $\gamma_{MTPA}$ of 24.9 degree.................135

Figure 6.6. Results of developed method to determine MEA during speed change from 575 rpm to 700 rpm at 6 A. (a) $d-$ and $q-$axis currents, DC current and DC voltage (secondary axis). (b) $d-$ and $q-$axis voltages. (c) Current angle iteration.................................................................136

Figure 6.7. Comparison of current angles at MTPA and MEA conditions for varying speeds and currents at 25°C operating temperature. (a) Current angles at 575 rpm and varying $I_m$. (b) Current angles at 700 rpm and varying $I_m$.................................137
Figure 6.8. Comparison of efficiency at MTPA and maximum efficiency current angle conditions for varying speeds and $I_m$ of 6 A and 10 A. (a) Efficiency at 575 rpm and varying $I_m$. (b) Efficiency at 700 rpm and varying $I_m$. .................................................................137

Figure 6.9. Comparison of system efficiency at MTPA and maximum efficiency current angle conditions for varying speeds and load currents. (a) System efficiency at maximum efficiency angle. (b) System efficiency at MTPA angle. .................................................................138

Figure 6.11. Three–phase PWM voltage measured at the motor terminals at 700 rpm for $I_m=10$ A and $\gamma=36.5^\circ$ and corresponding spectrum. (a) Phase voltage measurement of PWM waveform. (b) Harmonic spectrum of PWM voltage. .................................................................140

Fig. 6.12. System efficiency obtained using the developed method at 25°C and 65°C for varying $I_m$ and speed of 700 rpm. ..................................................................................................................140
NOMENCLATURE

A list of primary symbols is given here; there are more symbols used in this thesis, which
have been defined locally.

\( T_e \quad \) Electromagnetic torque

\( i_d, i_q \quad d-\text{ and } q-\text{axis current} \)

\( v_d, v_q \quad d-\text{ and } q-\text{axis voltage} \)

\( \lambda_d, \lambda_q \quad d-\text{axis flux linkage} \)

\( i_{md}, i_{mq} \quad d-\text{ and } q-\text{axis magnetizing currents} \)

\( R_i \quad \) Iron loss resistance

\( \lambda_{PM} \quad \) Permanent magnet flux linkage

\( L_d, L_q \quad d-\text{ and } q-\text{axis inductances} \)

\( \omega_s \quad \) Electrical angular speed

\( \omega_c \quad \) Carrier angular speed

\( I_m \quad \) Peak current

\( \phi_0 \quad \) Initial voltage vector phase

\( \theta \quad \) Electrical rotor position

\( \delta \quad \) Torque angle

\( P \quad \) Number of machine poles

\( J \quad \) Combined moment of inertia

\( R_s \quad \) Stator resistance

\( R_0 \quad \) Stator resistance at room temperature, \( T_i \)

\( \alpha \quad \) Copper thermal resistive coefficient

\( \beta \quad \) Magnet thermal resistive coefficient
\( f_s, f_c \)  Fundamental frequency, Carrier/switching frequency

\( F_s, F_r \)  Stator, Rotor MMF

\( F_{sh} \)  Harmonic Stator MMF

\( \varphi_s \)  Initial phase of winding current

\( \varphi_h \)  Initial phase of \( h^{th} \) winding current harmonic

\( \alpha_p \)  Pole arc coefficient

\( B_{hg} \)  Harmonic airgap flux density

\( \vartheta_s \)  Angular coordinate in the stator

\( \vartheta_r \)  Angular coordinate in the rotor

\( \mu_r \)  PM relative permeability

\( H_c \)  PM coercive force

\( h_m \)  Length of PM

\( g \)  Length of airgap

\( t_b, l_b \)  Thickness, length of the flux barrier

\( \vartheta_b \)  Angle of the flux barrier

\( k_t, k_y \)  Teeth width and yoke height coefficients

\( \gamma_{\text{actual}} \)  Current angle for actual maximum system efficiency

\( \gamma_{\text{actual,m}} \)  Current angle for actual maximum motor efficiency

\( A_t, A_y \)  Teeth and yoke areas

\( w_t, w_y \)  Teeth and yoke widths

\( \tau_s, \tau_p \)  Tooth and pole pitch

\( B_{htm} \)  Teeth flux density with harmonics

\( B_{hym} \)  Yoke flux density with harmonics
LIST OF ABBREVIATIONS

A list of key abbreviations is given here; there are more abbreviations used in this thesis, which have been defined locally.

EV Electric vehicle
FEA Finite element analysis
FOC Field oriented control
IPMSM Interior permanent magnet synchronous motor
IGBT Insulated-gate bipolar transistor
IPSO Improved particle swarm optimization
LSIPMSM Line start interior permanent magnet synchronous motor
MCM Magnetic circuit modeling
MTPA Maximum torque per ampere
MMF Magnetomotive force
MEA Maximum efficiency angle
MEPA Maximum efficiency per ampere
PM Permanent magnet
PWM Pulse–width modulation
RLS Recursive least square
RTD Resistance temperature detectors
GDA Gradient descent algorithm
CHAPTER 1
INTRODUCTION

1.1 Overview of Electric Motors and Drives in Traction Applications

Electric machines and drives used in electric vehicles (EVs) have been receiving increasing research and development interest in recent years due to environmental concerns and emphasis on global energy savings. Advancements in the areas of electric motor and drive design and control strategies have been in demand owing to the stringent requirements for performance and efficiency in automotive systems, more–electric aircraft and ships. Using new materials, innovative topologies and control strategies, it is possible to improve the efficiency of the electric machine and the power electronic components existing in the electric vehicle [1] – [5]. Figure 1.1 shows the block diagram of a commercial e–motor drive system used in Toyota Prius [2].

Any energy savings in the electric motor and the power electronic converter of the EV helps, which operates as an inverter during motoring operation, in obtaining longer distances per charge in the EV owing to lower energy consumption from the battery [3]– [5]. Hence, through advancements in electric motor testing and control methodologies, this
dissertation considers system–level efficiency improvement as the aim to extend the driving range and reduce the overall operational costs of the electrified vehicle.

The converter losses are composed of semiconductor conduction losses and switching losses, and are dependent on the converter topology, switching device characteristics and switching frequency of the devices [2]. Apart from these loss components, the power electronic converter also induces additional losses in the electrical machine as it produces a switched output voltage waveform with increased harmonic content. This leads to increased losses, additional heating and reduced lifetime of the machine. The electric motor component produces the highest electric losses in a motor and drive system consisting of the motor and the inverter [1]. The percentage of losses depends on the type of motor used. Proper selection of electric machine type is based on key features such as the energy source in the vehicle, space and vehicle dynamics, efficiency, reliability, cost, and the major operating requirements of the machine. The major operating requirements of the traction motor include a wide speed range, impulsive response, high efficiency over a wide torque and speed, high torque at low speeds, fault tolerance, and high–power density.

Among the major automakers, there is no general consensus as to the type of electric machine best suited for vehicles, but induction machines (IM) and permanent magnet synchronous machines (PMSM) are the two types currently used in EVs and are expected to continue to dominate the market [1], [2]. The PMSM machines have higher efficiency, torque density, and heat dissipation capability than their IM counterparts and are widely used in today’s EVs because of their superior performance over the induction machines. Out of the various configurations in PMSM such as the surface PMSM (SPMSM), interior PMSM (IPMSM) and inset PMSMs, the IPMSMs are widely used in commercial EVs and
continue to be the preferred choice compared to the other types [2]–[6]. Some of the commercial EVs/ Hybrid EVs using IPMSMs include Chevy Volt, Mitsubishi i–MiEV, Honda Accord, Nissan Leaf, Toyota Camry, Ford Focus, Toyota Prius, Lexus, etc. [2]. The motor losses in an IPMSM include mechanical losses, copper losses in the windings, iron/core losses in stator laminations, and magnet losses.

The efficiency map of the power components, that is the efficiency of the motor and inverter as a function of torque and speed, which determines the energy losses and consumption for vehicles, and the peak power characteristics, are important factors for high–performance demands [2]. Figure 1.2 shows efficiency maps of electric motor, inverter and system levels in 2014 Honda Accord [1]. It can be seen that the motor peak efficiency is close to 95%, inverter is 99% and the system efficiency, which is the combined efficiencies of the motor and inverter is 93%. It is to be noted that the component efficiencies as well as the system–level efficiencies are lower than 90% in certain operating conditions. The motor efficiency is also lower than 92% in high–speed, low–torque regions as well as the low–speed, high–torque regions. These operating points are frequent in an urban drive cycle and in some cases, highway driving cycle as well. Hence, it is important to consider design or control techniques that can improve the efficiency of the system over frequent operating points in a drive–cycle rather than achieving maximum efficiency at certain operating points for significant improvements in battery energy consumption.

Many control algorithms aimed towards the reduction of the electrical loss components during part load operation and increase in drive–cycle efficiency have been developed and reported in the literature. In the next section, current research areas and factors influencing system–level efficiency improvement and motivations of this dissertation are highlighted.
Figure 1.2. Efficiency maps for 2014 Honda Accord HEV at 300 V DC voltage. (a) Motor Efficiency. (b) Inverter Efficiency. (c) System Efficiency.
1.2 Review of Factors Influencing System Level Efficiency and Performance

Improvements in PM Based Electric Motor and Drive System

This research focuses on system level efficiency and performance improvements in PM based electric motor and drives. There is significant research on inverter topology and switch selection, pulse–width modulation type selection, etc. to improve the inverter efficiency that also affects the motor performance [5]. However, the scope of this dissertation is limited to improving the system level efficiency through improved motor modeling, control and testing techniques only.

In the study of PMSM motor drive efficiency improvement, it is important to consider the interaction of motor and drive and the behavior of one component with respect to the other. The major factors influencing control algorithms for improved efficiency in traction PM motor and drive can be summarized into the following:

1) Motor–drive interaction and accuracy of loss models
2) Parameter variations under varying operating conditions
3) Control algorithm and methodology of implementation

1.2.1 Motor–Drive Interaction and Accuracy of Loss Models

The electromagnetic field in the motor includes harmonics such as slot harmonics and carrier harmonics. They cause considerable losses in the stator and rotor. In case of stator laminations and rare earth magnets, the carrier harmonics produced by the pulse–width modulated (PWM) inverter may cause increase in stator core losses as well as magnet harmonic eddy–current losses, depending on the conductivity of the material and the impedance of the harmonic circuit. The mathematical analysis of losses in inverter–fed electrical machines should consider the space and time harmonics that distort the flux
density significantly and hence increase the losses and degrade performance [7]– [9]. The effect of PWM on losses and noise in electric motors have been widely studied in the literature [8]– [11]. It has been studied previously that eddy current losses are affected predominantly by PWM, especially in low-speed, low-torque regions of IPMSM [7], [9]. The main contribution factor is the time harmonics in both low frequency order as well as high frequency carrier order [9]. Another loss factor is space harmonics in the form of stator and rotor slotting [10]. The PWM based eddy current losses occur in stator and magnet [9]. Some of the applications and methodology considering PWM input in stator loss models include induction motors [11], finite element analysis (FEA) of IPMSM [9], [10] and extension of research conducted on steel specimens energized with PWM supply [11]. Regarding analytical techniques, winding function and semi-analytical method combined with simplified FEA have been developed to consider space harmonics and subsequently, effective motor design solutions have been proposed [14], but not including PWM carrier harmonics. Thus, there is a need of accurate analytical harmonic iron loss modeling in the stator core to study the influence of PWM harmonics on the motor and develop control methodologies to reduce these harmonics for wide operating regions.

1.2.2 Parameter Variations Under Varying Operating Conditions

The efficient performance of variable speed motor drives depends on the controller settings, which in turn depend on the accurate knowledge of the machine parameters that establish the correlation between the input excitation and the resulting torque [3]. The precise knowledge of machine parameters is beneficial not only for accurate control, but also for obtaining the best performance from the motor at various operating points and for fault tolerant control operation. In PMSMs, the resistive and magnetic properties vary with
operating conditions and temperature. Many techniques have been developed in the literature to identify varying parameters of IPMSMs as well as SPMSMs. The methods for identification performed experimentally can be classified as offsite experimental methods [15], on–site and off–line methods [16], [17] and on–line techniques [18]– [21]. Off–site experimental methods are either simple and include standard tests such as no–load, blocked rotor tests, etc. or require special experimental setup and are widely used for parameter determination in many types of electrical machines. However, the major drawback is the poor representation of real operating conditions and non–linearities associated with the machine. On–site and off–line test methods commonly make use of measurements from experiments performed on the motor connected to the drive in various operating conditions prior to actual operation and a search algorithm or a constrained optimization algorithm is used to identify the parameters off–line. In such methods, the estimation and updating of machine parameter information is not possible while it is in continuous operation, except from 2–D or 3–D look–up tables that are cumbersome if all conditions need to be incorporated. On–line parameter estimation methods are very popular and are used in sensor–less and adaptive control of PMSM. Various algorithms have been used to identify on–line, the inductances, stator resistance and PM flux linkage of the machine. Majority of identification problems are focused on identifying one or two parameters by keeping the other parameter as a constant. For instance, [19] identified inductances and winding resistance by keeping the PM flux linkage constant. Multi–parameter estimation is the estimation of all or many parameters simultaneously and some examples from literature include [20]– [22]. In [20], a recursive least square (RLS) method was used to simultaneously estimate all four parameters by using fast and slow segments without
dependence on off–line measurement for any of the parameters. In [21], additional measurements such as power and torque were used for the on–line multi–parameter identification problem. In [22], the problem of rank deficiency in identifying all parameters was performed by $d$–axis current injection and solving two sets of simplified PMSM equations by using Adaline neural network theory.

The stator resistance and PM flux predominantly depend on temperature. The stator winding temperature is dependent on operating conditions including load, operating frequency, cooling conditions, etc. On the other hand, the magnetic parameters, $d$– and $q$–axis inductances largely depend on the operating flux level of the machine. Most of the on–line multi–parameter identification techniques provided in the literature review neglect the impact of iron losses on the estimation process. Majority of estimation including iron losses in literature have been performed in induction motors [23] and a few in PMSMs [24]–[27]. The impact of iron losses has been observed to be significant in the estimation of stator resistance, magnitude changes in the back EMF and output torque calculations [25]. In sensor–less control, there is significant angle estimation error in case of models neglecting iron loss factor [26]. In PMSM, most of the work performed including iron loss is either off–line or neglects the simultaneous estimation of other parameters [25]–[28]. The iron loss factor is generally ignored in low speed operations, even in case of high torque conditions. However, the effect of iron loss could be significant in high torque conditions [18]. Hence, it is important to study the parameter variations accurately and also consider the iron loss factor at high torque conditions, especially in EV application where torque/ control performance is vital.
1.2.3 Control Algorithm and Methodology of Implementation

Of the various control methods, development of optimal control strategies to maximize efficiency have been of significant interest in the literature [32]– [48]. Recent research topics in control of PMSM to improve efficiency include improved maximum torque per ampere (MTPA) to reduce copper losses [32]– [39] and loss minimization (LM) through model based or search controller–based techniques [42]– [49]. The main difference between MTPA and other loss minimization strategies is that the former can only minimize copper losses whereas the latter can reduce the iron losses as well. In general, the necessary features to be addressed in developing control methods for efficiency improvement can be summarized as: (1) Applicability of the developed method in real conditions when parameter variations in motor is inevitable; (2) Scope of power loss reduction at component and system level; and (3) Ease and methodology of control implementation.

In PMSM control, MTPA is one of the most widely used techniques where the objective is to obtain an optimal current angle that consumes minimum stator current for a required output torque. In terms of applicability in real conditions, MTPA angle derivation has been performed using parameter–based approaches [32]– [34] and on–line techniques such as signal injection and DC power measurement which are robust against parameter variations [35]– [38]. In model–based derivations, inductance variations due to magnetic saturation and stator resistance and permanent magnet (PM) flux variation due to temperature are vital factors to be addressed. In [32], MTPA angle was calculated using look–up tables of parameters for varying operating condition. However, it is tough to include parameter variations considering saturation, cross–saturation and temperature simultaneously through look up tables. In [33], and [34], on–line parameter estimation techniques were used to
calculate the MTPA angle in real time. However, these methods can suffer from rank deficiency issues. Some of the solutions suggested use extra invasive measurements or methods such as torque sensor or temperature sensors [37] or current and voltage signal injection [36], [38] that could increase machine losses. In [39], DC power measurement was considered to compute optimal voltage. However, the method depends on motor model and is not entirely robust against parameter variations. In general, even though robustness against parameter variations can be considered, MTPA method does not provide a true optimal efficiency point as only copper losses are minimized. The motor iron and harmonic losses are ignored, even though they are significant at higher speeds [40], [41].

In loss minimization techniques towards better efficiency considering iron losses, the main difference from MTPA is the improvement in scope of optimal current distribution derivation in $d$– and $q$–axis by considering iron loss, stray loss, etc. Typically, loss minimization in PMSM can be classified into model based [42]– [44], search based [47], [48] and hybrid methodologies [49]. The model–based methods derive LM using motor model where the $d$– and $q$–axis currents are derived by solving constrained optimization problem using numerical methods or approximate analytical solutions [42]– [45]. The optimal values can be generated as look–up tables to be used in the control [42] or solved on–line [44]. The search–based methods aim to drive the control variable towards minimum power loss, regardless of the motor ratings or parameters [46]– [48]. However, search methods are slower than model–based techniques and are sensitive to current and voltage harmonics and cause torque ripple due to perturbations. Hybrid loss minimization methods were developed to comprehend the advantages of both model– and search–based methods [49]. Even though methods of improving the overall efficiency of the motor drive
system through control variables have been studied [50]– [52], a loss minimization method for PMSM that is robust and improves the scope of loss reduction by considering losses from a system level is not yet fully developed.

1.3 Motivation of the Thesis

Given the drawbacks in the methods developed in literature regarding the three major factors affecting control algorithms towards motor and drive efficiency improvement, this thesis contributes to novel modeling, testing and control methodologies aimed towards the energy efficiency improvement of a PMSM motor- and drive system. The approaches are: (1) Accurate analytical modeling and testing of loss components in PMSM considering inverter harmonics; (2) Easy–to–implement, accurate parameter determination techniques to understand variations in motor parameters due to saturation, cross–saturation and temperature; and, (3) Control methodologies to improve system level efficiency considering improved loss models and parameter variations.

The overall objectives are summarized as follows:

1. Understand the sources of losses in PM machines through comprehensive study using analytical models, numerical simulations and experimental tests.

2. Derive improved analytical model for air–gap flux density considering various sources of harmonics such as time and space caused by PWM–fed inverter and motor design parameters for improved loss models.

3. Identify control parameters affecting various losses in an insulated-gate bipolar transistor (IGBT)–fed PMSM and perform a study to understand loss behavior with respect to varying control parameter and propose the chosen control variable towards improved efficiency.
4. Develop methods of identifying varying parameters in IPMSM such as inductances due to saturation and PM flux due to temperature and understand their effects on system efficiency.

5. Validate developed loss models and parameter determination methods in laboratory PMSM through simulations and experimental investigations for various operating conditions.

6. Propose efficiency improvement method considering the improved loss models and variation of parameters under wide speed and torque regions.

7. Explore efficiency improvement from a system–level with parameter independence.

8. Analyze improvements in efficiency considering the developed methodologies in laboratory PMSM using numerical and analytical simulations and experiments.

1.4 Dissertation Outline and Research Contributions towards Improved System Efficiency in PMSM Drives

This dissertation proposes novel modeling, testing and control methodologies towards global improvement of efficiency in permanent magnet synchronous motor drives from a system level. This dissertation presents 5 chapters excluding this introductory chapter and the conclusion chapter that present the research conducted and novel research contributions made while working towards the overall objectives. The chapter outlines, and major contributions of this work are highlighted as follows:

Chapter 2 proposes the eddy current loss behavior due to time and space harmonics through a novel analytical modeling. An improved winding function theory incorporating armature reaction fields due to fundamental and harmonic stator magnetic fields has been used to
analyze the air–gap flux density including harmonics. Further study on stator and rotor eddy current losses has been performed. It was concluded that the PWM harmonics cause significant increase in losses and it is imperative to consider the dependency and controllability of the harmonic losses and methods of reduction through control techniques, especially in stator harmonic losses.

Chapter 3 proposes a detailed investigation into behavior of controllable losses in PMSM as functions of control variables such as current angle, DC link voltage, and switching frequency. The loss model proposed in Chapter 2, in addition to inverter loss models and fundamental loss models of the motor, have been used in a field–oriented control (FOC) based simulation to study the system level losses. The analysis suggested that the change in current angle can affect the overall system losses and that for every operating point there exists an optimal distribution of current that lead to maximum system efficiency.

Chapter 3 concludes that the optimal current value selection helps in improving efficiency of the system. However, it is vital to understand parameter variations with loading and speed in order to model the motor and drive efficiency close to actual conditions and further study system efficiency. Chapter 4 proposes methods of testing equivalent circuit parameters through off–line and on–line methods to study parameter variation due to saturation and temperature variations. Easy to implement off–line parameter determination using improved particle swarm optimization (IPSO) and on–line multi–parameter estimation through recursive least square (RLS) method have been developed. The methods suggested significant changes in motor parameters in the laboratory test motors. The developed on–line method was also used to study the influence of iron loss on output torque production.
Given the understanding of system efficiency behavior in a PMSM drive, an off–line loss minimization procedure was developed in Chapter 5 using a search–based approach towards minimization of system losses. The effects of parameter variation due to saturation and temperature have been considered using flux linkage maps to avoid the segregation of saturation and temperature effects in $d$–axis flux linkage. The losses in the motor were calculated considering these parameter variations. Improved optimization procedure to calculate optimal $d$– and $q$– axis currents considering the flux linkage–based loss models enables considering the effects of saturation, cross–saturation and temperature variation in stator and rotor. The inverter losses were included in deriving the optimal current angle. The comparison with conventional MTPA is provided along with experimental validations in laboratory test motor.

Chapter 6 proposes an on–line loss minimization procedure to improve the system efficiency considering DC power measurement and helps to simplify the consideration of parameter variations in real–time conditions. Firstly, the motor and inverter loss models were derived in such a way that the calculation can be performed using terminal measurements in a PMSM drive. Consequently, a gradient descent search–based algorithm is used to calculate the optimal current angle corresponding to maximum system efficiency from the input DC power measurement and output power based on the loss models. The developed method was compared against conventional MTPA and validated experimentally.

Chapter 7 summarizes the developed methods and their contributions towards the PMSM system efficiency improvement as well as proposes some future work to be considered as a follow–up of this dissertation
The test motor used for all analysis in this thesis is a 4.25 kW laboratory IPMSM except the off–line parameter estimation technique wherein a line–start IPMSM (LSPMSM) has been used. However, the knowledge from the tests have been used for the IPMSM study. All contributions, including loss models, parameter determination, control algorithms and corresponding experimental validations have been proposed keeping the IPMSM as a test case. It is to be noted that many of the contributions of this dissertation, especially the control algorithm improvements, can be easily applied to SPMSMs by modifying the equivalent circuit parameters accordingly and is expected to provide improved energy efficiency compared to conventional control techniques in SPMSMs also.
CHAPTER 2

COMPREHENSIVE ANALYTICAL MODELS TOWARDS STUDY OF HARMONIC MOTOR LOSS BEHAVIOUR IN A PWM–FED PERMANENT MAGNET MACHINE

2.1 Introduction

In this chapter, loss models are developed to study the behaviour of losses with respect to operating conditions. The behaviour of fundamental and harmonic losses in a PWM–fed IPMSM are studied. The time harmonics from inverter are significant in IPMSM and can cause increase in stator core eddy current losses. This chapter aims at modeling the harmonic losses accurately in an IPMSM, and subsequently study methods of reduction through control techniques. The outcomes of the loss modeling and analysis are used in Chapter 3 to study the behaviour of motor losses to varying control variables that can optimize the system efficiency. The proposed loss models include time harmonics from the PWM inverter and space harmonics from the motor design to analyze the fundamental and harmonic losses in the stator and rotor of the motor.

To study the behaviour of the aforementioned losses, a 4.25 kW laboratory IPMSM is used for investigations. The design parameters of the motor are given in Table 2.1.

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator outer, inner diameter</td>
<td>220.137 mm, 134.137 mm</td>
</tr>
<tr>
<td>Rotor outer, inner diameter</td>
<td>133.137, 85 mm</td>
</tr>
<tr>
<td>Number of poles ($p$)</td>
<td>8</td>
</tr>
<tr>
<td>Slot number, conductor per slot</td>
<td>48, 28</td>
</tr>
<tr>
<td>Pole arc/ pole pitch</td>
<td>0.82</td>
</tr>
<tr>
<td>Tooth width, slot pitch</td>
<td>4.96 mm, 8.78 mm</td>
</tr>
<tr>
<td>Magnet width, thickness</td>
<td>36 mm, 6 mm</td>
</tr>
<tr>
<td>Magnet electrical conductivity</td>
<td>62500 S/m</td>
</tr>
</tbody>
</table>

The ratings and equivalent circuit parameters of the test IPMSM are given in Table 2.2.
TABLE 2.2 PARAMETERS OF TEST IPMSM USED FOR VALIDATIONS

<table>
<thead>
<tr>
<th>Speed (rpm)</th>
<th>Power (kW)</th>
<th>Torque (Nm)</th>
<th>Stator Resistance (Ω)</th>
<th>PM flux (V.s/rad)</th>
<th>$L_{ds}$ (mH)</th>
<th>$L_{qs}$ (mH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>575</td>
<td>4.75</td>
<td>70</td>
<td>1 at 25°C</td>
<td>0.67</td>
<td>27.1 $(i_d = -7A,\ i_q = 5.1A)$</td>
<td>82.5 $(i_d = 7A,\ i_q = 5.1A)$</td>
</tr>
</tbody>
</table>

Figure 2.1. Three–phase IPMSM designed and prototyped in–house with 48 slots, 8 poles distributed winding, and NdFeB35 magnets in the rotor. (a) Cross–section of the IPMSM. (b) Flux density distribution at rated condition. (c) Experimental setup with the laboratory prototyped motor.
The validations of the models developed in this chapter have been performed using 2–D electromagnetic model/ finite element analysis (FEA) and experimental investigations. Figure 2.1(a) shows the 2–D FEA cross–sectional design of the motor, 2.1(b) shows the air–gap flux density distribution at rated condition and 2.1(c) shows the experimental setup used for the investigations.

The current control of the test motor is conducted using Opal–RT real time controller and IGBT inverter. In order to validate the accuracy of the 2–D FEA, back EMF tests were conducted at various speeds and similar conditions were provided in 2–D FEA. Figure 2.2 shows the experimental and 2–D FEA computed values of no–load back EMF at 1,000 rpm. The test setup consists of the IPMSM under test, which is in current controlled mode and is connected to a speed dynamometer. Figure 2.3 shows the stator and rotor of the motor under test.

![Figure 2.2](image)

**Figure 2.2.** No–load back–EMF experimental validation of accuracy of 2–D electromagnetic analysis model used for validating developed model at 1,000 rpm.

![Figure 2.3](image)

**Figure 2.3.** Stator and rotor of the prototype IPMSM under test.
The following sections in this chapter explain the loss model derivations, experimental investigations in the test motor and subsequent conclusions.

### 2.2 Sources of Harmonic Motor Losses

The electromagnetic field in a PWM–voltage fed motor includes harmonics such as slot harmonics from the motor design and carrier harmonics from the PWM inverter. In a sinusoidal, regular sampled PWM, the harmonic components of output voltage manifest at frequencies around multiples of carrier frequency, called as side–band harmonics and additional low frequency harmonic components called base–band harmonics [53]. An analytical expression for phase voltage of a VSI modulated by asymmetrical regular sampled sine triangle PWM can be written as in (2.1).

\[
V_{i,h}(t) = \frac{2V_{dc}}{\pi} \sum_{x=1}^{\infty} \frac{1}{x} J_0 \left( \frac{x \pi M_i}{2} \right) \cos(x \omega_s t) \sin \left( \frac{x \pi}{2} \right) + \\
\frac{2V_{dc}}{\pi} \sum_{x=1}^{\infty} \sum_{y=-\infty}^{\infty} \left( \frac{1}{x} J_y \left( \frac{x \pi M_i}{2} \right) \cos \left( (x \omega_c t) + y \left( \omega_c t + \theta_i \right) \right) \right) \times \sin \left( (x + y) \frac{\pi}{2} \right)
\]

(2.1)

where, \(V_{dc}\) is DC link voltage, \(x\) and \(y\) indices of sideband group and position of the sideband group respectively, \(J_y\) is the Bessel function of order \(y\), \(M_i\) is the modulation index, \(\omega_s\) is fundamental frequency, \(\omega_c\) is the carrier angular frequency and \(\theta_i\) is the phase shift for time \(t\). The harmonics have orders of \(m_1 f_c\) for carrier harmonics where \(m_1=1, 2, 3\ldots\) etc., and \(m_1 f_c + m_2 f_s\), for side–band harmonics where \(m_2=\pm1, \pm2, \pm3\), etc. [9]. The voltage harmonics lead to current harmonics corresponding to the motor impedance. These harmonics lead to eddy current losses in the stator and rotor. A sample Fourier spectrum showing the magnitudes of voltage components for frequency modulation of 50 is shown in Figure 2.4.
The design–based space harmonics that lead to harmonic losses are winding harmonics, slot harmonics and magnet harmonics due to the rotor structure [54]. In this work, the stator winding harmonics, rotor harmonics and carrier harmonics are taken into account to study the harmonic eddy current losses in the stator and rotor of an IPMSM through analytical derivation of air–gap flux density. The slot harmonic contribution to losses is minor compared to the other harmonics [9], [54], [55] and not controllable using control techniques and hence are neglected.

The following section elaborates on the derivation of air–gap flux density based on the reaction fields considering: (i) fundamental magnetomotive force (MMF), and (ii) sideband armature reaction field stator MMF with space harmonics.

2.3 Analytical Magnetic Flux Density Derivation in PMSM

An analytical method is proposed to study the effect of PWM parameters on eddy current losses in the stator and rotor using rotor MMF obtained from a magnetic circuit model (MCM) as a function of armature reaction stator MMF. Firstly, the harmonic currents resulting from the PWM inverter voltage have been derived using analytical and 2–D FEA models by considering appropriate impedance matrix. Furthermore, the harmonic stator
and rotor MMFs have been analytically derived for the corresponding current harmonics interacting with the space harmonics for specific operating conditions. Finally, harmonic flux density and the resulting stator iron losses due to carrier harmonics have been calculated. The rotor eddy current losses have been analyzed using a PWM harmonic loss factor to understand the impact of PWM harmonics on the magnet losses. The reference frame used to derive the MMFs and finally air–gap flux density is given in Figure 2.5.

![Figure 2.5. Representation of the rotor and reference axes in the IPMSM analyzed.](image)

2.3.1 **Stator MMF Computation Including Carrier Harmonic Ripple Current as a Function of PWM Parameters**

The phase voltage harmonics manifest in the form of current harmonics to create harmonic armature reaction field that increase the stator eddy current losses. The current harmonics from sideband harmonics can be expressed as given in (2.2) where $I_{mh}$ is the magnitude of harmonic current $h$, $m_1$ and $m_2$ express the order of the side–band harmonic as $m_1\omega_c \pm m_2\omega_s$ as explained in 2.2.

$$i_{ah}(t) = I_{mh} \cos \left( \left( m_1\omega_c \pm m_2\omega_s \right) t \mp \varphi_h \right)$$

$$i_{bh}(t) = I_{mh} \cos \left( \left( m_1\omega_c \pm m_2\omega_s \right) t \mp \frac{2m_2\pi}{3} \pm \varphi_h \right)$$

$$i_{ch}(t) = I_{mh} \cos \left( \left( m_1\omega_c \pm m_2\omega_s \right) t \mp \frac{4m_2\pi}{3} \pm \varphi_h \right)$$

(2.2)
The predominant side–band harmonics for phase currents in sine PWM fed IPMSM considered in this study are of first order, $\omega_c \pm 2\omega_s$, $\omega_c \pm 4\omega_s$, and second order, $2\omega_c \pm \omega_s$, $2\omega_c \pm 5\omega_s$ and $2\omega_c \pm 7\omega_s$. The currents are derived as a function of $V_{dc}$, torque angle, $\delta$, and $d$–$q$ axis inductances, $L_d$ and $L_q$ considering saturation [55] and coefficients of harmonic currents in (2.3) using (2.4) [56]. The analytical expressions of coefficients $C_1$ to $C_7$ using Bessel functions and modulation index, $M_i$ are given in Appendix A.

\[
i_s(\omega_c \pm 2\omega_s) = \frac{V_{dc}\sqrt{M^2_2 + N^2_1 + 2M_2N_1 \cos(2\varphi)}}{4(\omega_c \pm 3\omega_s)}
\]

\[
i_s(\omega_c \pm 4\omega_s) = \frac{V_{dc}\sqrt{M^2_1 + N^2_2 + 2M_1N_2 \cos(2\varphi)}}{4(\omega_c \pm 3\omega_s)}
\]

\[
i_s(2\omega_c \pm \omega_s) = \frac{V_{dc}C_1\cos^2(\varphi_0)L^2_q + \sin^2(\varphi_0)L^2_d}{4\omega_c\sqrt{L^2_d L^2_q}}
\]

\[
i_s(2\omega_c \pm 5\omega_s) = \frac{V_{dc}\sqrt{M^2_4 + N^2_3 + 2M_4N_3 \cos(2\varphi)}}{8(\omega_c \pm 3\omega_s)}
\]

\[
i_s(2\omega_c \pm 7\omega_s) = \frac{V_{dc}\sqrt{M^2_3 + N^2_4 + 2M_3N_4 \cos(2\varphi)}}{8(\omega_c \pm 3\omega_s)}
\]

\[
M_1 = \frac{C_2}{L_d} - \frac{C_2}{L_q}; M_2 = \frac{C_2}{L_d} + \frac{C_2}{L_q}; M_3 = \frac{C_5}{L_d} - \frac{C_5}{L_q}; M_4 = \frac{C_5}{L_d} + \frac{C_5}{L_q}
\]

\[
N_1 = \frac{C_4}{L_d} - \frac{C_4}{L_q}; N_2 = \frac{C_4}{L_d} + \frac{C_4}{L_q}; N_3 = \frac{C_7}{L_d} - \frac{C_7}{L_q}; N_4 = \frac{C_7}{L_d} + \frac{C_7}{L_q}
\]

The initial voltage vector phase, $\varphi_0$ changes with loading condition according to (2.5) where $\delta$ is the torque angle, $v_d$ and $v_q$ are $d$– and $q$–axis voltages.

\[
\varphi_0 = \pi + \arctan \left| \frac{v_d}{v_q} \right| ; \quad \arctan \left| \frac{v_d}{v_q} \right| = \delta
\]

The stator MMF of the IPMSM can be represented as in (2.6), where $k$ is number of phases, $N_\Phi$ is the winding function and $i_\Phi$ is the stator current.
The fundamental armature reaction field is caused by the fundamental current interacting with the winding and rotor space harmonics of order \( m \). The stator slot harmonics are neglected. The total stator MMF considering fundamental current, \( I_1 \) and winding function with space harmonics, \( m \) is derived as the summation of stator MMFs in each phase using (2.7) where the winding function with space harmonics and instantaneous currents are given in (2.8). The net phase stator MMF with fundamental current is defined in (2.9).

\[
F_{s1}(\theta_s, \omega t) = F_{a,s}(\theta_s, \omega t) + F_{b,s}(\theta_s, \omega t) + F_{c,s}(\theta_s, \omega t)
\]

(2.7)

\[
N_X(\theta_s) = \sum_{m=1,3,5...} N_{X,m} \left[ \cos m(\theta_s + Y) \right]
\]

(2.8)

where \( X \) represents the phases A–C, \( Y \) is 0, \(-2\pi/3\), and \(2\pi/3\), for phase A–C respectively and \( \gamma_i \) is the initial current angle measured between phase A and d–axis.

\[
F_{X,s}(\theta_s, \omega t) = \sum_{m} N_{X,m} \left( \cos m(\theta_s + Y) \right) I_X \cos(\omega t + Y + \gamma_i)
\]

(2.9)

The MMF given in (2.9) will be used towards the derivation of iron losses due to fundamental current. The other armature reaction field studied using the model is the side-band armature reaction field and the corresponding harmonic losses. Using (2.2), the stator MMF, \( F_{s,h} \) with side–band harmonics from (2.3) that can be used to calculate the harmonic reaction field and the net stator MMF including fundamental MMF, \( F_s \) are written as (2.10).

\[
F_{s,h}(\theta_s, \omega t) = \sum_{m} \sum_{h} F_{sh} \cos \left( mp\theta_s \pm (m_1\omega_c \pm m_2\omega_s) t \pm \phi_h \right)
\]

(2.10)
2.3.2  Rotor MMF as a Function of Armature Reaction Field from Stator Excitation and Air–gap Magnetic Flux Density Derivation

The analysis of airgap magnetic flux density and subsequent analysis of the machine losses have been performed based on the derivation of rotor MMF in [57], where stator iron losses were computed due to rotor geometry. The rotor MMF, $F_r$ induced by the stator MMF, $F_{s,h}$ can be used to analyze the magnet losses caused by harmonics in the stator excitation. The rotor MMF is fed by two sources, the PMs and the stator excitation. The dependency on PM is represented by $ST$, and a staircase function represented in Figure 2.6, rotating synchronously with the rotor and the dependence on stator excitation is represented as a reaction field rotor MMF, $U_r$.

![Figure 2.6. Staircase function representing rotor MMF due to PM for one flux barrier per pole.](image)

The number of stairs in $ST$ depends on the number of flux barriers per pole. The rotor MMF, $F_{rs}$ caused by the armature reaction field voltage drop, $U_r$ is given in (2.11), where $\alpha_p$ is the pole arc coefficient. The net rotor MMF, $F_r$ is calculated from (2.12).

$$F_r(9, \alpha) = U_r \times ST$$  \hspace{1cm} (2.11)
In order to derive the rotor MMF as a function of the armature reaction, the stator MMF has been calculated in rotor reference frame according to (2.13), applicable in steady state.

\[ p \Psi_s = p \Psi_r + \omega_m t \]  
\[ p \Psi_s = p \Psi_r + \omega_m t \]  \hspace{1cm} (2.13)

The net rotor MMF is a function of magnet MMF and armature reaction MMF from the stator. The rotor MMF, \( F_r \) is derived as a function of \( F_s \) as well as rotor design parameters using a MCM given in Figure 2.7.

Figure 2.7. Magnetic circuit network showing one pole pair used for calculation of rotor MMF.

The flux crossing the flux barrier for the corresponding armature reaction MMF at point \( F_r \) shown in the circuit can be solved according to (2.14) and the rotor MMF derivation is given in (2.15) and is used to derive \( B_g \).

\[ \Phi_r = \Phi_{rem} + \Phi_g \]  
\[ \Phi_r = \Phi_{rem} + \Phi_g \]  \hspace{1cm} (2.14)
\[ F_r(\theta_s, \omega t) = (\varphi_h + \varphi_{rem}) R_{fb} \]

\[
= \left[ \frac{n}{2p+n_{vtb}} \int_{\frac{n}{2p+n_{vtb}}}^{n} B_g(\theta_s, \omega t) L_s \frac{D}{2} d\omega t + \mu_0 \mu_r H_c h_m L_s \right] \frac{t_h}{\mu_0 l_h L_s} \]

\[
= \frac{D t_h}{2g l_b} \left[ \int_{\frac{n}{2p+n_{vtb}}}^{n} F_s(\theta_s, \omega t) d\omega t - 2g_h F_r(\theta_s, \omega t) d\omega t + \frac{2g}{D} \mu_r H_c h_m \right] \]

\[
= Y \left[ \frac{D t_h}{2g l_b} \left( \frac{n}{2p+n_{vtb}} \right) \sum_m \sum_k F_{sk} \cos \left( mp \theta_s \pm (m_1 \alpha_c \pm m_2 \alpha_s) t \pm \varphi_h \right) + \frac{2g}{D} \mu_r H_c h_m \right] \]

\[
Y = \left( \frac{D t_h}{2g l_b} \right) \left( \frac{n}{2p+n_{vtb}} \right) \]

(2.15)

where, \( D \) is the stator bore diameter, \( \Phi_{PM} \) is remnant flux, \( h_m \) is length of the PM, \( \mu_r \) is PM relative permeability, \( H_c \) is PM coercive force, \( g \) is airgap length, \( t_h, l_b \) and \( \theta_b \) are thickness, length and angle of the flux barrier.

Under loaded operation, the effective flux density can simply be expressed as the superposition of the rotor flux reflected in the stator and the flux due to the stator currents for a given phase. Thus, the net air-gap flux density from the fundamental and harmonic reaction fields, \( B_g \) is calculated as a function of \( F_s \) and \( F_r \) as given in (2.16). In (2.17), \( B_g \) is written based on the fundamental stator MMF, rotor MMF and higher stator harmonics including time and space harmonics.

\[
B_g(\theta_s, \omega t) = \frac{\mu_0}{g_c} \left[ F_s(\theta_s, \omega t) - F_r(\theta_s, \omega t) \right] \]

(2.16)
2.4 Analytical Derivation of Harmonic Iron Losses

In this section, the improved modeling of iron losses including harmonic eddy current losses is derived and explained.

2.4.1 Procedure to Determine Harmonic Iron Losses

The procedure to determine stator harmonic iron losses is as follows:

1) The harmonic currents that have been derived (2.3) – (2.5) from the harmonic voltage spectrum and impedance have been used to derive the harmonic air gap flux density using (2.14). The higher harmonics considered were of the orders of $m_1f_c$ for carrier harmonics where $m_1=1, 2, 3, \ldots$, and $m_1f_c+m_2f_s$, for side–band harmonics where $n=\pm 1, \pm 2, \pm 3, \ldots$ for fundamental frequency $f_s$ and carrier frequency $f_c$.

2) The fundamental and harmonic iron losses, especially stator eddy current losses have been calculated using air–gap flux density and analyzed for varying operating conditions.

The air–gap flux density derived from the armature reaction fields is used to calculate the teeth and yoke eddy current losses according to (2.18) and (2.19), where $B_{htm}$ and $B_{hym}$ are teeth and yoke harmonic amplitudes of flux density, $k_t$ and $k_y$ are teeth width and yoke height coefficients, $\alpha_s$ is one tooth pitch angle, $\tau_s$ and $\tau_p$ are one tooth pitch and pole pitch in air–gap, and $w_t$ and $w_y$ are width of teeth and yoke respectively.
The stator iron losses are calculated as the sum of eddy current \((P_{EC})\) and hysteresis losses \((P_{Hy})\) in teeth and yoke in (2.20). where, \(k_e\) and \(k_{hy}\) are variable eddy current and hysteresis loss coefficients, \(\rho\) is the mass density of steel, \(P_{EC}\) and \(P_{Hy}\) are total eddy current and hysteresis losses, \(B_{htm}\) and \(B_{hym}\) are harmonic teeth and yoke flux densities calculated using Fourier decomposition, \(V_t\) and \(V_y\) are teeth and yoke volumes, \(n\) is Steinmetz coefficient.

\[
P_{EC} = k_e \rho_i V_t \sum_{h=1}^{\infty} (h \omega_r B_{htm})^2 + k_e \rho_i V_y \sum_{h=1}^{\infty} (h \omega_r B_{hym})^2
\]

\[
P_{Hy} = k_{hy} \rho_i V_t \sum_{h=1}^{\infty} h \omega_r (B_{htm})^n + k_{hy} \rho_i V_y \sum_{h=1}^{\infty} h \omega_r (B_{hym})^n
\]

The assumptions of the developed model are: (a) Reaction field caused by eddy currents in the stator and the rotor has been ignored as it is negligible; (b) The slot harmonic and lower order current harmonics have not been included in as the primary focus was to segregate the core losses due to PWM time harmonics. However, these harmonics can easily be considered in the model; and (c) Saturation of iron path is neglected.

Figure 2.8 represents the results from the developed model in terms of the net MMFs developed at an operating condition of 575 rpm, 10 kHz carrier frequency, 70 Nm and 200 V DC link voltage. The results have been plotted for one pole and by varying the rotor position, \(\omega t\). The results represent only fundamental and higher order carrier harmonics.
It is seen that the envelope of stator MMF in rotor reference frame contains time harmonics that leads to increasing eddy current losses and torque ripples.

Figure 2.9 shows the comparison of air–gap flux density obtained from the developed model considering only fundamental time harmonic component and 2–D FEA using co–simulation. The results demonstrate that there is a significant deviation when time harmonics are not considered.

![Figure 2.8](image1)

(a) Stator MMF in rotor reference frame. (b) Rotor MMF considering fundamental and higher order carrier harmonics in input current.

![Figure 2.9](image2)

Figure 2.9. Comparison of air–gap flux density calculated with respect to rotor position using only fundamental component of armature reaction field in air–gap flux density analytical model with 2–D FEA for considering only fundamental current component.

Figure 2.10 shows the comparison of air–gap flux density obtained with respect to rotor position from the developed model considering space harmonics and side–band time
harmonic components and 2–D FEA using co–simulation. The analytical model closely follows the 2–D FEA results. The obvious difference in the $B_g$ is because the model deliberately ignores the lower order time harmonics and slot harmonics to separate the contribution due to carrier harmonics only. Figure 2.11 shows the comparison of air–gap flux density obtained with respect to time from the developed model considering space harmonics and side–band time harmonic components and 2–D FEA using co–simulation.

Figure 2.10. Comparison of air–gap flux density calculated with respect to rotor position using analytical model with 2–D FEA for considering space harmonics and side–band time harmonics in current.

Figure 2.11. Comparison of air–gap flux density calculated with respect to time using analytical model with 2–D FEA for considering space harmonics and side–band time harmonics in current.
2.4.2 Simulation Results of Harmonic Iron Loss Determination at Various Operating Conditions

The total losses in the motor were simulated to compare with the total losses obtained by the 2–D FEA co–simulation. This study was performed to understand the percentage of losses contributed by the harmonic iron losses from side-band harmonics. The iron losses were calculated using the developed model; contributions by side–band harmonics are shown as harmonic iron losses and that contributed by fundamental flux density is shown as fundamental iron losses. The copper losses were calculated as a function of peak current and stator resistance. The calculation of total iron losses from (2.20) was performed for 575 rpm and 700 rpm of the test IPMSM and peak currents of 2 A and 10 A for a fixed $V_{dc} = 650$ V, $f_c$ of 10 kHz and current angle, $\gamma$ of 30°. The results from the analytical calculation and comparison with 2–D FEA co–simulation are shown in Figure 2.12(a) for 2 A and Figure 2.12(b) for 10 A.

The results match closely with the 2-D FEA results. The slight discrepancy can be due to slot harmonic contributions in FEA and approximations in the MCM. It can be noticed that at lower speeds and currents, the harmonic losses are as significant as the copper losses whereas at higher loads, copper losses are predominant. The fundamental iron losses increase with speed and are predominant at low torque and high–speed region whereas percentage of harmonic iron losses with respect to total controllable losses is more at lower loads and speeds. Thus, it is imperative to consider harmonic iron losses also towards developing a global loss model towards maximum efficiency control strategies to include the effect of PWM harmonic iron loss in the prevalent low–torque, low–speed region [7].
2.5 Analytical Derivation of PWM Harmonic Factor Contributed by Magnet Eddy Current Losses

In the case of the rare earth magnets, the carrier harmonics produced by the PWM inverter may cause relatively large harmonic eddy–current losses because of the high conductivity of the magnet. This study aims at understanding the behaviour of magnet losses due to the parameters of the PWM such as DC link voltage and switching frequency. The procedure for calculating magnet loss factor is as follows:

Figure. 2.12. Calculation of iron and copper losses in IPMSM for 575 rpm and 700 rpm and comparison with 2–D FEA. (a) Results for 2A. (b) Results for 10 A.
(i) In order to study the effect of baseband, carrier and sideband harmonics, the analytical method was initially used to predict PWM loss factor for a known value of harmonic current. This method is termed current harmonic prediction. The orders of the major harmonics included in the PWM waveform with carrier frequency $f_c$ and fundamental frequency $f_s$ are $(2^{j-1}) \frac{f_c}{f_s}$ plus–minus second, $(2^{j-1}) \frac{f_c}{f_s}$ plus–minus fourth, $2j\frac{f_c}{f_s}$ plus–minus first, and $2j\frac{f_c}{f_s}$ plus–minus fifth ($j=1, 2, 3, \ldots$), etc. [37].

(ii) The magnitude of peak current ($I_m$) of the laboratory IPMSM being 15.5 A, harmonic current magnitudes were given as a percentage of peak current (order $2m/plus–minus one has maximum magnitude of 5\% of $I_m$) and the superposed current equation was subsequently used to calculate magnet loss for various operating conditions using the developed model.

(iii) Once the model was validated for assumed values of harmonic currents, a magnet loss factor calculation technique was followed for the laboratory IPMSM wherein the current harmonics derived in (2.3) were used. The magnet loss factors were calculated for varying $f_c$ from 2~7 kHz for fixed DC voltage. For validation of the developed method, 2–D FEA as well as experimental tests have been developed.

The time varying magnetic flux density, $B_{mil}$ that is proportional to the magnet loss caused by the interaction of time harmonics from PWM and the space harmonics from winding can be derived from (2.21) as the rate of change of flux density in the rotor caused by armature reaction rotor MMF, $F_r$, which is derived using (2.15).
\[ B_{nl}(\omega t) = \frac{H_0 H_L}{h_m} F_r \]  \hspace{1cm} (2.21)

\[ P_n \propto \frac{1}{\omega^2} \frac{2}{T} \left( \frac{d}{dt} \sum_{i=1}^{\infty} B_{nl}(\omega t) \right)^2 dt \]  \hspace{1cm} (2.22)

From (2.21) and (2.22), the time-varying magnetic flux density that determines the loss due to PWM supply, namely PWM loss factor is determined as a coefficient which is a function of armature reaction rotor MMF and current phase angle and can be used for understanding magnet loss behavior. Figure 2.13 shows the current harmonic magnitudes used to study magnet eddy current losses and Table 2.3 elucidates the PWM loss factor calculated for various \( f_c \) and corresponding validation from 2–D FEA. It can be concluded that the magnet eddy current losses decrease significantly with an increase in switching frequency. The same trend is obtained by the PWM loss factor.

![Figure 2.13. Current harmonic magnitudes used to study magnet eddy current losses.](image)

**Table 2.3 Magnet Losses for Various Carrier Frequencies with 200 VDC and 40 Hz Fundamental**

<table>
<thead>
<tr>
<th>Carrier Frequency (kHz)</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
</tr>
</thead>
<tbody>
<tr>
<td>PWM Loss Factor</td>
<td>0.235</td>
<td>0.136</td>
<td>0.118</td>
<td>0.099</td>
<td>0.076</td>
<td>0.032</td>
</tr>
<tr>
<td>2–D FEA (W)</td>
<td>17.1</td>
<td>12.8</td>
<td>12.1</td>
<td>9.05</td>
<td>8.3</td>
<td>7.2</td>
</tr>
</tbody>
</table>
2.6 Experimental Investigations

The developed analytical models have already been validated using 2–D FEA co-simulation model with PWM input and FOC control. However, to study the accuracy of the model and its applicability to real motors, experimental validations have been performed on the 4.25 kW laboratory prototype. The Opal–RT real time controller was used to control the IPMSM. The current angle was chosen at MTPA angle for respective load points. Figure 2.14 shows the total stator iron losses including harmonics at 200 rpm, 15 Nm, and varying DC link voltages at 7 kHz. The analytical results have been obtained using (2.20).

![Figure 2.14. Comparison of analytical and experimental results for total iron losses in stator under varying loading conditions for 200 rpm, 15 Nm and 7 kHz.](image)

The total loss measured indicates the power losses after copper losses and mechanical losses were segregated. To calculate the mechanical losses, which include the frictional and windage losses of test motor, no–load tests were conducted by running the test motor with a prime mover. The method of performing no–load tests with rotor of the motor prior to magnet installation to segregate mechanical losses is given in Appendix B. After segregation of the copper and mechanical losses, the fundamental and harmonic iron loss
and magnet eddy current loss can be determined. To understand the portion of magnet eddy current losses, a 2-D FEA co-simulation was performed. It can be seen that the iron losses, which includes fundamental and harmonic losses slightly decrease with decreasing DC link voltage. The magnet eddy current losses also decrease with decreasing DC link voltage. The total losses that were measured in the experiments match closely with the summation of iron losses calculated from the analytical model and magnet eddy current losses calculated using 2-D FEA. Thus, the developed method has been validated experimentally.

In addition to the loads test validations, additional blocked rotor test developed in [12] was performed to test for variation of losses due to carrier harmonics to validate the trend of the carrier harmonic loss results provided by analytical model. The test circuit and setup are developed in such a way that the rated harmonic iron losses can be represented and measured. The fundamental frequency of the PWM inverter is set to be zero. Subsequently, the DC currents including the inverter carrier are supplied to the test motor, whose rotor is locked. In this case, both the mechanical as well as output power are zero. The input power was measured using a power analyzer. The copper losses were segregated from the input power. The remaining power is contributed by the carrier harmonic iron losses and magnet eddy current losses due to PWM harmonics. The test circuit for input to the three-phase winding of IPMSM from inverter is given in Figure. 2.15(a) and the PWM losses including iron and magnet losses is plotted for increasing $f_c$ in Figure. 2.15(b).

The experimental results show the same trend as suggested by the developed model. The decreasing losses with increasing carrier frequency is mainly contributed by the decrease in magnet eddy current losses.
2.7 Summary

The chapter proposed an analytical model to calculate and analyze the stator iron losses as a function of PWM parameters through the combination of improved winding function and rotor MMF theories. The model incorporated carrier harmonics from a PWM inverter as well as space harmonics from the motor. The harmonic airgap flux density and hence, iron losses in stator were calculated using armature reaction–based rotor MMF as a function of stator harmonic MMF. The developed model suggested that the harmonic eddy current losses are predominant at low load conditions. The model was also used to calculate PWM loss factor for magnet eddy current losses due to carrier harmonics. The analytical model has been validated using 2–D FEA and experimental investigations. The novel contributions of this chapter are the analytical model incorporating time harmonics for an IPMSM and the derivations of harmonic iron losses considering the armature reaction field air–gap flux density to study the impacts of PWM parameters.
CHAPTER 3

INVESTIGATION INTO VARIATION OF PERMANENT MAGNET SYNCHRONOUS MOTOR AND DRIVE LOSSES FOR SYSTEM LEVEL EFFICIENCY IMPROVEMENT

3.1 Introduction

In this chapter, an investigation into the control methods towards efficiency improvement in PMSM drives is performed. The analysis results in two vital contributions: 1) comprehensively analyzing the behaviour of the motor and drive power losses to various inverter parameters and control variables such as current angle, DC link voltage and switching frequency through analytical and electromagnetic models of PMSM drive, and, 2) addressing the problem of finding the minimum power losses at the system level rather than individual component level. The overall system losses, measured from the DC link to the motor output mechanical power, are contributed by the losses shown in Figure 3.1.

![Figure 3.1. Schematic depiction of controllable losses in PMSM drive system.](image)

To analyze the system efficiency, it is imperative to consider the sensitivity of the losses of motor and inverter separately, as well as the system losses as a whole in order to find...
the best method that consumes least battery power through control methodologies. Some limited work existing on loss minimization for the motor and drive system as a whole, are presented in [58]–[61]. It is to be noted that most of the work considering motor harmonic losses as well as system level efficiency improvement has been addressed in case of induction motors [58], [60]. In [59], comprehensive losses of PMSM inverter for the accurate calculation of the overall system losses, with respect to switching frequency and the $d$-axis current of the drive system. However, the harmonic effects of the inverter on motor iron losses are not considered. In [61], an optimum switching frequency has been derived considering motor and inverter losses. An FEA model of the motor and analytical models for inverter were used to calculate the losses. The switching frequency and modulation type and index of the inverter affects not only the losses in the inverter, but also the fundamental and harmonic losses in the motor [41], [62]. The inverter parameters have been found to significantly affect the electrical losses in the stator and rotor of PMSMs [48]. For PMSM, variable DC link and switching frequency control methods have been proposed in literature to reduce losses in the inverter and in some cases, in both motor and inverter [48]. In drivetrain configuration with DC–DC converter that is connected between the battery and the inverter [48], it is possible to improve the motor output by increasing and controlling the inverter supply voltage, i.e., the system voltage without increasing the battery cost and size and simultaneously keeping the same motor size [63]. This topology also permits manufacturers to separately design the system voltage and battery, allowing for flexible system designs for vehicles with different output characteristics. DC link current minimization technique has also been studied in current source inverters [64]. Variable switching frequency methods have been studied in the literature for reducing the
current ripples in the inverter and subsequently improving the efficiency [65]. In [58],
direct torque control (DTC) and indirect field-oriented control (IFOC) were performed to
achieve maximum efficiency through flux selection for motor as well as system levels. The
results showed that the operating point for minimum system-level power loss is slightly
different from that for minimum machine power loss, but both can yield significant energy
savings. In order to analyze the system efficiency and to choose the appropriate control
variable, it is imperative to consider the sensitivity of the losses of motor and inverter
separately, as well as the motor–drive system losses as a whole in order to find the best
method to understand system efficiency improvement through changing the control
methodology.

Figure 3.2 shows the various control variables used to study the variation of system level
losses. This control diagram is also used for the FOC control of the test IPMSM to study
motor and system losses at specific operating conditions. First, the inverter loss models are
defined, followed by motor loss models. Subsequently, the three control variables, current
angle, switching frequency and DC link voltage are varied to study the behaviour of the
controllable losses. Based on the results of the proposed analysis, the chapter recommends
possible solutions for system–level loss minimization in PMSM drives.

Figure 3.2 Control scheme showing the variables for analyzing the behavior of system–level losses.
3.2 Mathematical Modeling of Inverter Losses

The dominating losses in hard–switched three–phase IGBT based inverters are conduction and switching losses in the IGBTs and freewheeling diodes. In this section, the losses in inverter have been calculated analytically assuming sinusoidal modulation. The temperature dependency of losses has been neglected and linear modulation is assumed.

3.2.1 Conduction Losses of VSI Inverter

The conduction losses of the six IGBT switches and freewheeling diodes in a two–level inverter can be calculated as the sum of average losses of the two components and depend on the load current, power factor angle, modulation type and consequently, modulation index as given in (3.1) [62], [66].

\[
P_{co} = \sum_n 6I_n \left( \frac{V_{cet}}{2\pi} + \frac{I_n r_{ce}}{8} + M_i \left( \frac{V_{cet}}{4} + \frac{I_n r_{ce}}{3\pi/2} \right) \cos \phi_n \right) \\
+ \sum_n 6I_n \left( \frac{V_{ft}}{2\pi} + \frac{I_n r_{f}}{8} + M_i \left( \frac{V_{ft}}{4} + \frac{I_n r_{f}}{3\pi/2} \right) \cos \phi_n \right)
\]  

(3.1)

where \( I_n \) is the magnitude of \( n \)th–order harmonic current, \( M_i \) is the modulation index, \( \cos \phi_n \) is the power factor of each harmonic order, \( V_{cet}, V_{ft} \) are the IGBT and diode threshold voltage, and \( r_{ce}, r_{f} \) represent the resistances of IGBT and diode respectively. In this study, only fundamental current is considered for the conduction losses.

3.2.2 Switching Losses of VSI Inverter

The switching loss occurs while turning the switching device on and off during the inverter operation. The switching losses in the IGBT and the diode are the product of switching energies and the switching frequency, \( f_s \). The switching losses in an IGBT inverter \( P_{sw} \), is given in (3.2).
\[
P_{sw} = \frac{6}{\pi} f_c \left( E_{on,I} + E_{off,I} + E_{err,D} \right) \left( \frac{V_{dc}}{V_r} \right)^k \left( \frac{I_L}{I_r} \right)^l
\]  

where \( f_c \) and \( V_{dc} \) are the switching frequency and DC link voltage, \( I_L \) is the peak line current, \( E_{on,I} \) and \( E_{off,I} \) are respectively the turn–on and turn–off energies of the IGBT, \( E_{err,D} \) is the turn–off energy of the power diode due to reverse recovery current, \( K_v \) and \( K_i \) are the exponent terms for non–linear voltage and current dependency of switching losses, \( V_{ref} \) and \( I_{ref} \) are the voltage and current reference values of switching loss measurements from data sheet of inverter. The properties of inverter used in this study are shown in Table 3.1.

**TABLE 3.1 PARAMETERS AND VALUES OF IGBT INVERTER USED FOR ANALYSIS**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>( V_{ce}, V_{ft} ) (V)</th>
<th>( r_{ce}, r_f ) (mΩ)</th>
<th>( E_{on,I}, E_{off}, E_{err} ) (mJ)</th>
<th>( K_v, K_i )</th>
<th>( V_{ref}, I_{ref} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>0.9, 2.2</td>
<td>2.88, 2.5</td>
<td>33, 56, 30.5</td>
<td>0.6, 0.6</td>
<td>400 A</td>
</tr>
</tbody>
</table>

3.3 **Mathematical Modeling of Motor Losses**

The losses in PMSM can be classified into fundamental losses, harmonic losses and mechanical losses. The fundamental electrical losses include copper losses in the winding, and core losses in the steel. The mechanical losses are classified into bearing losses and friction and windage losses. These losses depend on the speed of the motor, are electrically uncontrollable for a given torque, and speed condition. The harmonic losses, as described in the previous chapter, occur from time and space harmonics in the PMSM. In this section, analytical models of motor fundamental and harmonic stator eddy current losses used to study the behavior of these losses to control variables are elicited. The fundamental loss models are defined initially and subsequently, the analytical models developed in Chapter 2 for harmonic air gap flux density are used to calculate the harmonic iron losses in stator.
steel. The variation of the losses derived by the analytical models with control variables have been studied in section 3.4. The harmonic eddy current losses in the magnet have been derived using a coupled electromagnetic model in section 3.6.

3.3.1 Copper Losses

The fundamental copper losses where $I_{rms}$ is the RMS current and $R_{dc}$ is the DC resistance is defined as in (3.3).

$$P_{cu} = I_{rms}^2 R_{DC}$$ (3.3)

The harmonics in current, especially the baseband harmonics as well as skin and proximity effects lead to additional copper losses. The harmonic copper losses can be calculated by using the RMS current to calculate the total copper losses, $P_{Cu,total}$ in (3.4) and the losses induced by the skin and proximity effects can be represented as $R_{n,AC}$ which is calculated in (3.5) by multiplying $R_{dc}$ with AC skin and proximity effect gains, $K_{n,se}$ and $K_{n,pe}$ respectively.

$$P_{Cu,total} = I_{rms}^2 R_{DC} + \sum_{n=3}^{\infty} I_n^2 R_{n,AC}$$ (3.4)

$$R_{n,AC} = R_{DC} \left( K_{n,se} + K_{n,pe} \right)$$ (3.5)

3.3.2 Fundamental Iron Losses

The stator iron losses consist of eddy current, hysteresis and excess losses in the stator core. The iron losses are caused by time variation of flux density in the stator teeth and yoke. The total stator iron losses, $P_{iron,f}$ can be derived using (3.6) where $B_g$ is the magnitude of fundamental flux density in the airgap, $k_e$, $k_{hys}$ and $k_{exc}$ are the eddy loss coefficient,
hysteresis loss coefficient and excess loss coefficient, respectively, \( n \) is the order of the time harmonic, and \( f \) is the fundamental frequency.

\[
P_{\text{iron}, f} = k_{hys} B_{g}^{n} (f) + k_{e} B_{g}^{2} (f^2) + k_{exc} B_{g}^{1.5} (f^{1.5})
\]  

(3.6)

The fundamental iron losses given in (3.6) can be represented as a parallel resistance to the supply voltage in the PMSM equivalent circuit given in Figure 3.3. The parallel resistance, \( R_i \) can be determined by experiments, numerically as well as analytical modeling techniques. In this study, the determination of \( R_i \) has been performed by experimental tests. The advantage of experimental and equivalent circuit methods of determining \( R_i \) is that the motor design or steel parameters do not have to be known to implement the identification process. However, the experimental measurements should be conducted as accurately as possible in order to obtain \( R_i \) that is close to the real value. The determination of \( R_i \) has been performed using a no-load test performed at varying speeds [63]. A simple and widely

![Figure 3.3. d–and q–axis equivalent circuit model of IPMSM incorporating iron loss resistance. (a) d–axis model with iron loss. (b) q–axis model with iron loss.](image)
used method of determination of $R_i$ without knowledge of machine design details is to use results from no–load tests applied to the equivalent circuit model [63]. The no–load test has been conducted on the motor by keeping $i_d=0$ and varying $q$–axis current and speed. The value of $R_i$ has been obtained from (3.7).

$$R_i = \frac{V^2 - \omega_e^2 \left[ \lambda_{PM}^2 + (L_{sh} - L_{q})^2 \right]}{P_{in} - \omega_e \lambda_{PM} i_q - R_i i_q^2} \quad (3.7)$$

where, $V = \sqrt{(v_{ds}^2 + v_{qs}^2)}$ is the stator terminal voltage considering fundamental only, $P_{in} = V_q I_q$ when $i_d=0$. Using (3.7), the results of iron loss resistance calculation for the test motor is given in Figure 3.4.

![Figure 3.4. Iron loss resistance values for varying speeds from conducted on the laboratory IPMSM.](image)

Once the iron loss resistance is determined, the magnetizing currents, $i_{md}$ and $i_{mq}$ can be derived from the air–gap voltages in $d$–$q$ axis, $u_{md}$ and $u_{mq}$ as (3.8).

$$u_{md} = u_d - R_i i_d \quad u_{mq} = u_q - R_i i_q$$

$$i_{md} = i_d - \frac{u_{md}}{R_i} \quad i_{mq} = i_q - \frac{u_{mq}}{R_i} \quad (3.8)$$

The iron losses are assumed to be the same in direct and quadrature axes. Thus, the fundamental iron losses are written as a function of the magnetizing currents and
inductances, $L_d$ and $L_q$ and PM flux linkage, $\lambda_{PM}$ as in (3.9). The rated values of inductances and PM flux linkage are already given for the test motor in Table 2.1.

\[
P_{\text{iron},f} = \frac{\omega^2}{R} \left( (\lambda_{PM} + L_di_d) + (L_qi_q)^2 \right)
\]

(3.9)

### 3.3.3 Harmonic Iron Losses

The harmonic iron losses that have been derived in Chapter 2 is used in this Chapter to study the variation of harmonic iron losses with respect to control variables. The harmonic voltage spectrum that depends on the DC link voltage, $V_{dc}$, modulation index, $M_a$ and switching frequency, $f_c$ was given in (2.1) and the harmonic iron losses were derived using (2.20) by not considering the fundamental frequency and flux density.

### 3.4 Dependence of Inverter Losses on Control Variables

In this section, the inverter losses have been studied as functions on the current angle, switching frequency and DC link voltage. The conduction and switching losses in the inverter have been calculated using analytical equations (3.1) and (3.2) for varying line currents, switching frequencies and modulation index, therefore DC link voltage. The inverter parameters of the semiconductor switches and diodes in Table 3.1 for the inverter have been extracted from the data sheets of Semikron IGBT modules. For a given inverter, the conduction losses depend on the line current, modulation index and power factor angle whereas the switching losses depend on switching frequency and DC link voltage. The modulation index was varied from 0.2 to 1 and the line currents were varied from 2 A to 16 A based on the IPMSM motor ratings in the simulations. The power factor angle was swept from $-90^\circ$ to $90^\circ$. Figure 3.5(a) shows the results of inverter conduction losses calculated for varying modulation indices and load currents at fixed power factor angle and
Figure 3.5(b) shows the variation in conduction losses at different load currents and power factor angles at fixed modulation index. It can be seen that the conduction losses increase with increase in load current and modulation index but decrease with increase in power factor angle, $\phi$.

Figure 3.5. Conduction losses as a function of varying inverter parameters. (a) Conduction losses varying with respect to $M_i$ and load current. (b) Conduction losses as a function of power factor angle and load current.

Figure 3.6(a) shows the switching losses in the inverter as a function of varying $f_c$ and VDC. Figure 3.6(b) shows the variation of switching losses as a function of $f_c$ and load current.

Figure 3.6. Switching losses as a function of varying inverter parameters. (a) Switching losses varying with respect to $f_c$ and VDC. (b) Switching losses varying with respect to $f_c$ and load current.
From the analysis, it can be understood that the switching losses proportional to \( f_c \) as well as the load current. The results also suggest that the switching losses increase with increasing VDC. However, conduction losses tend to increase with an increase in modulation index. The phasor diagram of IPMSM in motoring mode is given in Figure 3.7 where \( \delta \) is the load angle, \( \theta \) is the power factor angle, \( E_{ph} \) is the induced EMF, and \( V_{ph} \) is the terminal voltage. The current angle, \( \gamma \) is related to the power factor angle using the relationship in (3.10), which means that for the same load angle, a higher current angle leads to a higher power factor angle, thus, an increase in current angle decreases the conduction losses.

\[
\gamma = \theta - \delta \tag{3.10}
\]

Figure 3.7. Phasor diagram of IPMSM in motoring mode.

The modulation index, \( M_i \) can be defined as the ratio of peak voltage to the DC link voltage according to (3.11).

\[
M_i = \frac{2V_{1.m}}{V_{DC}} = \frac{2}{V_{DC}} \left( \sqrt{V_d^2 + V_q^2} \right)
\]

(3.11)

With the DC link voltage constant, an increase in current angle changes the voltage operating point. The peak voltage decreases with increasing \( \gamma \). Thus, with increasing \( \gamma \), the modulation index decreases as well. It was seen from Figures 3.5 and 3.6 that conduction losses increase with increasing \( M_i \). Thus, an increase in \( \gamma \) can further decrease the
conduction losses. An increase in DC link voltage will reduce the modulation index for the same speed point. Thus, optimized DC link voltage can also reduce conduction losses. However, it is preferable to keep $M_i$ higher since the DC bus utilization will be higher and harmonic content will be lower. However, it is to be noted that the decrease of $M_i$ from increase in $\gamma$ is minor but helps in decreasing the inverter losses slightly.

Considering that the IGBT is chosen as the switching device, the switching losses can be reduced mainly by reducing the switching frequency. However, this selection depends on the operating speed of the motor and needs to be optimized considering the operating points. An increase in VDC also increases the switching losses in IGBTs and the diodes. Summarizing the effect of the control variables on inverter losses, switching losses can be reduced by decreasing the switching frequency or keeping a lower VDC. The reduction is also possible by choosing wide-band gap switches such as GaN and SiC but the study of effects of other switches is out of scope of this thesis. Conduction losses can be decreased by increasing the current angle as well as by choosing an optimized DC link voltage.

3.5 Dependence of Motor Losses on Control Variables

The fundamental copper and iron losses and harmonic iron losses depend on the motor parameters such as stator resistance, inductances and operating conditions such as current, frequency, and voltage. A simulation model of the motor with sine PWM fed drive was developed with the analytical equations given for motor losses in section 3.3 to study the effects of varying current angle on the losses. The dynamic equations used to model the motor and other details are given in Appendix C. The control diagram given in Figure 3.2 is used for the current control of PMSM. The simulation of motor losses with varying control variables and the behavior of fundamental and harmonic losses of the motor are
discussed in this section. This simulation model has been used to calculate the system level losses at a given load– speed condition to compare the performance at different control variable condition.

3.5.1 Simulation Results of Fundamental Motor Losses with Varying Current Angle

To study the fundamental losses, for a given $I_m$ value, the current angle was changed from $0^\circ$ to $60^\circ$ by keeping the DC link voltage at 650 V and switching frequency at 12 kHz. The losses were calculated for varying current angles, peak currents and operating speeds. The simulation was performed at 175 rpm and 575 rpm of the test motor and the fundamental iron losses were calculated using (3.9). The variation of iron losses with varying $I_m$ and current angle is given in Figure 3.8(a) and (b) respectively. The fundamental copper losses for varying $I_m$ and current angle is given in Figure 3.9(a). The total fundamental losses at 575 rpm is given in Figure 3.9(b).

![Figure 3.8](image.png)

Figure 3.8. Fundamental iron losses as a function of $\gamma$ and varying $I_m$ at switching frequency of 12 kHz and DC link voltage of 650 V. (a) Iron losses at 175 rpm. (b) Iron losses at 575 rpm.
Figure 3.9. Fundamental copper and copper plus iron losses as a function of $\gamma$ and varying $I_m$ at switching frequency of 12 kHz and DC link voltage of 650 V. (a) Copper losses at 175 rpm. (b) Total fundamental losses at 575 rpm.

The iron losses increase significantly with increase in speed as well as loading whereas copper losses only increase with loading. It can be understood that for a given $I_m$, the iron losses decrease with increasing current angle, whereas copper losses are the same for all current angles. This phenomenon can be used in optimizing the current angle to achieve maximum motor efficiency considering copper and iron losses.

3.5.2 Simulation Results of Fundamental Motor Losses with Varying DC Link Voltage

The DC link voltage at the inverter determines the line–to–line voltage input at the motor terminal. In a sine–PWM fed inverter that is considered in this thesis, the line–to–line fundamental voltage at the motor terminal is a function of the modulation index and DC link voltage and can be derived as (3.12). The line current is a function of the voltage and impedance of the motor as (3.13).

$$V_{l-l} = 0.613 * M_s * V_{DC}$$  \hspace{1cm} (3.12)

$$I_{l,rms} = \frac{V_{l-l,rms}}{R_s + jL_s}$$  \hspace{1cm} (3.13)
Figure 3.10(a) shows the variation of iron losses and copper losses under constant torque-region at 12 kHz, 575 rpm and 70 Nm for varying DC voltages. Figure 3.10(b) shows the variation of the iron and copper losses for 1,000 rpm and 40 Nm, where the IPMSM is operating at flux-weakening condition.

For constant torque region, if the DC link voltage is varied, for the same speed, the modulation is performed in such a way that the line–line voltage and hence line currents meet the speed and load demands. For an increase in DC link voltage, the modulation index varies, however, the line–line voltage remains the same to meet the speed demand. Thus, the fundamental iron losses do not vary with a change in DC link voltage. According to (3.13), for the same loading conditions, if voltage does not change, the line current does not change as well. Hence, fundamental copper losses do not vary with a change in the DC link voltage. However, for flux- weakening region, the rated speed varies when the DC link voltage is varied. It will be higher for a higher voltage and constant torque region is extended. This will create significant changes in the voltages at higher speeds between the
two DC link voltages. The currents are also different during flux weakening for the two DC link voltages. There is an increase in $d$-axis current with the initiation of flux weakening operation. This initial operating point will start earlier at 450 V compared to 650 V. Thus, the copper losses increase as well for a lower DC voltage. The iron losses slightly increase at 650 V due to a higher flux linkage.

3.5.3 Results of Fundamental Motor Losses with Varying Switching Frequency

The fundamental motor losses depend only on the fundamental flux linkage and fundamental current. The changes in switching frequency affects the harmonic spectrum and not the fundamental quantities. Thus, a change in switching frequency is not expected to cause any changes in the fundamental losses.

3.5.4 Simulation Results of Harmonic Motor Losses with Varying DC Link Voltage

The harmonic iron losses depend on the time harmonic spectrum according to (2.18). The modulation index plays a prominent role in the harmonic spectrum in a PWM–fed motor. The voltage ripples as well as current ripples vary with changes in the DC link voltage. With an increase in DC link voltage, the modulation index decreases, thus increasing the harmonics in the side bands and carrier frequency range. This leads to significant increase in harmonic iron losses. The simulation results of harmonic losses for varying DC link voltage at 275 rpm and torques of 35 Nm and 70 Nm are shown in Figure 3.11(a). The results for 575 rpm at the same loading conditions are given in Figure 3.11(b). The switching frequency was kept constant at 12 kHz in both cases. It is understood from Figures 3.11(a) and 3.11(b) that increasing DC voltage increases the harmonic iron losses.
Figure 3.11. Comparison of harmonic iron losses at 450 V and 650 V and torques of 35 Nm and 70 Nm for varying speeds. (a) 275 rpm. (b) 575 rpm.

The difference is more obvious at higher speeds and loading conditions. The increase in harmonic iron losses can be attributed to the decrease in $M_a$ to maintain the line voltage according to the speed demand. This decrease in $M_a$ causes increase in the harmonic distortion, thus increasing the harmonic iron losses. Thus, lower $M_a$ is not suitable for harmonic iron losses, especially at higher speeds and loading conditions.

3.5.5 Simulation Results of Harmonic Motor Losses with Varying Switching Frequency

The harmonic iron loss calculations were performed by varying the switching frequencies and keeping the DC voltage at 450 V for loading conditions of 575 rpm and 35 Nm. The simulation results are shown in Figure 3.12. It can be seen from the trend that the harmonic iron losses decrease slightly as the switching frequency increases. This is due to the increase in harmonic order of switching harmonics affecting the iron losses. According to (3.13), the higher the harmonic order, the higher the impedance and subsequently, lower the current harmonics.
3.5.6 Simulation Results of Harmonic Motor Losses with Varying Current Angle

The harmonic iron loss calculations were performed by varying the current angles and calculating the air-gap flux density with only carrier and side–band harmonics from analytical model in Chapter 2. The losses were studied for high and low loads and speeds for various current angle values and the results are shown in Figure 3.13.

Figure 3.13. Comparison of harmonic iron losses for varying current angles at varying speed and load conditions. (a) $I_m = 2$ A and varying speeds. (b) $I_m = 14$ A and varying speeds.

It can be seen that the trend of harmonic iron losses is dissimilar for different speeds and loads. The losses decrease with increasing $\gamma$ for low loads. However, the losses increase with increasing $\gamma$ at higher loads and this is predominant with increasing speeds. This
behavior can be explained by the trend of harmonic current magnitudes with \( M_i \) and thus, current angle [55], [56]. In first carrier frequency domain, the \( \omega_c \)-order harmonic component declines as modulation index increases, however the \( \omega_c \pm 2\omega_s \) and \( \omega_c \pm 4\omega_s \)-order components gradually increase instead. On the other hand, the \( 2\omega_c \pm \omega_s \), \( (2\omega_c \pm 5\omega_s) \) - and \( (2\omega_c \pm 7\omega_s) \) - order harmonics in second carrier frequency domain gradually ascend as modulation index rises. Thus, the harmonic iron loss trend varies with speeds and loads accordingly.

Considering an optimal current angle using the developed loss model will enable reduction of the harmonic iron losses.

3.6 Study of Magnet Eddy Current Losses with Varying Control Variables Using 2–D FEA Co-Simulation

The magnet eddy current losses have been calculated using a 2–D electromagnetic model for varying DC link voltage and switching frequencies. Figure 3.14(a) and 3.14(b) show the variation of magnet losses with switching frequency and DC link voltage respectively for 575 rpm and 70 Nm.

![Graph](image)

Figure 3.14. Comparison of magnet eddy current losses for varying switching frequencies and DC link voltage. (a) Varying switching frequency at 650 V. (b) Varying DC link voltage at 12 kHz.
The magnet losses decrease with an increase in switching frequency and increase with increasing DC voltage. When the switching frequency is high, the harmonic order is high, subsequently resulting in high impedance and low current harmonic value that causes the eddy current losses in the magnet. Thus, it is ideal to keep the switching frequency as high as possible to keep magnet losses minimum. A higher DC for the constant torque region leads to a decrease in $M_i$, thus increasing magnet losses.

3.7 **System Level Losses with Varying Control Variables**

The system level losses were calculated for varying loading conditions in a FOC based simulation of IPMSM. Except the case of varying current angle, the current angle was fixed at the traditional MTPA angle for a given load and speed. The subsequent subsections discuss the system–level loss behavior due to varying control variables.

3.7.1 **Current Angle**

The total losses were calculated for varying current angles at varying speeds and peak currents. Figure 3.15 shows the total system losses that were calculated for varying $I_m$ values at 575 rpm.

![Figure 3.15. System losses calculated for varying current angles at 575 rpm and varying $I_m$.](image)
Figure 3.16 shows the input, output powers and total losses at 575 rpm and $I_m = 2$ A. The total system losses are decreasing with increasing current angle. The input and output powers given in Figure 3.16(a) as well as the efficiency given in Figure 3.16(b) are a concave function with respect to current angle. This suggests that there exists an optimal current angle corresponding to maximum efficiency at the system level.

![Graph showing input power, output power, and total losses](image1)

![Graph showing system efficiency](image2)

(a) (b)

Figure 3.16. Input power, output power, losses and system efficiency calculated for varying current angles at 575 rpm and $I_m = 2$ A. (a) Input and output power and total losses. (b) System efficiency.

### 3.7.2 DC Link Voltage and Switching Frequency

Initially, the DC link voltage was varied from the rated voltage of 450 V to 650 V for varying speeds and loads. The switching frequency was kept a constant at 12 kHz and the current angle was chosen to be the MTPA angle according to the loading condition. Figure 3.17(a) and 3.17(b) show the system level losses for varying $f_c$ and DC link voltages at 575 rpm and 70 Nm, which is the rated condition of the test motor. It can be concluded that for increasing $f_c$, the increase in switching losses is more prominent compared to the decrease in harmonic losses in the test motor, which is a low-speed, high-torque motor. The harmonic iron and magnet eddy current losses increase with speed.
Figure 3.17. System level power variation at rated speed and torque. (a) Losses as a function of $f_c$ at 650 V. (b) Losses as a function of VDC at 12 kHz.

Therefore, at higher speeds, the sensitivity of harmonic losses to $f_c$ are expected to be more prominent. Thus, variable switching frequency technique can be used in high–speed motors to select an optimal $f_c$ to maximize system efficiency. However, it is not recommended for low–speed motors.

For changing VDCs, the switching losses as well as harmonic losses of the motor increase with increasing VDC for a particular speed. While considering system level losses, the sensitivity of switching loss and motor losses due to VDC is higher than that of the conduction loss due to VDC. Hence, maintaining a lower VDC given that the voltage is sufficient for the demanded speed is recommended to reduce system losses. However, as mentioned in section 3.5.2, over the entire operating condition of the motor, the DC voltage determines the rated speed and initiation of flux–weakening peformance. The losses for a higher DC voltage are lower in flux–weakening conditions as a much lower current compared to a lower DC voltage. Thus, considering the overall torque–speed characteristics, higher DC voltage are expected to provide lower losses. A variable DC
voltage control enables the selection of optimal DC link voltage according to the operating condition.

3.8 Experimental Validation in Laboratory Prototype

The motor is tested under load mode and the speed is set as required by a speed dynamometer. All tests have been conducted using MTPA control technique except the test to study system losses as a function of varying current angle. Three tests were conducted for analyzing and validating the analytical models used towards the study of sensitivity of system–level losses to varying control parameters: (1) Measurement of system losses at varying VDCs by keeping a constant \( f_c \); (2) Measurement of system losses at varying \( f_c \) by keeping constant VDC; and, (3) The measurement of system–losses by varying current angle, \( \gamma \) and finding out optimal current angle that provides best system efficiency for a particular peak current. The input power at the drive side was measured using DC current clamps and voltage probes that were connected at the inverter DC link. The motor speed was measured using a high–resolution encoder. A high–resolution torque transducer is used to measure the torque between the test IPMSM and dynamometer.

3.8.1 Power Loss Sensitivity

The results of Test 1 wherein the system power losses are calculated as a function of \( f_c \) for a fixed VDC of 200 V for an operating condition of 100 rpm, \( I_m=10 \) A at a load torque of 50 Nm and \( \gamma=30^\circ \) is given in Figure 3.18(a). The results of Test 2 where the VDC was varied by keeping \( f_c \) constant at 9 kHz for the same operating condition as in Test 1 is given in Figure 3.18(b).
Figure 3.18. System power loss variation as a function of VDC and carrier frequencies. (a) Power loss variation as a function of VDC for fixed switching frequencies. (b) Power loss variation as a function of \( f_c \) for fixed VDC.

The experimental results suggest similar trend as obtained from simulations at low speeds: the inverter losses increase with increasing switching frequency and decrease with increasing modulation index. The decrease in motor harmonic losses with decreasing \( f_c \) is not prominent owing to the low speed. The results from test 3 in Figure 3.19(a) and (b) show that the system losses increase with increasing \( \gamma \) whereas the input, output powers and efficiency are concave functions with \( \gamma \), as seen in the simulations. The \( \gamma \) corresponding to maximum system efficiency was 37.5°. The \( \gamma \) for MTPA control at the same operating condition was close to 36°, which gives a slightly lower efficiency. The efficiency difference is more significant at lower loading conditions.

Figure 3.19. Input power, output power, losses and system efficiency calculated for varying current angles at 575 rpm and \( I_m = 15 \) A. (a) Input and output power and total losses. (b) System efficiency.

61
3.9 Discussion on Feasible Solutions for Improved PMSM System Efficiency

The following feasible solutions have been suggested from the developed analysis in this chapter: (1) *Variable DC link voltage method* should be followed for motors having wide range of operating speed. Consequently, VDC limit of the drive can be set high (for operation at high-speed conditions), enabling the reduction of DC current limit, and therefore conductor and motor size. However, variable DC link method requires a DC/DC converter and extra closed–loop control for the modulation index. (2) *Optimized switching frequency method* can be used in high speed motors where the motor harmonic losses will be significant. However, for low speed motors such a direct–drive motors, keeping a low \( f_c \) is optimum as the harmonic iron losses variation with switching frequency is not significant. (3) *Loss minimization and optimization solutions with current angle*, by considering system losses can be used to decrease the system current consumption, hence losses for low–and high–speed and load conditions. This method works for all types of motors. The method is expected to work better for high–speed motors.

3.10 Conclusions

This chapter proposes a comprehensive analysis of system level losses in a PMSM motor drive by considering various control variables and feasibility of minimization of the overall losses. The mathematical models used to analyze motor and the inverter losses as well as the experimental validation on test IPMSM have shown that DC link voltage, switching frequency and current angle can be controlled to reduce the system losses. The feasible solutions of loss minimization depending on motor type and requirements have been suggested. It was concluded that optimal current angle derivation considering system level losses is an ideal solution for improving the system efficiency in all types of IPMSMs.
CHAPTER 4

STUDY OF PARAMETER VARIATIONS IN PM MACHINES CONSIDERING SATURATION AND TEMPERATURE VARIATIONS

4.1 Introduction

In this chapter, parameter determination methods have been developed and tested in laboratory PMSMs to study the effects of saturation and temperature variations on the PMSM parameters. Two methods have been developed: (1) Off–line metaheuristic optimization–based parameter identification technique that considers saturation; and, (2) On–line recursive least square–based parameter determination method that considers saturation as well as temperature variations. The aims of this chapter are to develop accurate methods to determine the varying parameters of a PMSM with respect to load. It was concluded from Chapter 3 that optimizing the current angle through control techniques can lead to improvements in system efficiencies. The variation of optimal current angle with respect to speed and torque leads to varying current operating points. Through the development of parameter determination techniques, this chapter studies the importance of considering the effects that varying load has on parameters and the consideration of the same in efficiency calculations.

The developed methods have been validated in laboratory IPMSMs. Initially, an off–line parameter determination technique has been studied on a line–start IPMSM (LSIPMSM). Through the combination of experimental test methods conducted on the inverter connected LSIPMSM under varying operating conditions and a metaheuristic algorithm, parameters such as stator and magnetizing inductances and damper parameters have been identified for all conditions. Furthermore, an on–line RLS technique has been applied to
identify on–line, the inductances as functions of operating currents and permanent magnet
flux linkage and stator resistance as a function of operating temperature. The effect of iron
losses on the accuracy of estimation of machine parameters has been studied under varying
operating conditions and the performance has been compared with conventional estimation
without iron losses. The developed identification technique improves the precision of on–
line estimation.

4.2 Off–line Parameter Identification based on Metaheuristic Optimization
Considering the importance of parameter determination over a wide range of operating
conditions and the suitability of on–site off–line techniques, this section proposes an
innovative parameter determination technique for LSIPMSM using a constrained
optimization technique. The characterization procedure consists of the analysis of the
collected data and aims at fitting the predictions of the model to the observed behavior.
This primarily involves solving an optimization problem, where the parameters of the
model parameters are independent variables that may be adjusted to minimize the
prediction error. Initially, a conventional dynamic model of LSIPMSM has been used and
improved particle swarm optimization (IPSO) has been used as the constrained
optimization algorithm. The dependence of magnetizing inductances to magnetizing
current has been included in an improved dynamic model of LSIPMSM.
The results of this study are applied towards the identification of leakage and magnetizing
inductances, which are otherwise difficult to determine accurately from experiments or
finite element methods.

4.2.1 Improved LSIPMSM model Considering Saturation
Dynamic modeling and identification procedure has been applied to a 7.5 hp, 4–pole
LSIPMSM to determine the variable inductances as well as the damper circuit parameters.

The dynamic equations of the machine are elicited from (4.1) to (4.3), and the equivalent circuit representation for direct and quadrature axes are given in Figure 4.1. The symbols have standard notations as in [67]. From (4.1)–(4.3) and Figure 4.1, the parameters of the machine to be determined are \( R_s, L_d, L_q, L_{md}, L_{mq}, R_{kd}, L_{kd} \) and \( L_{kq} \).

\[
\begin{bmatrix}
    v_{ds} \\
    v_{qs} \\
    0 \\
    0
\end{bmatrix} = \begin{bmatrix}
    R_s + L_{ds}p & -\omega_r L_{qs} & L_{md}p & -\omega_r L_{mq} \\
    \omega_r L_{ds} & R_s + L_{qs}p & \omega_r L_{md} & L_{mq}p \\
    L_{md}p & 0 & R_{kd} + L_{kd}p & 0 \\
    0 & L_{mq}p & 0 & R_{kq} + L_{kq}p
\end{bmatrix}
\begin{bmatrix}
    i_{ds} \\
    i_{qs} \\
    i_{kd} \\
    i_{kq}
\end{bmatrix} + \begin{bmatrix}
    0 \\
    1 \\
    0 \\
    1
\end{bmatrix}
\]

(4.1)

\[
T_e = \frac{3P}{4} \left[ \lambda' i_{qs} + i_{ds} i_{qs} \left( L_{ds} - L_{qs} \right) \right] + \frac{3P}{4} \left[ L_{md} i_{kd} i_{qs} - L_{mq} i_{kq} i_{ds} \right]
\]

(4.2)

\[
p\omega_c = \frac{P}{2J} (T_e - T_L)
\]

(4.3)

Figure 4.1. Equivalent circuit illustration of conventional LSIPMSM dynamic model. (a) Direct axis. (b) Quadrature axis.

From these parameters, \( R_s \) and \( \lambda' \) are considered to be constant and are determined from conventional experimental tests, dc drop test for \( R_s \) and no-load test under constant speed by running the machine as a generator for \( \lambda' \). All other parameters are considered to be variables and are determined by IPSO along with experimental tests. In order to include the effect of magnetizing currents on inductance, the dynamic modeling has been improved to incorporate saturation effects in \( d- \) and \( q- \)axis magnetizing inductances and modified
for LSIPMSM. In [68], novel test methods for experimental parameter determination of the LSIPMSM considered in this study were developed. The measured values from [68] have been used for validation of the developed identification algorithm for line–start condition.

In order to determine machine parameters closer to the real values, it is imperative to use a model that represents the machine as efficiently as possible. Conventional two–axis model neglects saturation present in the motor ferromagnetic material. However, owing to the non–linear properties of the core material, at high currents, flux does not have a proportional relationship with the magnetizing current. In this study, an improved model has been proposed to include magnetizing inductance saturation. The effect of saturation in IPMSM can be modeled by considering the \( d \)– and \( q \)–axis inductances as functions of the \( d \)– and \( q \)–axis currents. Equation (4.4) represents the dependence of saturated inductances as a function of magnetizing currents \( i_{mx} \) [69], where \( i_{mx} \) is the current where saturation begins. In (4.4), \( x \) represents direct or quadrature axis and \( L_{mx} \) represents actual magnetizing inductance;

\[
L_{mx} = \begin{cases} 
L_{mxi} & i_{mx} \leq i_{mxi} \\
\frac{L_{mxi}}{1 + \zeta L_{mxi} i_{mxi} \left( \frac{1}{i_{mxi}} - \frac{1}{i_{mx}} \right)^2} & i_{mx} > i_{mxi}
\end{cases}
\]

(4.4)

\( L_{mxi} \) and \( i_{mxi} \) are the \( d \)– or \( q \)–axis magnetizing inductances and magnetizing currents respectively for unsaturated condition.

The value of \( \zeta \) in (4.4) can be obtained by two methods: (1) Estimation through optimization: \( \zeta \) can be included as unknown for the optimization algorithm to estimate. However, this would require knowledge of initial and final values of \( \zeta \); (2) Experimental
determination—For accurate results, experimental determination of saturation inductances with respect to magnetizing currents can be performed and a non-linear curve fitting algorithm can be used to find $\zeta$. In this study, $\zeta$ has been determined experimentally through DC excited static measurements [70], [71]. The DC static measurement method of inductance determination is well motivated compared to single-phase AC test owing to the representation of magnetizing flux linkage in steady-state and transient conditions similar to the real conditions [72].

Figures 4.2(a) and 4.2(b) illustrate the $d$– and $q$–axis current rise as a function of time. The experimental method used is similar to the method as mentioned in [71]. The DC currents used for determining the inductances corresponded to the peak value of the working AC current, thus representing saturation condition in the motor. Using (4.5), the values of inductances are calculated from magnetic flux based on the variation of the terminal voltage in a specific period of time.

$$L_d(t) = \frac{\int [v(x) - \frac{3}{2} R_s x i(x)] dx}{i(t)}$$
$$L_q(t) = \frac{\int [v(x) - 2 R_s x i(x)] dx}{i(t)}$$

(4.5)

where, $x$ is the variable of the integral and denotes time, $v(x)$ is the instantaneous terminal voltage, and $i(x)$ is the instantaneous terminal current of the LSIPMSM. Figures 4.3(a) and 4.3(b) show the dependence $d$– and $q$–axis inductances on magnetizing currents $i_d$ and $i_q$ respectively in the test motor. A non-linear interpolation method was used to determine the value of $\zeta$. Once $\zeta$ is determined, the condition given in (4.4) was used in the model in a loop with the optimization algorithm.
Employment of Optimization Algorithm and Parameter Identification

A new parameter identification approach using IPSO has been applied to the non-linear system of LSIPMSM to determine the varying inductances as a function of current as well as the damper parameters. PSO has been used to solve a wide range of optimization problems and is the chosen stochastic search algorithm owing to its easy implementation as well as high convergence when inertia and acceleration coefficients are selected accordingly [73], [74]. The flow diagram of the optimization procedure used has been provided in Figure 4.4.

In Figure 4.5 for each time step, and particle, the position is updated in (4.6) [73]– [76]:

\[ x_{k+1}^i = x_k^i + v_k^i + 1 \]  

(4.6)
Figure 4.4. Flow chart of the IPSO algorithm developed for parameter determination.

Figure 4.5. Particles updating in a circular behavior

Each particle in PSO is associated with a pseudo velocity of

\[ v^i_{k+1} = w_k v^i_k + c_1 r_{i,k} (p^i_k - x^i_k) + c_2 r_{2,k} (p^g_k - x^i_k) \]  

which represents the rate of change of position for the particle.

\[ (4.7) \]
Equation (4.7) is used to calculate each particle’s new velocity $v^i_{k+1}$, based on its previous velocity $v^i_k$, and the distances of its current position, $x^i_k$, from its own best experience (position) and the best experienced position of its own informants, $p^g_k$. Here, subscripts indicate a pseudo time increment and the number of particles, respectively.

Variables $r_1$ and $r_2$ represent uniform random numbers between 0 and 1, which will be regenerated in each iteration, $c_1$ and $c_2$ are two positive constants, called the cognitive and social parameters, respectively, (in this study, they are dynamically varying, $c_{1\text{first}}=1.5$, $c_{1\text{end}}=2.5$ and $c_{2\text{first}}=2.5$, $c_{2\text{end}}=1.5$), this is called improved PSO (IPSO) and has been used to achieve a more reliable and global result compared to conventional PSO [73]– [75].

The inertia weight, $w_k$ in (4.7) should neither be too large, which could result in premature convergence, nor too small, which may slow down the convergence excessively. They are chosen as $w_{\text{min}}=0.6$, $w_{\text{max}}=1.2$, holding the value at the beginning of each simulation cycle and increasing linearly until the end [18]. In this study, the information links between the particles were defined once and kept unchanged throughout the simulation. Each particle has a set of informants of fixed size, $k$. The neighborhood of size, $k$, of a particle is obtained from the virtual circle by recruiting alternately on the right and left of its position until a total of $k-1$ neighbors are obtained. The particle itself is also included, i.e., $k=8$. In this study, a swarm with thirty particles is used.

The IPSO algorithm is performed at each time step to determine if the calculated values of the variables meet the constraints. Subsequently, it is compared with the best result of the objective function $P_{\text{best}}$ and the best among the entire swarm is stored as $g$. The inputs to IPSO are phase currents, $I_a$, $I_b$, $I_c$ and angular velocity, $\omega_r$. The sensitivity analysis of the phase currents and angular speed of the machine for parameter variation has already been
studied for PMSMs [8]. According to algorithm, the possible answers will be analyzed, and the optimal value will be chosen from the possible answers. The dynamic equations of the machine given in (4.1)–(4.3) have been modeled in the state space form given in (4.8) and applied to IPSO. In (4.8), $x$ includes the state variables, $p$ includes the unknown variables that are determined, $u$ is the input, and $y$ is the output and includes voltages $V_d$ and $V_q$.

$$px = f(p, x, u)$$

$$y = g(p, x)$$

(4.8)

The fitness function used in the study is given (4.9).

$$CF(L_d, L_q, L_{md}, L_{mq}, L_{kd}, L_{kq}, R_k, R_f) = \sum_{k=1}^{N} \left[ (i_{ma}(k) - i_a(k))^2 + (i_{mb}(k) - i_b(k))^2 + (i_{mc}(k) - i_c(k))^2 + (\omega_{mr}(k) - \omega_r(k))^2 \right]$$

(4.9)

where CF is the cost function, $i_{ma}$ and $i_a$ are the measured and calculated values of phase A current respectively and $\omega_r$ is the angular velocity of the rotor in rad/s and are chosen as the outputs from the model. The same representation holds for phases B and C. The measured values of $i_a, i_b, i_c$ and $\omega_r$ and the calculated values for the same in system model are integrated with IPSO optimization to minimize the objective function given in (4.9).

4.2.3 Experimental Setup and Validations

The detailed block diagram of the experimental setup and the optimization process is shown in Figure 4.6 and an illustration of the experimental setup using the LSIPMSM is provided in Figure 4.7. The details of the motor used in this study is given in Table 4.1.
The LSIPMSM under test was a three-phase, four pole wye connected machine with rated voltage of 200 V under 50 Hz. The motor was driven by a Semikron voltage source inverter (VSI) and was loaded through a 7.5 hp DC generator supplying a variable resistive load. The motor was run as a line connected machine at 60 Hz as well as from the controlled Semikron VSI at frequencies from 10 Hz to 50 Hz at variable voltages. Due to the limitations of voltage from the DC supply in the laboratory, a maximum DC link voltage of 230 V could be supplied for control purposes. The motor phase currents, $I_a$, $I_b$ and $I_c$ as
well as angular velocity, $\omega_r$ were recorded for tests at various frequencies and voltages. For experiments at rated load, a purely resistive load consisting of bulbs of 1,000 W each was used. At maximum loaded condition wherein torque of 18 Nm is obtained while supplying current of 11.5 A, the output power of DC generator was observed to be 3.2 kW.

### 4.2.4 Results and Discussions

The optimization routine was adopted for variable frequencies and voltages under various loading conditions. The results of identified parameters generated by IPSO for various conditions by using the improved model are provided in Table 4.2.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>10Hz/47 V</th>
<th>20Hz/85V</th>
<th>40Hz/100V</th>
<th>60Hz/240V Model 2</th>
<th>Measured 60Hz/240V</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R(\Omega)$ (DC test)</td>
<td>0.6</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$L_d$ (mH)</td>
<td>37.67</td>
<td>42.2</td>
<td>43.2</td>
<td>18.93</td>
<td>25.18</td>
</tr>
<tr>
<td>$L_q$ (mH)</td>
<td>54.98</td>
<td>43.6</td>
<td>50</td>
<td>35.15</td>
<td>41.96</td>
</tr>
<tr>
<td>$L_{nd}$ (mH)</td>
<td>35.42</td>
<td>37</td>
<td>37.8</td>
<td>15.7</td>
<td>22.88</td>
</tr>
<tr>
<td>$L_{mq}$ (mH)</td>
<td>50.36</td>
<td>37.4</td>
<td>43.4</td>
<td>32.7</td>
<td>36.66</td>
</tr>
<tr>
<td>$L_{kd}$ (mH)</td>
<td>36</td>
<td>35.8</td>
<td>39.4</td>
<td>18</td>
<td>27</td>
</tr>
<tr>
<td>$L_{kq}$ (mH)</td>
<td>50</td>
<td>25.5</td>
<td>49.5</td>
<td>35</td>
<td>58</td>
</tr>
<tr>
<td>$R'<em>{kd}=R'</em>{kq}$ ((\Omega))</td>
<td>0.37</td>
<td>0.41</td>
<td>0.41</td>
<td>0.45</td>
<td>0.4</td>
</tr>
</tbody>
</table>

The measured values that are provided in [15] for 60 Hz, 240 V supply are also shown for comparison. For validation of the identified parameters at various conditions, the behavior of currents in the model simulated with identified parameters has been compared to that of the measured currents. The validation figures have been provided for supply frequency of 20 Hz with a supply voltage of 85 V in Figure 4.8 and also for a supply voltage of 240 V with a frequency of 60 Hz.
Figure 4.8. Model validation for 20Hz, 85V supply. (a) Measured phase currents $I_a$, $I_b$ and $I_c$. (b) Calculated phase currents $I_a$, $I_b$ and $I_c$ using identified model.

The current, torque and speed behavior of the motor has been given for 60 Hz line–start condition. In Figure 4.9(a) to 4.9(c), the measured three–phase currents, calculated three–phase currents using conventional model and calculated three–phase currents using improved model are shown respectively for full load condition. Table 4.3 shows the comparison of parameters identified using conventional and improved model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>60Hz/ 240 V Model 1</th>
<th>60Hz/ 240V Model 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R$ ($\Omega$) (DC test)</td>
<td>0.6</td>
<td></td>
</tr>
<tr>
<td>$L_d$ (mH)</td>
<td>25.93</td>
<td>18.93</td>
</tr>
<tr>
<td>$L_{sd}$ (mH)</td>
<td>38.35</td>
<td>35.15</td>
</tr>
<tr>
<td>$L_{md}$ (mH)</td>
<td>24.55</td>
<td>15.7</td>
</tr>
<tr>
<td>$L_{sq}$ (mH)</td>
<td>35.96</td>
<td>32.7</td>
</tr>
<tr>
<td>$L_{kd}$ (mH)</td>
<td>18</td>
<td>18</td>
</tr>
<tr>
<td>$L_{kq}$ (mH)</td>
<td>35</td>
<td>35</td>
</tr>
<tr>
<td>$R'<em>{kd}=R'</em>{kq}$ ($\Omega$)</td>
<td>0.42</td>
<td>0.45</td>
</tr>
</tbody>
</table>
The model used in Figure 4.9(b) (Model 1) is a conventional model that does not incorporate saturation effects and the model used in Figure 4.9(c) (Model 2) is the improved model. It can be seen that the transient times and steady state current magnitudes in Figure 4.9(a) and Figure 4.9(c) are similar to each other when comparing Figure 4.9(a) and Figure 4.9(b).

![Figure 4.9. Model validation for 60 Hz, 240V supply. (a) Measured phase currents $I_a$, $I_b$ and $I_c$. (b) Calculated phase currents $I_a$, $I_b$ and $I_c$ using identified model without saturation (conventional) (c) Calculated phase currents $I_a$, $I_b$ and $I_c$ using identified model with improved model accounting for saturation (developed model).](image-url)

This is because the improved model has reduced inductances owing to saturation. This was achieved by deriving the decrease in inductance due to increased current from experimental
tests. The conventional model does not take into account this change in inductance, thus obtaining a faster synchronism. In Figure 4.10, the calculated phase A current, torque and speed behavior of the machine have been provided for 60 Hz, 1,800 rpm under full load condition for models 1 and 2.

![Figure 4.10. Calculated results for identified model for 60Hz, 240 V supply. (a) Torque and speed characteristics. (b) Calculated phase current $I_a$. Model 1: Conventional model without saturation, Model 2: Improved model incorporating saturation](image)

It can be seen that the start–up performance of the LSIPMSM identified using Model 1 shows a large magnetizing inductance.

4.2.5 Conclusions on Off–line Parameter Determination

This section proposes an innovative technique to determine the parameters of a LSIPMSM by using experimental measurements in conjunction with an improved model that incorporates saturation characteristics in the magnetizing inductances. When the flux differs from rated level, the magnetizing inductance changes due to the non–linear behavior
of the iron core and it is seen that the improved model depicts the motor in a better manner, hence achieving a more accurate representation of the motor. The non–linear model of the laboratory LSIPMSM has been identified as a function of the phase currents and rotor position with a constrained optimization procedure using IPSO algorithm. Even though it has been concluded that the identification procedure was able to represent the motor closely with its actual performance, starting and steady–state currents in the identified model still deviate in magnitude when compared to the experimental values. This could be attributed to the change in magnet operating temperature and iron losses in the stator [77].

In the next section, an on–line parameter determination approach is proposed that takes into account saturation as well as magnet temperature variations. The effect of core loss on the parameter determination is also studied.

4.3 On–line Parameter Identification based on Multi–Parameter Estimation Considering Iron Losses

In this section, an on–line parameter determination method is proposed for IPMSM to consider the variations in inductances due to saturation, stator resistance and PM flux due to temperature as well as the effect of iron losses in the steel on these parameters.

4.3.1 Equivalent Circuit Modeling of PMSM Incorporating Iron Losses for On–line Identification of Parameters

The $d$– and $q$– axis equivalent circuit of an IPMSM incorporating iron losses is shown in Figure. 4.11, where $v_{ds}$, $v_{qs}$, $i_{ds}$, and $i_{qs}$ are the $d$– and $q$–axis voltages and currents respectively, $R_s$ is the stator resistance; $L_{ds}$ and $L_{qs}$ are the $d$– and $q$–axis inductances; $\lambda_{PM}$ is the magnet flux linkage; $i_{md}$ and $i_{mq}$ are the magnetizing currents. $R_i$ is the iron loss resistance used to depict the eddy current losses in the stator core [78].
The currents flowing through \( R_l \) are represented as \( i_{id} \) and \( i_{iq} \). The iron loss resistance in the parallel circuit depends on electrical angular velocity, \( \omega_e \) and flux linkages in the \( d- \) and \( q- \) axis. The dynamic equations used for solving the IPMSM model considering iron loss are expressed as in (4.10)– (4.13).

\[
\begin{align*}
\begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix} &= \begin{bmatrix} R_e + pL_{ds} & -\omega_e L_{qs} + \frac{\omega_e L_{ds} L_{qs}}{R_i} \\ \omega_e L_{ds} - p \frac{\omega_e L_{ds} L_{qs}}{R_i} & R_e + pL_{qs} \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + \begin{bmatrix} \frac{\omega_e^2 L_{qs} \lambda_{PM}}{R_i} \\ \omega_e \lambda_{PM} \end{bmatrix} \\
\omega_e(L_{qs} i_{mq} + L_{ds} i_{mq}) \\
\end{align*}
\]  

where \( p = \frac{d}{dt} \) and \( R_e = R_i + \left( \frac{\omega_e^2 L_{ds} L_{qs}}{R_i} \right) \)

\[
T_e = 3P/4 \left[ (L_d - L_q) i_{md} i_{mq} + \lambda_{PM} i_{mq} \right] 
\]

\[
p\omega_r = \frac{P}{2J} \left( T_e - T_l \right) 
\]

Figure 4.11. \( d- \) and \( q- \) axis equivalent circuit model of IPMSM incorporating iron loss resistance. (a) \( d- \) axis model with iron loss. (b) \( q- \) axis model with iron loss.

where \( P \) is the number of poles, \( J \) is the moment of inertia, and \( T_l \) is the load torque.
4.3.2 Determination of Iron Loss Resistance to Incorporate Iron Loss in Identification

In Chapter 3, iron loss resistance determined from no–load tests performed with $i_d=0$ control was used for calculating magnetizing currents and the fundamental iron losses during operation. The rated inductances were used to perform the loss calculations and analyze iron losses. However, in this study, the aim is to identify the parameters considering the iron loss effect. Thus, an alternate iron loss resistance method which does not require the information of the equivalent circuit parameters has been utilized. From the no–load tests with magnetized rotor, the sum of iron losses and mechanical losses of the IPMSM can be obtained after subtracting the copper losses from the input power. The mechanical losses calculated using Appendix B are used to segregate the iron and mechanical losses. Thus, the iron loss resistance calculation for the same IPMSM without the knowledge of $L_{ds}$, $L_{qs}$ and $\lambda_{PM}$ can be performed by dividing the back–EMF voltage by the iron loss value. The values from method 2, given in Figure 4.12, has been used in the controller as a 2–D lookup table.

![Graph showing iron loss resistance values vs. speed.](image)

Fig. 4.12. Iron loss resistance values for varying speeds from method 2 tests conducted on test IPMSM.
4.3.3 On–line Multi–Parameter Determination through Two– Stage RLS Estimation Algorithm

The on–line determination of the resistive and magnetic parameters of the PMSM has been performed using (4.13) and (4.14).

\[ v_{ds} = R_s i_{ds} - \omega_e L_{qs} i_{mq} \]
\[ v_{qs} = R_s i_{qs} + \omega_e L_{ds} i_{md} + \omega_e \lambda_{PM} \] (4.13)

\[ i_{ds} = \frac{1}{R_s} \left( i_{md} - \omega_e L_{qs} i_{mq} \right) \]
\[ i_{qs} = \frac{1}{R_s} \left( i_{mq} + \omega_e L_{ds} i_{md} + \omega_e \lambda_{PM} \right) \] (4.14)

The parameters have been determined in two stages: stage 1 and stage 2, from the current, voltage and position measurements.

A. Stage 1: Determination of \( L_{ds} \) and \( L_{qs} \)

In stage 1, the values of magnetic parameters, \( L_{ds} \) and \( L_{qs} \) have been determined by using \( R_s \), \( R_i \) and \( \lambda_{PM} \) as known initial values from prior off–line experimental parameter determination tests. The \( d \)– and \( q \)–axes inductance estimation has been performed by using calculated values of \( i_{md} \) and \( i_{mq} \). From (4.15) and (4.16), the \( \omega_e L_{qs} i_{mq} \) and \( \omega_e L_{ds} i_{md} \) terms can be written as:

\[ \omega_e L_{qs} i_{mq} = R_s i_{ds} - v_{ds} \] (4.15)
\[ \omega_e L_{ds} i_{md} = v_{qs} - R_s i_{qs} - \omega_e \lambda_{PM} \] (4.16)

The values of \( i_{md} \) and \( i_{mq} \) have to be calculated as they cannot be measured as terminal quantities. Hence \( i_{md} \) and \( i_{mq} \) are determined using the measured terminal values, \( v_{ds} \), \( v_{qs} \), \( i_{ds} \), and \( i_{qs} \), by substitution in (4.17).
\[ i_{md} = i_{ds} - \frac{v_{ds}}{R_s} - \frac{1}{R_i} i_{ds} \]
\[ i_{mq} = i_{qs} - \frac{v_{qs}}{R_s} - \frac{1}{R_i} i_{qs} \]  
\hfill (4.17)

The value of \( R_s \) is obtained from temperature measurements and will be explained in the next section whereas \( R_i \) is obtained off-line and stored in a lookup table. Finally, the values of \( L_{ds} \) and \( L_{qs} \) are calculated using (4.18) and (4.19).

\[ L_{ds} = \frac{v_{qs} - R_i i_{qs} - \omega \lambda_{PM}}{\omega_c \left( i_{ds} - \frac{v_{ds}}{R_s} + \frac{R_s}{R_i} i_{ds} \right)} \]  
\hfill (4.18)

\[ L_{qs} = \frac{R_i i_{qs} - v_{ds}}{\omega_c \left( i_{qs} - \frac{v_{qs}}{R_s} + \frac{R_s}{R_i} i_{qs} \right)} \]  
\hfill (4.19)

B. Stage 2: Determination of \( R_s \) and \( \lambda_{PM} \)

Once the values of \( L_{ds} \) and \( L_{qs} \) have been determined, the values of \( R_s \) and \( \lambda_{PM} \) are recalculated from the initial values in the second stage of identification problem. In this case, the values of \( R_s \) and \( \lambda_{PM} \) are calculated and the updated parameter set is provided to the controller and further continuing stage 1 of estimation. This method of estimating \( R_s \) through temperature provides accurate estimation results [18]. In this stage, the estimation of \( R_s \) is performed as a linear function of winding temperature and updated periodically using (4.20), where \( R_{s0} \) is the initial value of stator resistance measured at room temperature (considered as 25°C), and \( \alpha \) is the temperature coefficient of winding resistance.

\[ R = R_{s0} \left[ 1 + \alpha (T - T_0) \right] \]  
\hfill (4.20)

Thermocouples have been embedded in the phase windings of the test motor to obtain the measurements through a temperature measuring unit (TMU).
Using the periodically updated values of $R_s$, the magnetizing currents, $i_{md}$ and $i_{mq}$ are recalculated according to (4.17). Hence, the value of $\lambda_{PM}$ can be calculated from (4.21).

$$\lambda_{PM} = \frac{v_{qs} - R_s i_{qs} - \omega_e L_{ds} i_{md}}{\omega_e}$$  \hspace{1cm} (4.21)

C. Employment of Two–step RLS Estimation towards Multi–parameter Estimation in Investigated IPMSM

This section employs RLS algorithm to solve the two stages of identification problems due to its simplicity and robustness and wide suitability in case of identification problems [20]. The RLS equations used for implementing stage 1 and stage 2 of identification are given in (4.22).

\[ \begin{align*}
\theta_{(k)} &= \theta_{(k-1)} + K_{(k)} \ast \epsilon_{(k)} \\
\epsilon_{(k)} &= y_{(k)} - \phi_{(k)}^T \ast \theta_{(k-1)} \\
K_{(k)} &= P_{(k-1)} \ast \phi_{(k)} \left[ \lambda I_d + \phi_{(k)}^T P_{(k-1)} \phi_{(k)} \right]^{-1} \\
P_{(k)} &= \left[ I - K_{(k)} \ast \phi_{(k)} \right] \ast \frac{P_{(k-1)}}{\lambda} \end{align*} \] \hspace{1cm} (4.22)

where $y$ is the output matrix, $\theta_{est}$ is the estimated parameter vector, $\phi$ is the feedback matrix, $\lambda$ is the forgetting factor, $I$ is the identity matrix, $\epsilon$ is the estimation error, and $K$ and $P_{(k)}$ are correction gain matrices. The machine inductances that are a function of $d$–and $q$–axis current magnitudes and hence, level of saturation in the iron, change dynamically and have been included in the first step of identification that has a higher sampling rate. The estimation of $L_{ds}$ and $L_{qs}$ is performed using RLS according to (4.23).

\[ \begin{align*}
\phi &= \begin{bmatrix} 0 & -\omega_e i_{mq} \end{bmatrix}^T, \quad \theta = \begin{bmatrix} L_{ds} \\ L_{qs} \end{bmatrix} \\
\epsilon &= y - \phi^T \theta_{(k-1)} \\
K &= \left[ I - \phi \theta_{(k-1)} \right] \ast \frac{P_{(k-1)}}{\lambda} \\
P &= \left[ I - K \phi \right] \ast \frac{P_{(k-1)}}{\lambda}
\end{align*} \] \hspace{1cm} (4.23)
The currents \( i_{md} \) and \( i_{mq} \) given in (4.17) are calculated simultaneously by the algorithm. The details of experimental study and related measurements will be explained in section VI. In stage 2 of the estimation, RLS estimation is performed by updating the values of \( L_{ds} \) and \( L_{qs} \) obtained from previous step. The equation for identification including \( \lambda_{PM} \) from (4.21) is performed using RLS according to (4.24).

\[
\begin{align*}
\mathbf{y} &= \left[ v_{iq}(k) - R_s \cdot i_{iq}(k) - \omega_s(k) L_{ds} \cdot i_{md} \right]; \\
\varphi^T &= \left[ \omega_s \right]; \quad \theta = \left[ \lambda_m \right]
\end{align*}
\] (4.24)

The values of \( R_s \) and \( R_i \) are updated from temperature measurements and 2–D look–up table obtained from off–line experiments respectively. In stage 2, the matrix inversion required by the RLS algorithm is of size one, which is simple to implement numerically, thus, making the RLS estimation less computationally intensive and easy to implement. Once stage 1 and stage 2 are performed, the controller is updated with the new set of parameters and the stage 1 identification implementation continues. The structure of the proposed two–step RLS algorithm is shown in Figure 4.13.

Figure. 4.13. Structure of proposed two–step RLS estimation for multi–parameter estimation incorporating iron losses
4.3.4 Mathematical Modeling and Simulation of Proposed Algorithm for Parameter Identification

The identification procedure has been implemented in a mathematical model to study the effects of considering iron losses in the multi-parameter estimation procedure at various simulated operating points and subsequently in experimental study to verify the results in real conditions. In the mathematical study, the actual value given into the machine model has been obtained from the 2-D FEA model of the IPMSM that has been prototyped in the laboratory. Table 2.2 in Chapter 2 provides details of the investigated IPMSM that was prototyped in-house. The values of inductances have been provided from 2-D FEA result as well as off-line experimental tests conducted at two current points in the test motor.

4.3.5 Results and Investigation of Identified Parameters from Mathematical Model

The two-step identification procedure has been applied in the IPMSM with MTPA control at various speeds and optimal current angles. The results of identification of parameters with and without iron losses from the mathematical study has been compared with actual value obtained from the 2-D FEA model. For stage 1, the comparison results for $I_m=15.55$ A, $\gamma=30.5^\circ$ at a speed of 600 rpm is given in Table 4.4.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>With iron loss</th>
<th>Without iron loss</th>
<th>2-D FEA</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{ds}$ (mH)</td>
<td>31.5</td>
<td>33.6</td>
<td>31.135</td>
</tr>
<tr>
<td>$L_{qs}$ (mH)</td>
<td>67</td>
<td>66.66</td>
<td>66.9</td>
</tr>
</tbody>
</table>

It is observed that the error from the actual value in case of identification without iron loss is more than the case considering iron losses. For detailed analysis, a comparison of the
identified inductances with and without iron losses is given in Figures 4.14 and 4.15 for varying $I_m$ at optimal current angles, $\gamma$ in MTPA control operation at 600 rpm. The relative error of parameters from actual value is provided in Figure 4.16. From Figures 4.14 to 4.16, it can be seen that the error in estimation is more for the identification case without iron losses, especially in higher current values, i.e., higher torque values. The error in $L_q$ is minimized when iron loss is considered whereas the error in $L_d$ is lower for lower currents but slightly higher in higher current values, for example, 7 A. This can be attributed to measurement errors in $R_c$ determination leading to lower value of $i_{ind}$ or slight error in voltage values in the model that can significantly impact the estimation.

Figure 4.14. $d$–axis inductance vs current estimation with and without iron losses through mathematical model for IPMSM and comparison with 2–D FEA.

Figure 4.15. $q$–axis inductance vs current estimation with and without iron losses through mathematical model for IPMSM and comparison with 2–D FEA.
Figure 4.16. Relative error from actual value in estimation of $d$– and $q$–axis inductances with and without iron losses using mathematical model.

4.3.6 Impact of Iron Losses at Various Operating Points

To study the impact of considering iron losses at various conditions and to perform a comprehensive analysis, the results of inductance estimation at various load torques and two points of speeds at 200 rpm and 600 rpm are provided in Figures 4.17 and 4.18. The error in estimation is given in Figures 4.19 and 4.20. Generally, the estimation at low currents and low speeds in both cases, with and without iron losses, give inaccurate values of $L_{ds}$ owing to very less value of $V_d$. However, the error is lesser for the estimation case with iron losses, for example, 20.3% and 28.9% for with and without iron loss consideration respectively. The relative error in estimation considering iron loss for $L_{qs}$ is much lesser compared to the case without iron losses, especially at higher current values. The error in $L_{ds}$ and $L_{qs}$ without considering $R_i$ is very high in lower values of current and hence will create more error in the torque command than the case considering $R_i$. The error in estimation without iron losses in case of high current and speed values is higher due a significant value of $i_{id}$ and $i_{iq}$ through the $R_i$ branch that has been neglected.
Figure 4.17. $d$–axis inductance vs current estimation from mathematical model for IPMSM with and without iron losses and comparison with actual value.

Figure 4.18. $q$–axis inductance vs current estimation from mathematical model for IPMSM with and without iron losses and comparison with actual value.

Figure 4.19. Relative error of $L_{ds}$ from mathematically obtained values of $d$–axis inductance with and without iron losses at varying load torque conditions.
4.3.7 Experimental Validation of On–line Multi–Parameter Estimation and Results

The developed on–line method was validated in the IPMSM developed in–house shown in Figure 2.1(b). For the tests in this chapter, the IPMSM was controlled by an Opal–RT real time controller with MTPA control from a PWM–fed voltage source IGBT based inverter operating at a switching frequency of 5 kHz from a DC link source of 650 VDC. The temperature is monitored using thermocouples embedded into the windings and the value of $R_s$ is updated based on (4.20). The on–line identification procedure was performed using the experimental setup as explained under varying speed and torque conditions. To study the variation in estimated values for the models with and without iron losses, the tests were conducted at low torque and high torque as well as similar speed conditions, capable of being delivered by the test motor. The measured currents, voltages and angular velocity were used in the RLS estimation algorithm considering iron losses and compared with the estimation without iron losses. The measured three–phase currents and corresponding $I_d$ and $I_q$ at $I_m = 11.16$ A, $\gamma = 20.38^\circ$ at a speed of 600 rpm for which the estimation was performed are given in Figure 4.21 and Figure 4.22 respectively.
Figure 4.21 Three–phase currents obtained from experimental identification process for $I_m=10$ A and $\gamma=20^\circ$ at a speed of 600 rpm.

The comparison results of the experimental inductance estimation with and without iron losses at $I_m=10$ A, $\gamma=20$ deg and speeds of 200 rpm and 600 rpm is given in Table 4.5.

**Table 4.5 Comparison of Experimental Inductance Estimation Results at $I_m=10$ A, $\gamma=20$ Deg and 200 RPM and 600 RPM**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Speed (rpm)</th>
<th>With iron loss</th>
<th>Without iron loss</th>
<th>Actual</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{ds}$ (mH)</td>
<td>200</td>
<td>30.02</td>
<td>26.2</td>
<td>29.5</td>
</tr>
<tr>
<td></td>
<td>600</td>
<td>25.2</td>
<td>19.95</td>
<td>28</td>
</tr>
<tr>
<td>$L_{qs}$ (mH)</td>
<td>200</td>
<td>88.52</td>
<td>89.14</td>
<td>85.06</td>
</tr>
<tr>
<td></td>
<td>600</td>
<td>81.6</td>
<td>81.8</td>
<td>82</td>
</tr>
</tbody>
</table>

Figure 4.22. Measured $I_{ds}$ and $I_{qs}$ currents for identification process at $I_m=10$ A and $\gamma=20^\circ$ at a speed of 600 rpm.
The RLS estimation result with and without iron losses for \( L_{ds} \) and \( L_{qs} \) is given in Figure 4.23 and that for PM flux linkage (\( \lambda_{PM} \)) is given in Figure 4.24. The error in identification with and without iron losses compared to actual value in case of \( L_{ds} \) is more obvious compared to the identification of \( L_{qs} \). The error in \( L_{ds} \) is higher due to the higher value of \( i_{qs} \) and hence more difference between \( i_{qs} \) and \( i_mq \) that is used for \( L_{ds} \) estimation compared to \( i_{ds} \) and \( i_{md} \) that is used for \( L_{qs} \) estimation. Thus, the higher the current values are, higher the estimation error in both \( L_{ds} \) and \( L_{qs} \) will be. In flux weakening conditions, where the value of \( i_{ds} \) is higher than constant torque region, estimation without iron losses will give significant error compared to the one with iron losses. The error is also expected to increase significantly in case of high speed machines owing to higher \( R_i \) value. In case of \( \lambda_{PM} \), the errors in case of model with iron loss resistance is slightly lesser compared to the one without iron losses. Since \( R_s \) estimation is similar in both cases, the values will be the same.

The measured temperature over time is given in Figure 4.25.

![Figure 4.23. Comparison of \( L_{ds} \) and \( L_{qs} \) obtained experimentally from stage 1 of identification at 600 rpm with and without iron losses for \( I_m = 10 \) A and \( \gamma = 20^\circ \).]
4.3.8 Conclusions on on–line parameter determination

The RLS based parameter estimation method incorporating iron losses has been used to estimate all the variable parameters of an IPMSM and eliminates the rank deficiency issue in multi–parameter estimation for IPMSMs. The effect of including iron loss resistance has been studied and the results suggest a significant difference in estimation results with and without iron losses with the former giving results close to the actual value that was obtained in 2–D FEA model of the test motor. Hence, the developed algorithm for estimation with iron losses has been validated and gives better results than the conventional model without iron losses. The results suggest that the error is more obvious at very low currents as well as high currents and hence torque. Thus, considering iron losses in estimation is an important factor, especially for EV motors with high–speed and high–torque operation.
4.4 Conclusions

The developed methods have multi–faceted applications in the overall objective of this thesis, which is to study methods of improving PM motor and drive efficiency: (1) Studying the inductance, resistance and PM flux variation in the test motor to develop parameter maps for use in control methodology study, and, (2) Study the influence of iron loss in parameter variations and hence, torque output.
CHAPTER 5
IMPROVED MAXIMUM EFFICIENCY CONTROL OF PERMANENT MAGNET SYNCHRONOUS MACHINES CONSIDERING EFFECTS OF CORE SATURATION AND TEMPERATURE VARIATION

5.1 Introduction
Considering the results obtained from previous studies on the behavior of motor and inverter losses with varying control variables, and parameter variations in a PMSM, it can be concluded that an optimized current angle control considering parameter variations can improve the system efficiency significantly compared to traditional MTPA control of a PMSM. This chapter proposes a novel method of maximizing the efficiency per current in an IPMSM through the selection of optimized phase angle during operation by considering parameter and temperature variations under load. Firstly, models have been developed for considering motor and inverter losses and the effects of parameter variations due to saturation and temperature are studied from a combination of analytical models and practical experiments. The models are then used for employing an optimal current angle search corresponding to maximum efficiency for varying operating conditions. Experimental investigations are performed on the laboratory test IPMSM for validating the developed control through interpolation of the improved look-up tables. The efficiencies have been measured at motor, inverter and system stages for varying speed, torque and temperature conditions to validate the results from the analytical model. The effectiveness of the developed method in improving system and motor efficiency is also verified and compared with conventional maximum torque per ampere (MTPA) control that considers saturation.
5.2 Development of Non–linear Model Based Efficiency Improvement Method

The goal of the proposed model–based loss minimization is to identify the current angle that corresponds to the maximum efficiency point for a speed, torque and motor operating temperature. The optimal current angle is computed off–line using an iterative procedure by considering the effects of parameter variations due to core saturation and temperature. The proposed method uses flux maps with respect to currents in the $d$– and $q$– axis to calculate the voltages and losses in the motor. Figure 5.1 depicts the flux linkage equations with respect to the currents, inductances and PM flux linkage in $d$– and $q$– axis. The saturation and cross–saturation phenomenon are represented by variations in inductances and the temperature variation in permanent magnets is represented by a change in the PM flux value. It is difficult to segregate the effects of saturation and temperature in the flux behavior. Hence, in order to incorporate both phenomenon into account, flux maps can be used in the optimization procedure.

$$
\begin{bmatrix}
\lambda_d \\
\lambda_q
\end{bmatrix} =
\begin{bmatrix}
L_d(i_d, i_q) & L_{dq}(i_d, i_q) \\
L_{qd}(i_d, i_q) & L_q(i_d, i_q)
\end{bmatrix}
\begin{bmatrix}
i_d \\
i_q
\end{bmatrix}
+ \lambda_{PM} \begin{bmatrix}
0 \\
0
\end{bmatrix}
$$

Figure 5.1. Flux linkage representation with respect to inductances and PM flux linkage depicting saturation and temperature variations effects.

The following sections explain the derivation of flux maps including saturation and cross–saturation and temperature variations in the 4.25 kW laboratory IPMSM. Furthermore, the optimization of current angle based on these properties is explained.
5.2.1 Voltage Equation Based Flux Linkage Mapping Considering Saturation and Cross–Saturation

One of the straightforward methods of calculating the flux linkage in a voltage source inverter (VSI) fed motor is by using the voltage and current values at a specific operating condition and solving $d$– and $q$– axis voltage for steady state condition [79], [80]. The flux linkage in the $d$– and $q$–axis can be calculated from the average values of the measured voltage and current commands of the VSI. The relationship between flux linkage and current is non–linear and the saturation is one of the important non–linearity to be considered while developing improved control algorithms [81]. The flux linkage considering saturation equation and cross coupling can be written as:

$$
\lambda_d(I_d, I_q) = \frac{-V_q + R_q I_q}{\omega_e}
$$

$$
\lambda_q(I_d, I_q) = \frac{V_d - R_d I_d}{\omega_e}
$$

In order to take saturation and cross coupling into account and derive flux linkage maps for the entire operating region, a polynomial function of flux linkage according to (5.1) can been used. A non–linear least squares regression function, which produces good estimates of the unknown parameters in the model with relatively small data sets can be used to fit the values in (5.2) [80].

$$
\lambda_d(I_d, I_q) = a_0 + a_1 I_d + a_2 I_q + a_3 I_d I_q + a_4 I_d^2 + a_5 I_q^2
$$

$$
\lambda_q(I_d, I_q) = b_0 + b_1 I_d + b_2 I_q + b_3 I_d I_q + b_4 I_d^2 + b_5 I_q^2
$$

(5.2)

where $a_0$–$a_5$ and $b_0$–$b_5$ represent the coefficients of flux linkage in the $d$– and $q$– axis respectively; $a_0$ is the PM flux linkage. Experimental method of flux linkage calculation has been developed. Figure 5.2 shows the block diagram of the test setup and the required
measurements and Figure 5.3 shows the sampling of measured signals that is used for the calculation. The tests were conducted for 5 values of $i_q$ and 3 values of $i_d$, giving a total of 15 current values.

![Block diagram representing test method for determination of flux maps experimentally.](image)

Figure 5.2. Block diagram representing test method for determination of flux maps experimentally.

![Procedure and measurements of current control used for flux linkage fitting](image)

Figure 5.3. Procedure and measurements of current control used for flux linkage fitting

The generalized process of obtaining the flux linkage map follows the given steps:

1) Choose operating points for $d$- and $q$-axis currents sufficient to solve (5.2). The operating points for current are chosen in such a way that the entire region within current limit circle is covered and the accuracy of voltage measurement is high. A speed close to and lower than base speed is preferred to reduce the influence of stator resistance in the calculations and obtain more accurate command voltages.
2) Perform tests at selected speeds by keeping \(i_d\) constant and sweeping \(i_q\) over selected values and repeat for all selected values of \(i_d\). Simultaneously obtain the \(v_d, v_q, i_d, i_q, \) and \(\omega\) values from measurements.

3) Calculate average values of the measured values. The controllers try to maintain desired operating point continuously; hence, an acceptable time window of 1 s has been used for averaging.

4) Using (5.2) and a non–linear least squares regression function, the flux linkage map can be obtained.

The voltage measurements have been taken from the command voltage given to the VSI. It is well known that owing to the non–linearity of the inverter, the controller command voltage do not match the measured voltage exactly [82]. Therefore, average distortion compensation voltages \(V_{dq, err}\) are used for compensating the non–linearity in \(d–\) and \(q–\)axis voltages according to (5.3) [82].

\[
\begin{align*}
V_d^* &= V_d + V_{d, err}^*; \quad V_{d, err}^* = D_d V_{dead} \\
V_q^* &= V_q + V_{q, err}^*; \quad V_{q, err}^* = D_q V_{dead}
\end{align*}
\]  

(5.3)

where, \(D_d\) and \(D_q\) are periodical functions of rotor position, \(\theta\) and \(\gamma\) given in (5.4) and are constants for constant \(\gamma\). The distorted voltage, \(V_{dead}\) is 0.5 V for the test inverter.

\[
\begin{align*}
D_d &= 2\sin\left[\theta - \text{int}\left\{3\left(\theta + \gamma + \frac{\pi}{6}\right)/\pi\right\} \times \frac{\pi}{3}\right] \\
D_q &= 2\cos\left[\theta - \text{int}\left\{3\left(\theta + \gamma + \frac{\pi}{6}\right)/\pi\right\} \times \frac{\pi}{3}\right]
\end{align*}
\]  

(5.4)

The flux linkage coefficients derived using the aforementioned procedure in the \(d–\) and \(q–\)axis is given in Table 5.1. Figures 5.4(a) and 5.4(b) show the \(d–\) axis and \(q–\)axis flux linkage maps respectively obtained for varying currents at room temperature.
TABLE 5.1 FLUX LINKAGE COEFFICIENTS AT 25 °C

<table>
<thead>
<tr>
<th></th>
<th>( \lambda_d )</th>
<th>( \lambda_{q} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( a_0 )</td>
<td>0.6632</td>
<td>( b_0 ) 0.3794</td>
</tr>
<tr>
<td>( a_1 )</td>
<td>-0.00246</td>
<td>( b_1 ) -0.0027</td>
</tr>
<tr>
<td>( a_2 )</td>
<td>0.00929</td>
<td>( b_2 ) 0.0357</td>
</tr>
<tr>
<td>( a_3 )</td>
<td>-7.216e-6</td>
<td>( b_3 ) -0.00028</td>
</tr>
<tr>
<td>( a_4 )</td>
<td>-0.00139</td>
<td>( b_4 ) 1.929e-05</td>
</tr>
<tr>
<td>( a_5 )</td>
<td>0.000119</td>
<td>( b_5 ) 0.00098</td>
</tr>
</tbody>
</table>

Figure 5.4. Flux linkage maps obtained in the \( d- \) and \( q- \) axis for room temperature. (a) \( d- \) axis flux linkage. (b) \( q- \) axis flux linkage.
5.2.2 Flux Linkage Considering PM Flux Variation Due to Temperature

During machine operation, machine losses cause an increase in the permanent magnet temperature. This reduces the PM flux linkage, according to the properties of PM [83]. The linear thermal model of PM flux linkage can be denoted as (5.5) [83].

\[
\lambda_{PM(T2)} = \lambda_{PM(T1)} \left(1 + \beta \Delta T\right)
\]

where \(T_1\) and \(\lambda_{PM(T1)}\) denote permanent magnet temperature and PM flux linkage at room temperature, considered as 25°C, \(T_2\) is the temperature at which the PM flux linkage is calculated, \(\lambda_{PM(T2)}\) denotes the PM flux at \(T_2\), \(\Delta T\) is the difference between \(T_2\) and \(T_1\), and \(\beta\) is the PM flux thermal coefficient. For the NdFeB magnet material considered in the analytical studies as well as experimental investigations in this thesis, \(\beta\) is about \(-0.12\%/°C\) [84], [85]. To understand the behavior of reduction of PM flux in the motor with temperature, preliminary experiments were conducted on the IPMSM. A back–EMF test was initially conducted before loading to determine the PM flux linkage at room temperature. The motor was loaded to increase the temperature. The stator winding temperature was measured using RTDs connected to a data acquisition system and the rotor temperature was measured using a thermal imager. The motor was prototyped in such a way that the magnets can be accesses through holes on the motor housing for the thermal imager to capture the temperatures. Figure 5.5 shows some of the images captured at varying operating temperatures. The back–EMF tests were repeated at various magnet operating temperatures and the PM flux linkage was calculated. Figure 5.6 shows the PM flux linkage variation with temperature. The variation affects torque output of the motor, thus affecting the efficiency. In this work, the variation of PM flux with temperature is considered by conducting the flux linkage determination technique used in section 5.2.1 at
a higher operating temperature. The stator resistance used to calculate the flux linkages using $d$– and $q$– axis voltages in (5.1) increases with temperature and was also updated based on stator winding temperature measurement according to (5.6).

$$R_{s(T_f)} = R_{s(T_i)} \left(1 + \alpha \left(T_f - T_i\right)\right)$$  \hspace{1cm} (5.6)

Figure 5.5. Permanent magnet temperature measured using thermal image camera at various operating conditions. (a) Room temperature. (b) Magnet temperature measured at 47ºC. (b) Magnet temperature measured at 60ºC.

Figure 5.7 shows the measurement of stator temperature and corresponding increase in stator resistance that was considered in the flux linkage determination.

Figure 5.6. Permanent magnet flux variation with temperature.
5.3 Implementation of Optimal Current Angle Computation

This section explains the methodology of derivation of optimal current angle corresponding to maximum system efficiency. Initially, the overall methodology is explained, followed by the motor and inverter loss models that are used in deriving the optimal current angle.

5.3.1 Overview of Current Angle Derivation

The procedure of deriving the optimal current angle for a given speed and peak current at temperature $T_1$, considered as the room temperature is given in Figure 5.8. The current angle is derived for various speeds and a look–up table can be generated for various loads and speeds. The procedure is repeated with flux linkage maps obtained at operating temperature, $T_2$, to derive the optimal current angles for varying loads and speeds at $T_2$, which is considered as 85°C in this study. Likewise, the optimal current angles can be derived for other temperatures. The operating temperature is defined as the stator temperature, for which the measurements are non–invasive and easily available. There is a difference between the rotor temperature and stator temperature; however, two temperatures can be assumed to be varying linearly, thus the stator and rotor temperatures
can be assumed to be always proportional. In the next sub-section, derivation of loss models of the motor and inverter are elaborated.

![Diagram](image)

**Figure 5.8. Procedure of deriving optimal current angle for a specific speed and peak current at room temperature, \( T_1 \).**

### 5.3.2  Loss Models for Inverter and Motor

The losses in the inverter can be defined as switching losses in the IGBT and conduction losses in the IGBT and freewheeling diodes [62], [66]. The equations for switching losses, \( P_{\text{inv,sw}} \) and conduction losses \( P_{\text{inv,con}} \) have been used from (3.1) and (3.2) in chapter 3. The switching losses can be rewritten as (5.7) in terms of turn on and turn off energies. The conduction losses can be rewritten as (5.8) and (5.9) to represent IGBT and diode conduction losses with respect to \( I_m \).

\[
P_{\text{inv,sw}} = \frac{6}{\pi} \left( T_{\text{on}} + T_{\text{off}} \right) f_s V_{dc} I_s
\]  

(5.7)
The modulation index, \( M_i \), can be calculated using the \( \nu_d \) and \( \nu_q \) using (5.10), and the power factor, \( \cos \theta \), can be calculated using (5.11) from load angle calculations for a fixed \( \gamma \). The expression of conduction losses with respect to the power factor ensures taking the effect of varying current angle into account in loss calculations.

\[
M_s = \frac{2V_{i_m}}{V_{dc}} = \frac{2\left(\sqrt{\nu_d^2 + \nu_q^2}\right)}{V_{dc}} \tag{5.10}
\]

\[
\theta = \delta + \gamma; \quad \delta = \arctan \left(\frac{\nu_d}{\nu_q}\right) \tag{5.11}
\]

The total inverter losses, \( P_{loss, inverter} \), is an addition of the losses given in (5.7) to (5.9).

To calculate the motor losses and output power, the equivalent circuit for IPMSM has been used. For the current angle derivations and comparisons with conventional MTPA technique, only fundamental copper and iron losses and harmonic iron losses have been considered. The switching frequency is kept at 12 kHz, subsequently, magnet eddy current losses and harmonic copper losses can be kept minimum and are ignored. The equivalent circuit used in Chapter 4 can be redrawn in Figure 5.9 to represent the flux linkage. The \( R_s \) determines the copper losses and \( R_i \) determines the iron losses and the magnetizing currents, \( i_{md} \) and \( i_{mq} \). The \( R_s \) is updated according to the operating temperature using the relationship in (5.6) and the \( R_i \) value is calculated according to (5.12).

\[
P_{Cu(T)} = 3I_m^2 R_{s(T)} = \frac{3}{2} \left( i_d^2 + i_q^2 \right) R_{s(T)} \tag{5.12}
\]
Figure 5.9. $d$– and $q$–axis equivalent circuit model of IPMSM incorporating iron loss resistance with respect to flux linkage. (a) $d$–axis model with iron loss. (b) $q$–axis model with iron loss.

The $R_i$ value represents the iron losses in the equivalent circuit and can be calculated as a function of the air–gap voltage and iron loss power calculated using the flux linkages in the $d$– and $q$–axis. The iron losses $P_{Fe,f}$ depends on the time variation of flux density in stator teeth and yoke and thus, the magnitude of peak stator core flux density, $B_m$ according to (5.13) and (5.14). The losses can also be defined in terms of the teeth and yoke losses in $d$– and $q$–axis, $P_{Fe,dq}$ and $P_{Fe,y,dq}$ and corresponding volumes, $V_t$ and $V_y$ as in (5.15) [15].

$$dP_{Fe,f} = K_{eddy} f^2 B_m^2 + K_{Hys} fB_m^x$$

(5.13)

$K_{Eddy}$ and $K_{Hys}$ are the eddy current and hysteresis loss coefficients, $x$ and $\chi$ are the constants dependent on core material property, $d$ is lamination thickness, and $f$ is the fundamental frequency and $V$ is the volume of the core.

$$K_{eddy} = \frac{V\pi^2 d^2}{6\rho}$$

(5.14)

$$P_{Fe,f} = dP_{Fe,dq} V_t + dP_{Fe,y,dq} V_y$$

(5.15)
In (5.16), the $d$– and $q$–axis flux density in teeth, $B_{t,dq}$ and yoke, $B_{y,dq}$ corresponding to the $d$– and $q$–axis flux linkages $\lambda_{dq}$ are written as a function of the corresponding teeth and yoke areas, $A_t$ and $A_y$.

$$B_{t,dq} = \frac{\lambda_{dq}(i_d,i_q,T)}{A_t}; \quad B_{y,dq} = \frac{\lambda_{dq}(i_d,i_q,T)}{A_y}$$

(5.16)

Thus, by substituting the flux density in (5.13) and deriving the equations in $d$– and $q$–axis, updated iron loss equation with respect to the flux linkages in $d$– and $q$–axis can be written from (5.13)– (5.16) in (5.17) to calculate the fundamental iron losses, $P_{Fe,f}$. The iron loss resistance can be derived as a function of $P_{Fe,f}$ as (5.18). The updated coefficients, $k_{EC}$ and $k_{Hy}$ have been determined from 2–D FEA model of the IPMSM as 0.3 and 0.015.

$$P_{Fe,f} = \frac{3}{2} \left[ k_{EC} f^2 \left( \lambda_d \left( i_d,i_q,T \right)^2 + \lambda_q \left( i_d,i_q,T \right)^2 \right) + k_{Hy} f \left( \lambda_d \left( i_d,i_q,T \right)^2 + \lambda_q \left( i_d,i_q,T \right)^2 \right) \right]$$

$$k_{EC} = \frac{V \pi^2 d^2}{6 \rho} \left( \frac{V_t}{A_t} + \frac{V_y}{A_y} \right); \quad k_{Hy} = \chi \left( \frac{V_t}{A_t} + \frac{V_y}{A_y} \right)$$

(5.17)

$$R_i = \frac{3}{2} \left( \frac{\omega^2}{P_{Fe,f}} \right)$$

(5.18)

The calculated $R_i$ can be used to calculate the magnetizing currents that can be subsequently used to calculate the electromagnetic output torque in the next subsection. The harmonic iron loss calculations from Chapter 2 have been incorporated in the loss calculations for each loading condition. In addition to the controllable electrical losses, mechanical losses, $P_{mech}$, are used in calculating total motor losses to compare with actual experimental values. The mechanical losses were calculated in the no–load test for varying speeds with non–magnetized rotor as explained in Appendix B. Thus, total motor losses, $P_{loss,\text{motor}}$ can be written as in (5.19).
\[ P_{\text{loss,motor}} = P_{\text{Cu}(T)} + P_{\text{Fe},f} + P_{\text{Fe},h} + P_{\text{mech}} \quad (5.19) \]

### 5.3.3 System Efficiency Computation for Optimal Current Angle Derivation

The system efficiency can be defined as the percentage of mechanical output power divided by the DC input power at the inverter in (5.20).

\[ \eta_{\text{system}} = \left( \frac{P_{\text{out,mechanical}}}{P_{\text{in,DC}}} \right) \times 100\% \quad (5.20) \]

The output power can be derived as a function of the electromagnetic torque and speed, whereas the DC input power is rewritten in terms of the motor and inverter losses in (5.21).

\[ \eta_{\text{system}} = \frac{1.5 \times T_e \omega \left( \frac{2\pi}{60} \right)}{1.5 \times T_e \omega \left( \frac{2\pi}{60} \right) + P_{\text{loss,motor}} + P_{\text{loss,inverter}}} \times 100\% \quad (5.21) \]

The electromagnetic output torque considering iron losses can be derived in (5.22).

\[ T_e = \frac{3P}{4} \left[ \left( \lambda_d \left( i_d, i_q, T \right) i_{mq} - \lambda_q \left( i_d, i_q, T \right) i_{md} \right) \right] \quad (5.22) \]

The magnetizing currents, \( i_{md} \) and \( i_{mq} \) used in \( T_e \) derivation can be derived using \( R_i \) as (5.23).

\[ i_{md} = i_{dk} \frac{\left( v_{dq} - R_i i_{dq} \right)}{R_y} \quad i_{mq} = i_{qs} \frac{\left( v_{dq} - R_i i_{qs} \right)}{R_y} \quad (5.23) \]

Hence, using (5.23) in (5.22) and then in (5.21), the system efficiency can be derived from the \( d \)- and \( q \)-axis stator resistance, iron loss resistance, magnetizing currents and flux linkages. For every speed and peak current, the current angle is swept until the system efficiency is maximum according to search process in Figure 5.8. The analytical results from the developed method for varying torque–speed conditions are given in section 5.4.
5.4 Analytical Results from the Developed Method

The procedure selects the current angle, $\gamma_{MEPA}$ that corresponds to the maximum system efficiency defined as maximum efficiency per ampere (MEPA) angle. The test motor was the 4.25 kW IPMSM introduced in Chapter 2. For a given speed and load peak current, an optimum current angle reference is computed based on the search–based approach that takes into account the corresponding system losses. By repeating this procedure for all $n$ points between 0 and $I_{\text{max}}$, a 1–by–$n$ look–up table can be obtained for the MEPA approach. The MEPA flux depends on the motor temperature as well as the motor speed. The calculations have been repeated for the $\gamma_{MEPA}$ computation procedure for several speed values and to build a 2–D LUT valid for a specified temperature. In this case, the MEPA angle is obtained as an interpolation according to the load demand and rotor speed. This study used load–to–current angle LUTs of 16 points ($n=16$) for an operating speed. The number of speed points were selected as 6, from 175 rpm to 675 rpm with an interval of 100 rpm. Several 2–D LUTs can be obtained for different motor operating temperatures.

5.4.1 Results of Motor and System Efficiency Under Varying Operating Conditions

The developed search based procedure has been implemented on the test motor for various load and speed points. The current angle corresponding to maximum system efficiency, $\gamma_{MEPA}$ is shown for varying speeds and loads in Figure 5.10. It can be seen that the as current and speed increase, the current angle values also increase. The increase due to speed is due to the increase in core losses and the increase due to current is due to the inverter losses as well as the core and copper losses. Figure 5.11(a) shows the variation of system and 5.11(b) shows the motor efficiencies with respect to the current angle for varying speeds and $I_m=2A$. It can be seen that the motor and system efficiencies peak at different angles.
Figure 5.12(a) shows the efficiency map obtained for the various speed and load points. Figure 5.12(b) shows the 2–D plot of system efficiency variation with respect to $I_m$ for varying speed points. Figure 5.13(a) and 5.13(b) show the efficiency map and 2–D plot of motor efficiency variation with respect to $I_m$ for varying speed points.

![Graph showing system efficiency variation](image1)

Figure 5.10. Simulated values of current angle, $\gamma_{MEPA}$ with respect to peak current and speed.

![Graph showing simulated values](image2)

Figure 5.11. Simulated values of efficiency variation with current angle showing maximum efficiency angles for varying speeds and $I_m = 2$ A. (a) System efficiency. (b) Motor efficiency.
Figure 5.12. System efficiency corresponding to simulated values of current angle, $\gamma_{MEPA}$ in 3–D and 2–D forms. (a) Surface map of system efficiency. (b) System efficiency with respect to $I_m$ for various speeds.

5.5 Experimental Validation using Laboratory PMSM

The proposed maximum efficiency detection method has been tested and validated on a laboratory IPMSM motor–drive system. The current angles obtained off–line were used at the corresponding load–speed points in the control diagram at $T_1^\circ C$ shown in Figure 5.14. The efficiency at $\gamma_{MEPA}$ loading point was measured using a power analyzer and torque transducer for which the details are given in sub-section 5.5.1. The setup and experimental
validation of the developed method under varying speed and load conditions are elaborated in subsequent subsections.

Figure 5.13. Motor efficiency corresponding to simulated values of current angle, $\gamma_{MEPA,m}$ in 3–D and 2–D forms. (a) Surface map of motor efficiency. (b) Motor efficiency with respect to $I_m$ for various speeds.

Figure 5.14. Control diagram for implementation of the developed maximum efficiency method in laboratory IPMSM drive.

110
5.5.1 Experimental Setup

The 4.25 kW test motor for which the details are given in Chapter 2 is under torque/current control mode controlled using a field programmable gate array (FPGA) based real–time simulator and is driven by a dynamometer which is in speed control mode. The IGBT inverter is sine PWM modulated and has a constant switching frequency of 12 kHz. The DC link voltage is set a constant at 650 V. For the speed and position sensing, a high–resolution encoder with a resolution of 2,500 cycles–per–revolution has been used. The current sensor used for DC power measurement has nominal current of 12 A and accuracy of ±0.084 A at 25°C. The DC voltage sensor is a closed loop Hall Effect current transducer with an accuracy of ±0.8% at 25°C and can measure up to 1,000 V. The current sensors for AC current measurements have a nominal rms current of 25 A with a resolution of ±0.2% at 25°C. A power analyzer was used for current/voltage measurements at the DC link and the three–phase motor terminals for validation of the proposed method. A torque transducer was used at the shaft between the test IPMSM and dynamometer to measure output power and validate the developed method and efficiency calculations. The accuracy of AC power measurement of the power analyzer is ±0.04% of the reading value and that of DC power measurement is ±0.01% of the reading value. The torque transducer has a resolution of 0.1 Nm with a maximum torque measurement capability of 100 Nm. The stator resistance is updated on–line using measurements from RTDs attached to the stator windings.

5.5.2 Tests at Varying Load–Speed Points Using $\gamma_{MEPA}$

The control methodology shown in Figure 5.14 was implemented in the Opal–RT real time controller to test the actual efficiency of the motor at $\gamma_{MEPA}$ and compared with the simulated efficiency. Figure 5.15 shows the comparison of experimental efficiency and
simulated efficiency for various $I_m$ and speeds. It is seen that the actual and the simulated efficiencies follow closely in a wide range of loading points. The slight difference in the efficiencies can be attributed to the magnet and stray losses and the off–line efficiency model accuracy.

![Figure 5.15. Comparison of experimental efficiency and simulated efficiency for various $I_m$ and speeds at room temperature.](image)

### 5.5.3 Sweep tests for Optimal Current Angle Benchmarking

In addition to the measurement of actual efficiency of the motor with $\gamma_{\text{MEPA}}$ condition and comparing with simulated values, a few preliminary sweep tests have also been conducted in the test motor by varying current angle in steps for different current levels and operating speeds. The objective of the sweep test is to detect the actual current angles corresponding to maximum torque and maximum efficiency for a given speed and load current. Thus, the validation using results from sweep tests also confirms the accuracy of the loss models and the simulation results. The current angle values obtained from the sweep test for maximum torque, $\gamma_{\text{MTPA}}$, and maximum efficiency, $\gamma_{\text{MEPA,actual}}$ will be used for validation of the developed method. The $\gamma_{\text{MEPA,actual}}$ will be compared against the optimal $\gamma$ value $\gamma_{\text{MEPA}}$ obtained off–line from the proposed method. The direct measurement of efficiency in the
sweep tests is performed using the power analyzer for DC and AC inputs at the inverter and motor terminals respectively and the torque transducer for the output mechanical power. The actual system efficiency and torque with respect to current angle showing the maximum efficiency and torque angles obtained for a sample point at $\omega_r=500$ rpm and $I_m=10$ A is given in Figures 5.16(a) and 5.16(b), respectively. Table 5.2 shows the current angles obtained for maximum system efficiency, $\gamma_{MEPA,\text{actual}}$, maximum, motor efficiency, $\gamma_{MEPA,m,\text{actual}}$, and maximum torque, $\gamma_{MTPA}$ at varying currents and speeds along with the comparison of simulated value, $\gamma_{MEPA}$.

![Efficiency vs current angle](image1)

![Output torque vs current angle](image2)

Figure 5.16. Results from sweep test at 500 rpm and $I_m=10$ A. (a) Efficiency vs current angle. (b) Output torque vs current angle.

<table>
<thead>
<tr>
<th>Speed (rpm)</th>
<th>700 rpm</th>
<th>575 rpm</th>
<th>375 rpm</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_m$ (A)</td>
<td>2</td>
<td>6</td>
<td>10</td>
</tr>
<tr>
<td>$\gamma_{MEPA,\text{actual}}$</td>
<td>18.5</td>
<td>29.9</td>
<td>36.5</td>
</tr>
<tr>
<td>$\gamma_{MEPA,m,\text{actual}}$</td>
<td>20</td>
<td>31</td>
<td>37</td>
</tr>
<tr>
<td>$\gamma_{MTPA}$</td>
<td>14.5</td>
<td>23.5</td>
<td>31</td>
</tr>
<tr>
<td>$\gamma_{MEPA}$</td>
<td>17.5</td>
<td>29</td>
<td>34.5</td>
</tr>
</tbody>
</table>

The comparison shows that the $\gamma_{MEPA}$ and $\gamma_{MEPA,\text{actual}}$ values are higher than $\gamma_{MTPA}$ values for all speeds and loads. This shows that there is an optimal current angle which minimizes the losses considering both copper and iron losses. The fundamental iron losses decrease with
increasing current angle and harmonic iron losses have individual trends for each load–speed point. Once the actual current angles were obtained from the sweep tests, system efficiencies obtained at $\gamma_{MEPA,\text{actual}}$, $\gamma_{MEPA}$ and $\gamma_{MTPA}$ for various operating conditions have been compared in subsequent sub–sections.

5.5.4  Effects of Developed Method Considering Saturation

The effect of saturation due to loading and the ability of the developed method to take into consideration is discussed in this sub–section. The current angle for varying load and speed was computed by keeping the flux linkages as a function of rated inductance for the test motor (in Chapter 2). The inductance was kept a constant for all loading conditions. This current angle, $\gamma_{MEPA,\text{unsat}}$ is defined as the unsaturated current angle and is compared with the $\gamma_{MEPA,\text{actual}}$ and $\gamma_{MEPA}$ values for varying $I_m$ at 575 rpm in Figure. 5.17. It can be seen that as the loading increases, the unsaturated current angle deviates from the actual value from sweep tests significantly whereas the current angle from the developed method follows the actual current angle closely. Thus, saturation is well represented.

![Figure 5.17. Comparison of $\gamma_{MEPA,\text{actual}}$, $\gamma_{MEPA}$ and $\gamma_{MEPA,\text{unsat}}$ for varying $I_m$ at 575 rpm.](image_url)
5.5.5 Effects of Developed Method Considering Temperature Variation

The search for optimal angle performed at operating temperatures of 25°C and 85°C yields $\gamma_{MEPA, 25^\circ C}$ and $\gamma_{MEPA, 85^\circ C}$. The values were incorporated in the experimental tests at varying speeds and loads and the actual efficiencies were measured. The search for current angle corresponding to maximum efficiency from the simulation at 25°C and 85°C is given in Figure 5.18. The values of system efficiency obtained in the experiments conducted at 85°C with uncompensated current angle $\gamma_{MEPA, uncomp}$ wherein there is no change in $\gamma_{MEPA}$ value from that at 25°C and the compensated current angle, $\gamma_{MEPA, 85^\circ C}$ is given in Figure 5.19 for various speeds and 14 A and 2 A. The uncompensated current angle is lower than the compensated current angle and it is seen from Figure 5.19 that the developed method is able to compensate for the change in temperature, and increase the overall system efficiency compared to uncompensated current angle.

Figure 5.18. Efficiencies at 25°C and 85°C with temperature compensation with respect to $\gamma$ and $I_m$.
(a) 25°C. (b) 85°C.
5.5.6 Comparison with Conventional MTPA Control

The MTPA angles corresponding to the peak currents were derived for the test motor. Subsequently, experiments were conducted at different speeds and loads for MTPA and MEPA angles. The comparison of the developed control method and conventional MTPA control for wide load torque points are given in Figures 5.20 and 5.21. Figure 5.20 shows the system efficiency comparisons for MTPA, $\eta_{MTPA}$ and developed method, $\eta_{MEPA}$ for various speeds and loads. Figure 5.21 shows the motor efficiency comparison $\eta_{MTPA,m}$ and $\eta_{MEPA,m}$ at various loads and speeds. It can be seen that the developed method yields increased system efficiency compared to the MTPA method, mainly at low loads and high speeds. The improvement is close to 1.5% for 2 A and 700 rpm as compared to 0.04% at 700 rpm and 14 A. The motor efficiency also improves close to 1.4% and the improvements are significant for all loading conditions and higher speeds. The method can give substantial efficiency improvements in high–speed motors where the ratio of iron loss to copper loss is much higher that the test IPMSM.

Figure 5.19. Efficiencies obtained experimentally at 85°C with and without temperature compensation as functions of speed at 14 A and 2 A.
Figure 5.20. Efficiency comparison the developed method with MTPA for varying loading and speeds.

Figure 5.21. Motor efficiency comparison of the developed method with MTPA for varying loads and speeds.

5.6 Discussions and Conclusions

In this chapter, an off–line search–based technique for deriving the optimal current angle corresponding to maximum system efficiency has been developed. The method considers saturation, cross–saturation as well as temperature variations in the equivalent circuit parameters that are used to calculate the efficiency in the simulations. An iron loss model that considers the flux linkages containing saturation information is developed based on
iron loss resistance calculation. The motor and inverter losses were used to calculate the efficiency at a system level and the current angle corresponding to maximum efficiency. The simulation results were validated using experimental investigations on the 4.25 kW IPMSM. Firstly, the simulation results were validated using experimental results by keeping the current angle the same value as obtained from the simulations. Secondly, sweep tests were conducted to obtain the actual current angle for maximum system efficiency. The various validations suggested the following conclusions:

(i) The system efficiency can be improved by considering optimized current angle at the system level. The proposed method improved system efficiency by more than 1.5% compared to MTPA control.

(ii) The maximum motor efficiency point is not necessarily the optimal system efficiency point, hence it is important to consider the system level losses instead of component level losses.

(iii) The consideration of saturation and temperature improves the system and motor efficiency significantly. The results with unsaturated inductances overestimates the current angle, leading to a false optimal angle. The results with uncompensated PM temperature leads to a lower current angle for which the experimental efficiency was lower than the compensated current angle. Hence, the developed method improves applicability in real scenario of PMSM motor and drive operation.
CHAPTER 6
ON–LINE METHOD USING DC POWER MEASUREMENT FOR ENERGY EFFICIENCY IMPROVEMENT IN PMSM MOTOR DRIVE SYSTEM

6.1 Introduction

This chapter proposes a novel method of efficiency improvement in a vector controlled PMSM through system level maximum efficiency point determination using current angle as a control variable. The models are derived such that the system losses can be calculated through available terminal measurements with minimum predetermined parameters and non–invasiveness, which is one of the advantages of the proposed method. Loss models for the inverter and the motor fundamental and harmonic losses, which are capable of being solved on–line using available terminal measurements in the system are initially developed. The loss parameters are all obtained from measurements; thus, this method does not require the detailed information of motor design parameters. The loss models and DC link power measurement are then used to seek the maximum efficiency angle for different operating conditions using a gradient descent optimization algorithm (GDA). The GDA algorithm is used to search for maximum efficiency operating point with current angle as the control variable by minimizing the ratio of total motor losses to the DC link input power at an operating condition. By considering the input power at DC link, this method extends the scope of loss reduction to a system level that is significant in EV applications. The developed method is robust against changes in inductances due to saturation and cross–saturation with loading conditions as well as temperature effects. The effectiveness of the developed method in improving the system efficiency is verified and compared with conventional maximum torque per ampere method. The proposed strategy has been
validated on a laboratory interior PMSM, and the efficiency has been calculated for
different speed and torque conditions. The experimental validations confirm the
effectiveness of the proposed solution in improving the motor drive system energy
efficiency.

6.2 Loss Models for On–line System Efficiency Improvement

This section describes the loss models used to represent controllable losses that can be
minimized in the PMSM drive system.

6.2.1 Fundamental Losses in PMSM

The fundamental losses in PMSM include copper, $P_{Cu,f}$ can be written as in (6.1) and core
losses per volume, $dP_{Fe,f}$ can be represented using (6.2) and (6.3) respectively. The $P_{Cu,f}$
depends on peak phase current magnitude and $P_{Fe,f}$ depends on the time variation of flux
density in stator teeth and yoke, thus, the magnitude of peak stator core flux density [86].

\[
P_{Cu,f} = 3I_{rms}^2 R_s = \frac{3}{2}(i_d^2 + i_q^2) R_s
\]

\[
dP_{Fe,f} = K_{eddy} f^2 B_m^2 + K_{Hys} f B_m^x
\]

\[
K_{eddy} = \frac{\pi x^2 d^2}{6\rho}; K_{Hys} = \chi V
\]

where, $I_{rms}$ is the fundamental rms phase current, $R_s$ is the stator resistance, $B_m$ is the peak
flux density in the core, $K_{Eddy}$ and $K_{Hys}$ are the eddy current and hysteresis loss coefficients,
\(x\) and $\chi$ are the constants dependent on core material property, $d$ is lamination thickness,
and $f$ is the fundamental frequency and $V$ is the volume of the core. The core losses can be
defined in terms of the constituting teeth and yoke losses in $d-$ and $q-$axis, $P_{Fe,t,dq}$ and
$P_{Fe,y,dq}$ and corresponding volumes, $V_t$ and $V_y$ as in (6.4) [40], [87].
\[ P_{Fe,f} = dP_{Fe,dq} V_t + dP_{Fe,ydq} V_y \]  \hspace{1cm} (6.4)

In (6.5), the \( d \)– and \( q \)–axis flux density in teeth, \( B_{t,dq} \) and yoke, \( B_{y,dq} \) corresponding to the \( d \)– and \( q \)–axis flux linkages \( \lambda_{dq} \) are written as a function of corresponding teeth and yoke areas, \( A_t \) and \( A_y \).

\[
B_{t,dq} = \frac{\lambda_{dq}}{A_t}; \quad B_{y,dq} = \frac{\lambda_{dq}}{A_y} \hspace{1cm} (6.5)
\]

The updated core loss equation with respect to the flux linkages in \( d \)– and \( q \)–axis by substituting for flux density in (6.3) and (6.4) using (6.5) can be written in (6.6) to calculate the fundamental core losses, \( P_{Fe,f} \).

\[
P_{Fe,f} = \frac{3}{2} \left[ k_{EC} f^2 (\lambda_d^2 + \lambda_q^2) + k_{Hy} f (\lambda_d^2 + \lambda_q^2) \right]
\]

\[
k_{EC} = \frac{V \pi^2 d^2}{6 \rho} \left( \frac{V_t}{A_t} + \frac{V_y}{A_y} \right); \quad k_{Hy} = \chi \left( \frac{V_t}{A_t} + \frac{V_y}{A_y} \right) \hspace{1cm} (6.6)
\]

The updated constants \( k_{Hy} \) and \( k_{EC} \) required to calculate core losses as a function of flux linkage have been derived from experimental tests. Various samples of slightly varying peak flux linkage values were created by conducting tests at very low \( i_q \) values just enough to compensate for the mechanical losses and zero \( i_d \) at various speeds, like the core loss determination method in [90]. However, unlike [90], in this study, only the constants are determined from the preliminary tests whereas the core losses are calculated on–line for varying operating conditions using (6.6) through measurements, providing a better core loss model. The constants \( k_{Hy} \) and \( k_{EC} \) change with operating speed and a fitting function has been used to derive the average values from tests conducted at four speeds of the test motor. Figure 6.1 shows the results of core loss with small variations in flux linkage in test motor for various speeds.
Figure 6.1. Iron losses calculated from no–load tests at varying speeds and flux linkages to determine average hysteresis and eddy current loss coefficients.

The average $k_{HY}$ and $k_{EC}$ derived from the test and used in this analysis are 0.35 and 0.02 respectively. The determination of the updated constants $k_{HY}$ and $k_{EC}$ through experimental methods provides more advantages and accuracy compared to the method using motor design parameters in [40] since such specific motor design parameters are not always available and measurements increase the practicality of the calculation by considering the effects of manufacturing.

6.2.2 Harmonic Copper and Core Losses

The PMSM drives are fed with pulse width modulated supply, in which the inverter generates time harmonics in the range of the selected carrier frequency level that are quite significant, especially at high speed condition [30]. Hence, it is imperative to consider the harmonic losses, $P_{Cu,h}$ and $P_{CL,h}$ caused by PWM harmonics to improve the efficiency through control techniques. The harmonic copper losses are caused by the current ripple from the inverter. The harmonic iron losses, as explained and derived in Chapter 2, are mostly eddy current losses influenced by harmonics in the flux density waveform [52], as induced circulating currents are functions of the rate of change of the flux density. For higher switching frequencies (>10 kHz), $P_{Cu,h}$ in PMSM can be neglected and the harmonic
iron losses $P_{CL,h}$ are predominantly considered to be eddy current losses. The increase in hysteresis losses due to these harmonics are insignificant [88], hence only the harmonic eddy current losses are considered. The harmonic eddy current losses are calculated from the radial and tangential flux densities as in (6.7) where $D$ is the density of the electrical steel plate, $h$ is the time harmonic order, and $B_{r,h}$ and $B_{t,h}$ are the $h^{th}$ harmonics of the radial and tangential components of the flux density [86].

$$P_{CL,h} = \int \sum_{h=1}^{\infty} K_{iron} D(hf)^2 \cdot \left\{B_{r,h}^2 + B_{t,h}^2\right\} dv$$

(6.7)

The $P_{CL,h}$ can be calculated from the Fourier series expansion of harmonic voltages, $v_2$ as in (6.8) where $\omega$ is the fundamental angular frequency, $a_h$ and $b_h$ are Fourier coefficients, $V_h$ and $\phi_h$ are harmonic voltage and phase angle.

$$v_2(t) = \sum_{h=1}^{\infty} v_h(t) = \sum_{h=1}^{\infty} (a_h \cosh \omega t + b_h \sinh \omega t)$$

$$= \sum_{h=1}^{\infty} \left(V_h \sin (h \omega t + \phi_h)\right)$$

(6.8)

Using (6.9) and (6.10), $P_{CL,h}$ is written as a function of ripple voltage rms, $\Delta V_{rms}$, and measurable quantities, DC link voltage, $V_{dc}$ and modulation index, $M_a$ [88].

$$P_{CL,h} = k_{h,eddy}^* \sum_{h>1} V_h^2$$

$$= k_{h,eddy}^* \left(V_{rms}^2 - V_{f,rms}^2\right) = k_{h,eddy}^* \Delta V_{rms}^2$$

(6.9)

$$\Delta V_{rms}^2 = \frac{V_{dc}^2}{3} \left(\frac{2}{\pi} M_a - \frac{1}{2} M_a^2\right)$$

(6.10)

The harmonic loss constant $k_{h,eddy}^*$ representing harmonic eddy current losses is 0.002 and can be calculated from finite element co–simulations with inverter and motor or measurements by segregating harmonic losses [88]. The magnet eddy current losses can be
kept to a minimum by keeping high switching frequency. Thus, the switching frequency in
the tests are kept at 12 kHz and the magnet eddy current losses are ignored.

6.2.3 Inverter Losses

The predominant losses in hard–switched three–phase insulated–gate bipolar transistor
(IGBT) based inverters are the conduction and switching losses in the IGBTs and
freewheeling diodes. The conduction losses of the six IGBT switches and freewheeling
diodes in a two–level inverter are calculated as the sum of their average losses [62], [89].
A trade–off is observed between the accuracy and simplicity of the models representing
inverter losses for control purposes. The switching losses and the conduction losses are
calculated as a function of diode and IGBT properties and \( I_s \) in (6.11) and (6.12) [89].

\[
P_{\text{inv,sw}} = \frac{6}{\pi} \left( E_{\text{on}} + E_{\text{off}} + E_{\text{rr}} \right) f_s V_{\text{dc}} I_s
\]

(6.11)

\[
P_{\text{inv,con}} = 6 \left\{ \left( \frac{1}{\pi} V_{\text{on}} + \frac{1}{4} R_{\text{on}} \right) I_s \right\}
\]

(6.12)

where \( E_{\text{on}} \) and \( E_{\text{off}} \) are the turn–on and turn–off energies of the IGBT, \( E_{\text{rr}} \) is the turnoff
energy of the power diode due to reverse recovery current, \( R_{\text{on}} \) and \( V_{\text{on}} \) are the average slope
resistance and average forward threshold voltage of the diode and IGBT. Table 6.1
provides the details of the IGBT parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>( V_{\text{on}} ) (V)</th>
<th>( R_{\text{on}} ) (mΩ)</th>
<th>( E_{\text{on},t}, E_{\text{off}}, E_{\text{rr}} ) (mJ)</th>
<th>( f_s ) (kHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>0.9, 2.2</td>
<td>2.88, 2.5</td>
<td>33, 56, 30.5</td>
<td>12</td>
</tr>
</tbody>
</table>
6.2.4 Total Losses

The total losses in the motor and inverter and related to the system efficiency can be written as given in (6.13) and (6.14) where $V_{dc}$ is the DC link voltage and $I_{dc}$ is the DC link current.

\[
P_{\text{loss, total}} = P_{\text{Cu,f}} + P_{\text{Fe,f}} + P_{\text{CL&A}} + P_{\text{inv.sw}} + P_{\text{inv.con}}
\]

\[
= \frac{3}{2} (i_d^2 + i_q^2) R_s + \frac{3}{2} (k_{EC} f^2 (\lambda_d^2 + \lambda_q^2) + k_{Hy} f (\lambda_d^2 + \lambda_q^2))
\]

\[
+ \frac{V_{dc}^2}{3} \left( \frac{2}{\pi} M_a - \frac{1}{2} M_a \right) + \frac{6}{\pi} (E_{on} + E_{off} + E_{rr}) f_s V_{dc} I_s
\]

\[
+ 6 \left\{ \left( \frac{1}{\pi} V_{on} + \frac{1}{4} R_{on} \right) I_s \right\}
\]

\[
\eta = \frac{P_{in} - P_{\text{loss, total}}}{P_{in}} ; \quad P_{in} = V_{\text{DC}} I_{\text{DC}} \tag{6.14}
\]

In the next section, the method of deriving maximum efficiency point during operation is discussed.

6.3 Proposed Gradient Descent Algorithm Based Maximum Efficiency Optimization Method

For a steady–state operating point described by load torque, speed, and temperature, this section describes the method of deriving the optimal current angle that corresponds to the maximum system efficiency. An overview of the search method is described first, followed by the implementation procedure.

6.3.1 Search for Maximum Efficiency Angle Using Gradient Descent Algorithm

The aim of the search method is given in (6.15) as minimization of $P_{\text{loss, total}}$ derived in (6.13). The problem can be translated to a maximization problem by maximizing efficiency in (6.14) with respect to $\gamma$, given that the constraints in (6.16) is met.

\[
\max_{\gamma} \frac{\partial \eta}{\partial \gamma} , \text{ s.t. } 0 \leq \gamma \leq \pi/2, \text{ and } (6.16) \tag{6.15}
\]
A computationally efficient gradient descent–based method is used to iteratively update $\gamma$ using the DC input power and loss values from measurements until it reaches the current angle corresponding to maximum efficiency point according to (6.17), where $k$ denotes time instant, and $\beta$ is the weight of gradient descent algorithm controlling the convergence speed.

Equation (6.17) can be written as a maximization problem where, using (6.18), the current angle at time $k+1$ is updated using measurements at time instant $k$. 

$$
\gamma_{k+1} = \gamma_k + \beta \frac{\partial (V_{DC}I_{DC} - P_{loss,total})}{\partial \gamma} 
$$

(6.17)

where $k$ denotes time instant and $\beta$ is the weight controlling the convergence speed. The derivative term in (6.17) can be discretized in the form of (6.19) at time $k$.

$$
\frac{\partial (\eta)}{\partial \gamma} = \frac{\eta_k - \eta_{k-1}}{\gamma_k - \gamma_{k-1}} 
$$

(6.19)

Substituting (6.19) into (6.18), $\gamma$ at time $k$ is given in (6.20)

$$
\gamma_{k+1} = \gamma_k + \beta \left( \frac{\eta_k - \eta_{k-1}}{\gamma_k - \gamma_{k-1}} \right) 
$$

(6.20)

The efficiency, $\eta$ at any instant is written in terms of input power and losses as given in (6.14). From (6.14) and (6.18), the gradient descent based $\gamma$ derivation is rewritten in (6.21).
\[ \gamma_{k+1} = \gamma_k + \beta \frac{\left( \frac{V_{DC}I_{DC} - P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_k - \left( \frac{V_{DC}I_{DC} - P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_{k-1} \} \]

\[ = \gamma_k + \beta \frac{\left( 1 - \frac{P_{loss,\text{total}}}{V_{DC}I_{DC}} \right) - \left( 1 - \frac{P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_{k-1} }{\gamma_k - \gamma_{k-1}} \]  

Substituting (6.21) into the discretized form given in (6.20),

\[ \gamma_{k+1} = \gamma_k + \beta \frac{\beta \frac{\left( P_{loss,\text{total}} \right)_{k+1}}{\left( V_{DC}I_{DC} \right)_{k+1}} - \left( P_{loss,\text{total}} \right)_k}{\gamma_k - \gamma_{k-1}} \left( \frac{P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_{k-1} - \left( \frac{P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_k \]  

Using (6.22), the developed method uses only a few loss parameters and terminal measurements for maximum efficiency angle detection. The method is robust against the changes in motor parameters such as inductance and PM flux linkage since only terminal measurements are used to calculate flux linkage and the parameters are not included in calculations. A pre–defined value, \( \epsilon \) is selected as in (6.23) wherein the satisfaction of (6.23) stops the \( \gamma \) search for maximum efficiency point at a given operating condition.

\[ |\gamma_k - \gamma_{k-1}| < \epsilon \]  

where \( \epsilon \) is a predefined small positive value. In this study, a simplified approach is used as a stop criterion in which the term \( \gamma_k - \gamma_{k-1} \) is replaced by its sign in (6.24) to avoid the dependence of the developed method on \( \epsilon \).

\[ \gamma_{k+1} = \gamma_k + \text{sign}(\gamma_k - \gamma_{k-1}) \beta \frac{\left( \frac{P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_{k+1} - \left( \frac{P_{loss,\text{total}}}{V_{DC}I_{DC}} \right)_k}{\gamma_k - \gamma_{k-1}} \]  

where \( \text{sign}(\gamma_k - \gamma_{k-1}) \) is a function as given in (6.25)

\[ \text{sign}(\gamma_k - \gamma_{k-1}) = \begin{cases} 1 & \text{if } \gamma_k - \gamma_{k-1} > 0 \\ 0 & \text{else} \\ -1 & \text{if } \gamma_k - \gamma_{k-1} < 0 \end{cases} \]
The implementation of search for maximum efficiency point using motor terminal measurements is explained further.

6.3.2 Implementation of Developed Method in Test Motor

The developed method is implemented in the 4.25 kW test IPMSM motor. According to (6.24), the method requires the motor speed, $d$– and $q$–axis currents and voltages, DC link current and voltage, and $M_a$ from terminal measurements to calculate the efficiency and derive the current angle corresponding to maximum efficiency angle. The average values of the measurements are used to update the output current angle using the gradient descent algorithm until optimal angle is reached. The copper loss calculation is a direct method using the $d$– and $q$–axis currents. The fundamental core losses are calculated using (6.6) by replacing the flux linkage calculation as in (6.26) by measuring the speed $\omega_e$, and $d$– and $q$–axis currents and voltages.

$$\begin{align*}
\lambda_d (I_d, I_q) &= \frac{V_q - R_s I_q}{\omega_e} \\
\lambda_q (I_d, I_q) &= -\frac{V_d + R_s I_d}{\omega_e}
\end{align*}$$

(6.26)

The voltages references from the proportional–integral (PI) current controller are used as the voltage measurements. The stator resistance, $R_{s,T}$ at temperature $T$ is updated using a resistance temperature detector (RTD) in the stator winding and updated using (6.27), where $\alpha$ is the temperature coefficient of copper, $R_{s,T_0}$ is the resistance at initial temperature, $T_0$.

$$R_{s,T} = R_{s,T_0} \cdot (1 + \alpha(T - T_0))$$

(6.27)

The $d$– and $q$–axis voltages from the outputs of PI controller are affected by non–linearity in the PWM voltage source inverter and there is a difference between the reference and
actual voltages [82] that will in turn affect the core loss determination in (6.6), hence (6.24) also. Therefore, average distortion compensation voltages $V_{dq}^{err}$ are used for compensating the non-linearity in $d$– and $q$–axis voltages according to (6.28) [82].

$$
V_d^* = V_d + V_{d}^{err}; \quad V_{d}^{err} = D_d V_{dead}
$$

$$
V_q^* = V_q + V_{q}^{err}; \quad V_{q}^{err} = D_q V_{dead}
$$

(6.28)

where $D_d$ and $D_q$ are periodical functions of rotor position, $\theta$ and $\gamma$ given in (6.29) and are constants for constant $\gamma$.

$$
D_d = 2 \sin \left[ \theta - \text{int} \left\{ \frac{3(\theta + \gamma + \frac{\pi}{6})}{\pi} \times \frac{\pi}{3} \right\} \right]
$$

$$
D_q = 2 \cos \left[ \theta - \text{int} \left\{ \frac{3(\theta + \gamma + \frac{\pi}{6})}{\pi} \times \frac{\pi}{3} \right\} \right]
$$

(6.29)

The distorted voltage term, $V_{dead}$ is a constant of 0.5 V for the test inverter.

The modulation index, $M_a$ in (6.30) is used to calculate the harmonic iron losses. The inverter losses are calculated from its loss parameters, DC link voltage and $I_s$ as in (6.11) and (6.12).

$$
M_a = \frac{2V_{lm}}{V_{DC}} = \frac{2\left(\sqrt{V_d^2 + V_q^2}\right)}{V_{DC}}
$$

(6.30)

Algorithm 1 summarizes the developed current angle detection method. The flow chart of implementation and control diagram are given in Figure 6.2. The control diagram is given in Figure 6.3.

Algorithm 1: Gradient descent algorithm based optimal efficiency control using $\gamma$

1. Initiate current angle, $\gamma_{init}$ and motor loss parameters and set $k=1$
2. Read $V_{dc}$, $I_{dc}$, $V_{dq}$, $I_{dq}$, $M_a$, $\omega_r$, and $R_s$ with respect to temperature from terminal measurements

129
3. Calculate losses according to (6.13) and $\gamma_{k+1}$ using (6.24)

4. If $|\gamma_k - \gamma_{k-1}| < \varepsilon$

   Output maximum efficiency angle, $\gamma_{MEA} = \gamma_{k+1}$

   Else

   Set $k = k+1$ and Go to 2

---

Figure 6.2. Flowchart representing implementation of gradient descent optimization for maximizing efficiency in PMSM drives considering system losses.

The measurements used to calculate the controllable losses are all available terminal measurements except the added DC power measurement, which is simple, cost–efficient and non–invasive.
By using the direct controllability of fundamental and inverter losses and indirect controllability of harmonic losses, the algorithm minimizes system losses. An initial $\gamma_{\text{init}}$ is set and the gradient descent algorithm uses the real-time DC power and loss calculations from measurements to detect maximum efficiency angle. Once the optimal angle is detected, the algorithm remains inactive unless the load/speed changes. It is to be noted that during transients, the loss calculations from measurements and the corresponding $\gamma$ update can be affected. Hence, the proposed approach is beneficial mainly under steady state conditions.

### 6.4 Experimental Investigations and Validations of the Proposed Maximum Efficiency Control Method

The proposed maximum efficiency detection method has been tested and validated on the
laboratory 4.25 kW IPMSM motor and drive system. The experimental setup and procedure to obtain the maximum efficiency angle for varying speed and load conditions are elaborated in this section.

6.4.1 Implementation of Developed Method in Test Motor

The test motor is under torque/current control mode controlled using a field programmable gate array (FPGA) based real-time simulator and is driven by a dynamometer which is in speed control mode. The IGBT inverter is sine PWM modulated and has a constant switching frequency of 12 kHz. The DC link voltage is set a constant at 660 V. For the speed and position sensing, a high-resolution encoder with a resolution of 2,500 cycles-per-revolution has been used. The current sensor used for DC power measurement has nominal current of 12 A and accuracy of ±0.084 A at 25°C. The DC voltage sensor is a closed loop Hall Effect current transducer with an accuracy of ±0.8% at 25°C and can measure up to 1,000 V. The current sensors for AC current measurements have a nominal rms current of 25 A with a resolution of ±0.2% at 25°C. A power analyzer was used for current/voltage measurements at the DC link and the three-phase motor terminals for validation of the proposed method. A torque transducer was used at the shaft between the test IPM and dynamometer to measure output power and validate the developed method and efficiency calculations. The accuracy of AC power measurement of the power analyzer is ±0.04% of the reading value and that of DC power measurement is ±0.01% of the reading value. The torque transducer has a resolution of 0.1 Nm with a maximum torque measurement capability of 100 Nm. The stator resistance is updated on-line using measurements from RTDs attached to the stator windings.
6.4.2 Sweep Tests for Optimal Current Angle Benchmarking

A few preliminary sweep tests have been conducted in the test motor by varying current angle in steps for different current levels and operating speeds. The objective of the sweep test is to detect the actual current angles corresponding to maximum torque and maximum efficiency for a given speed and load current. Thus, the validation using results from sweep tests also confirms the accuracy of the loss models and calculations. The current angle values obtained from the sweep test for maximum torque, $\gamma_{\text{MTPA}}$, and maximum efficiency, $\gamma_{\text{MEA,actual}}$, will be used for validation of the developed method. The $\gamma_{\text{MEA,actual}}$ will be compared against the optimal $\gamma$ value $\gamma_{\text{MEA,calc}}$ obtained from the proposed method. The direct measurement of efficiency in the sweep tests is performed using the power analyzer for DC and AC inputs at the inverter and motor terminals respectively and the torque transducer for the output mechanical power. The maximum torque and efficiency angles obtained for an example point at $\omega_r=700$ rpm and $I_m=10$ A is given in Figures 6.4(a) and 6.4(b), respectively. The variations in DC value of output torque and the efficiency with respect to $\gamma$ are depicted. The DC voltage and current that are used for the efficiency calculations are shown in Figures 6.4(c) and 6.4(d). In the tests, speed points higher than the rated speed were chosen to understand the extent of improvement at higher speed through reduction of core losses. The difference in $\gamma_{\text{MEA,actual}}$ from $\gamma_{\text{MTPA}}$ were significant at higher speeds, owing to the increase in core losses. The experimental results shown in the next sub–section will focus more on high–speed regions of the test IPMSM, for which the effects of the developed method are more profound.
6.4.3 Experimental Results on Efficiency Improvement in IPMSM

In this test, the proposed methodology is validated at various speeds and loads to detect the maximum efficiency angle using gradient descent algorithm. The accuracy of the developed method has been validated at varying speeds and loads. Figure 6.5 shows the iterations in tracking maximum efficiency angle when the test speed was 575 rpm and the current was changed from $I_m=6\ \text{A}$ to 10 A. The initial $\gamma$ was set as the known MTPA angle of 24.9 degree. The gradient descent algorithm was turned on at steady state and the current angle settled at a maximum efficiency angle of 26.25 degree. Figure 6.6 shows the performance of the developed method when speed changed from 575 rpm to 700 rpm at a constant $I_m$ of 10 A. The initial angle was set as the previous maximum efficiency angle of 26.25 degree. Figure 6.6(a) shows measured $d-$ and $q-$axis currents and DC link current
Figure 6.5. Current angle iteration to determine maximum efficiency angle at 575 rpm and current change from 6 A to 10 A; initial γ was $\gamma_{MTPA}$ of 24.9 degree. 

and voltage, Figure 6.6(b) shows the measured $d-$ and $q-$axis voltage during the change in current angle. The average values of $d-$ and $q-$axis currents and voltages that are used by the gradient descent algorithm are also shown. As seen from Figure 6.6(c), the current angle settles at the optimal maximum efficiency angle of 31.5 degree in six iterations.

The change in $\gamma$ from 26.25 degree to 31.5 degree is due to the change in fundamental and harmonic iron losses with speed. The actual MEA from sweep tests for 700 rpm and 6 A was found to be 29.9 degree. Thus, the proposed method is effective in representing the maximum efficiency control. A sample result for the summary of calculated maximum efficiency angle values, $\gamma_{MEA,\text{calc}}$ from the developed method for varying $I_m$ values at 700 rpm is given in Table 6.2 along with the actual MEA values, $\gamma_{MEA,\text{actual}}$ obtained from sweep tests for comparison.

<table>
<thead>
<tr>
<th>$I_m$ (A)</th>
<th>2</th>
<th>6</th>
<th>10</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\gamma_{MEA,\text{calc}}$</td>
<td>17.5</td>
<td>31.3</td>
<td>36.2</td>
</tr>
<tr>
<td>$\gamma_{MEA,\text{actual}}$</td>
<td>18.5</td>
<td>29.9</td>
<td>36.5</td>
</tr>
</tbody>
</table>
Figure 6.6. Results of developed method to determine MEA during speed change from 575 rpm to 700 rpm at 6 A. (a) $d$– and $q$–axis currents, DC current and DC voltage (secondary axis). (b) $d$– and $q$–axis voltages. (c) Current angle iteration.

The developed method has been compared with MTPA control method at various speeds and loads. The MTPA angle was derived from the sweep tests for varying $I_m$. To compare the obtained system efficiency using MTPA and developed control, the sample current angles at operating temperature of 25°C for varying stator currents and speeds of 575 rpm and 700 rpm are presented in Figure 6.7. It can be seen that the current angles increase with the load current in both MTPA and MEA but only maximum efficiency angle changes with speed. The difference in MTPA to MEA angles for the same operating condition is more obvious at 700 rpm and lower currents.
Figure 6.7. Comparison of current angles at MTPA and MEA conditions for varying speeds and currents at 25\(^\circ\)C operating temperature. (a) Current angles at 575 rpm and varying \(I_m\). (b) Current angles at 700 rpm and varying \(I_m\).

Figure 6.8. Comparison of efficiency at MTPA and maximum efficiency current angle conditions for varying speeds and \(I_m\) of 6 A and 10 A. (a) Efficiency at 575 rpm and varying \(I_m\). (b) Efficiency at 700 rpm and varying \(I_m\).

Figure 6.8 shows the efficiency comparisons using MTPA and MEA control for 575 rpm and 700 rpm and \(I_m\) of 6 A and 10 A. Figure 6.9 shows further comparisons of system efficiency at speeds from 175 rpm to 575 rpm and varying loads. Figure 6.9(a) and 6.9(b) show the system efficiency maps corresponding to maximum efficiency current angle and MTPA angle respectively. The results suggest that at higher speeds, the proposed method is beneficial in improving the system level efficiency. At low speeds and high loads, the developed method had minimal effect on the system losses. In most of the torque speed plane, the proposed method provides improved system efficiency as compared to MTPA.
method. The proposed method is expected to improve the system efficiency more significantly in high-speed IPMSMs.

Figure 6.9. Comparison of system efficiency at MTPA and maximum efficiency current angle conditions for varying speeds and load currents. (a) System efficiency at maximum efficiency angle. (b) System efficiency at MTPA angle.

Figure 6.10 shows the comparisons of efficiencies of the system and motor by choosing MTPA and MEA angles for 575 rpm and varying $I_m$. The copper losses do not vary with $\gamma$, thus remain the same in both cases. However, the fundamental core losses decrease with increase in $\gamma$. There was a slight increase in harmonic losses, which can be due to a decrease in $M_a$ with increasing $\gamma$ [9]. The PWM phase voltage waveforms measured at operating condition corresponding to maximum efficiency angle of $\gamma=36.5^\circ$ at 700 rpm and $I_m=10$ A is given in Figure 6.11(a) and Figure 6.11(b) depicts the corresponding harmonic spectrum. The total rms ripple voltage caused by carrier harmonics of first and second order sidebands for switching frequency of 12 kHz are the predominant cause of the harmonic iron losses. The total motor losses decreased when maximum efficiency angle was chosen compared to MTPA angle. There was a slight decrease in the inverter losses with increase in $\gamma$. This is mainly due to an increase in power factor with increase in $\gamma$. 
Figure 6.10. Comparison of motor and system efficiencies at MTPA and maximum efficiency current angle conditions at 575 rpm and varying $I_m$.

<table>
<thead>
<tr>
<th>$I_m$ (A)</th>
<th>2</th>
<th>6</th>
<th>10</th>
</tr>
</thead>
<tbody>
<tr>
<td>Methodology</td>
<td>MTPA</td>
<td>MEA</td>
<td>MTPA</td>
</tr>
<tr>
<td>Controllable Losses (W)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Copper</td>
<td>28.5</td>
<td>28.5</td>
<td>60</td>
</tr>
<tr>
<td>Core</td>
<td>57</td>
<td>51</td>
<td>69.8</td>
</tr>
<tr>
<td>Harmonic iron</td>
<td>14</td>
<td>14.9</td>
<td>16</td>
</tr>
<tr>
<td>Inverter</td>
<td>72.1</td>
<td>60.2</td>
<td>96.9</td>
</tr>
<tr>
<td>% improvement in system efficiency due to MEA</td>
<td>4.5</td>
<td>1.8</td>
<td>0.14</td>
</tr>
</tbody>
</table>
Figure 6.11. Three–phase PWM voltage measured at the motor terminals at 700 rpm for \( I_m = 10 \, \text{A} \) and \( \gamma = 36.5^\circ \) and corresponding spectrum. (a) Phase voltage measurement of PWM waveform. (b) Harmonic spectrum of PWM voltage.

To validate the developed method in considering operating temperature deviations, Figure 6.12 shows the efficiencies for varying \( I_m \) and 700 rpm at 25°C and 65°C by using the obtained corresponding optimal angles.

Fig. 6.12. System efficiency obtained using the developed method at 25°C and 65°C for varying \( I_m \) and speed of 700 rpm.

The developed method was validated under various loads and speeds. The results show that the efficiency improvement using the developed method was significant in higher speeds and loads below the rated current. An efficiency improvement of 4.5% was achieved at 700 rpm and 2 A and for half the rated current, improvements up to 2% were observed. The use of MEA can be of significant advantage in traction PMSM with high operating speeds. At
higher currents and lower speeds, the scope of efficiency improvement was minimal. The close match of $\gamma_{MEA,calc}$ and $\gamma_{MEA,actual}$ conclude the accuracy of the loss models in depicting the real motor losses. The slight difference in the angles is due to the increase in magnet eddy current losses due to space harmonics and PWM inverter harmonics, which is minor in the test motor.

6.5 Discussions on Adaptations of Developed Maximum Efficiency Control Method

Some of the adaptations of the developed method used for traction motor drive applications are briefed.

6.5.1 Dynamic Response and Improved Adaptations

Owing to the continuous search for optimal efficiency, the proposed algorithm can tend to fail in transient states when the torque or speed command are suddenly changed. One solution is to disable the algorithm in transient conditions and use only in steady–state conditions by tracking $I_m$ and speed.

6.5.2 Look–up Table Generation

The developed method can be used to replace traditional off–line loss minimization procedures owing to its multiple advantages and combined with less noisy look–up table implementations [45]. Unlike conventional loss minimization approaches, the method is straightforward and does not involve tedious numerical methods to solve the loss minimization problem.

6.6 Conclusions

An enhanced control approach towards system–level efficiency improvement in PMSM drives is presented in this study. The developed method incorporates a current vector selection towards maximizing drive system efficiency using DC power measurement and
motor and inverter loss models solved on–line. The inclusion of DC power in a gradient descent–based search towards maximum efficiency improves system efficiency by decreasing motor core losses and inverter losses. The algorithm does not need knowledge of motor parameters and the losses are solved on–line, hence the influence of saturation and temperature variations are taken into consideration automatically. Experimental validation with actual efficiency measurements in the sweep test and comparison with MTPA control method showed superior performance and that the system efficiency can be improved by using the maximum efficiency angle, mainly at higher speeds.
CHAPTER 7
CONCLUSIONS AND FUTURE WORK

7.1 Conclusions

In this dissertation, comprehensive loss models, parameter testing and control methodologies have been proposed for improvement in efficiency and performance of an interior permanent magnet synchronous machine (IPMSM) based motor and drive system. Chapter 1 discussed the state–of–the–art electric motor used in commercially available EVs and justified the need for methods of testing and control techniques towards improved efficiency from a system level considering real driving conditions of EVs.

Chapter 2 proposed the loss models for IPMSM taking the effects of PWM inverter into consideration. The aim of this chapter was to understand the interaction of the motor and drive and to study the behavior of harmonic losses in the motor, which are not considered in many control techniques. The contribution of this chapter was the improved analytical model to derive stator and rotor harmonic losses considering PWM effects in an IPMSM, which was missing in literature.

Chapter 3 proposed an investigation into the behavior of fundamental and harmonic losses in the IPMSM with respect to various control variables, such as DC link voltage, switching frequency and current angle. A field oriented control (FOC) based simulation model of the motor was developed including the loss models from Chapter 2 and inverter losses as functions of the control variables. Out of the three feasible options, this dissertation chose variable current angle control method for efficiency improvement. The loss behavior and values were validated experimentally on the test motor. The novel contribution of this
chapter was to study the loss and control variable behavior from a system level rather than a component level.

Chapter 4 proposed testing methods towards the understanding of parameter variation with respect to saturation, cross-saturation and temperature. The off-line and on-line methods suggested significant changes in inductances of the motor due to saturation and a decrease in PM flux linkage with increase in temperature. The torque calculated with updated parameters correlated with the measured value. The novel contribution of this chapter was improved methods of determining all the parameters of the motor simultaneously and the consideration of the effect of iron losses in output torque.

Based on the understanding of loss behavior in Chapter 3 and the parameter determination methods in Chapter 4, Chapter 5 proposed an off-line search based optimal current angle selection and control method to improve the system efficiency considering parameter variations and inverter and fundamental motor losses. The system efficiency was improved by more than 1.5% in comparison to conventional MTPA method. The novel contribution of this chapter was improved efficiency control algorithm considering system level losses and parameter variations due to temperature and saturation simultaneously.

The derivation of off-line optimal current angle requires preliminary tests to characterize the parameter maps prior to search based optimal angle derivation. In Chapter 6, an easier-to-implement, on-line method including motor harmonic losses was developed for optimal current angle search using gradient descent algorithm (GDA). The novel contribution of this chapter was improved online efficiency control algorithm considering system level losses. The algorithm does not need the knowledge of motor parameters and the losses are solved on-line, hence the influence of saturation and temperature variations are taken into
consideration automatically.

Both the control methods from Chapter 5 and 6 were validated experimentally on the test motor and improvements in efficiency of the system and motor were observed compared to conventional MTPA method. The developed control methods showed superior performance and improved system efficiency by using the maximum efficiency angle, mainly at higher speeds.

7.2 Future Work

The future work to be considered towards improvements in the developed methods and further efficiency increase include: (i) derivation of optimal control considering flux–weakening region of operation; (ii) considering the iron loss model based on short–circuit conditions for improved accuracy during deep flux weakening conditions; (iii) derivation of equivalent circuit based harmonic iron and magnet loss derivations for easier implementation in off–line optimal current derivations; (iv) temperature effect on inverter losses to provide a holistic approach considering cooling; and (v) improvements in system efficiency through changes in inverter topology by using wide–band gap switches (GaN/SiC) or multi–level inverter topologies along with the maximum efficiency control method.
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APPENDICES

Appendix A: Details of Side–Band Harmonic Current Derivation

The derivation of coefficients of side–band harmonic components based on Fourier series expansion by involving Bessel functions is given in this section.

A.1 Bessel Function used in Derivations

The $k$th order Bessel formula can be expressed as:

$$J_k(x) = \sum_{n=0}^{\infty} \frac{(-1)^n}{n! \Gamma(k+n+1)} \left( \frac{x}{2} \right)^{2n+k}$$

$$\Gamma(k+n+1) = (k+n)$$

(A.1)

The parameters $C_1, C_2, C_4, C_5,$ and $C_7$ in (2.3) and (2.4) are determined by

$$C_1 = \frac{4}{\pi} \left( J_2 \left( \frac{M \pi}{2} \right) J_0 \left( \frac{M \pi \xi}{2} \right) - J_1 \left( \frac{M \pi}{2} \right) J_1 \left( \frac{M \pi \xi}{2} \right) \right)$$

$$C_2 = -\frac{2}{\pi} \left( J_1 \left( \frac{M \pi}{2} \right) J_0 \left( M \pi \xi \right) + J_2 \left( \frac{M \pi}{2} \right) J_1 \left( M \pi \xi \right) - J_4 \left( \frac{M \pi}{2} \right) J_1 \left( \frac{M \pi \xi}{2} \right) \right)$$

$$C_4 = \frac{4}{\pi} \left( J_1 \left( \frac{M \pi}{2} \right) J_1 \left( \frac{M \pi \xi}{2} \right) \right)$$

$$C_5 = -\frac{2}{\pi} J_2 \left( \frac{M \pi}{2} \right) J_1 \left( M \pi \xi \right)$$

$$C_7 = -\frac{2}{\pi} J_4 \left( \frac{M \pi}{2} \right) J_1 \left( M \pi \xi \right)$$

(A.2)

Appendix B: Mechanical Loss Determination

The mechanical losses of the machine was experimentally measured using no–load tests conducted a magnet–less rotor. Figure B–1(a) shows test bench used for the loss determination and B–1(b) shows the mechanical loss results obtained by inserting the designed rotor without the magnets into the stator and running the machine at different
speeds by a prime mover. The speed of the prime mover and torque at the shaft was measured and the input mechanical power was calculated from torque and speed. Since there is no current in the windings and magnets in the rotor, core and copper losses can be avoided. Hence, only mechanical loss component prevails in the machine.

Figure B.1 Mechanical loss calculation from no–load tests on a magnet–less rotor. (a) Test bench setup. (b) Mechanical loss results for varying speeds.

Appendix C: Dynamic Equations for PMSM Modeling

The dynamic equations representing the PMSM operation $d$–$q$ frame are given in (C.1).

\[ v_d = R_s i_d + \frac{d\lambda_d}{dt} - \omega_e \lambda_q \]
\[ v_q = R_s i_q + \frac{d\lambda_q}{dt} + \omega_e \lambda_d \]  

(C.1)

where $v_d$ and $v_q$ are the $d$–and $q$–axis voltages, $i_d$ and $i_q$ are the $d$– and $q$–axis currents, $\lambda_d$ and $\lambda_q$ are the $d$– and $q$–axis flux linkages, $R_s$ is the stator resistance, and $\omega_e$ is the rotor
electrical angular speed. The currents in the $d-$ and $q-$axis based on current angle, $\gamma$ between the stator current and $q-$axis, and the electromagnetic torque, $T_e$ are as in (C.2)–(C.4).

\[
\begin{align*}
  i_d &= -i_m \sin \gamma \\
  i_q &= i_m \cos \gamma \\
  T_e &= \frac{3P}{4} (\lambda_d i_q - \lambda_q i_d)
\end{align*}
\]  

(C.2)

(C.3)

\[
T_e - T_l - B \omega_m = J \frac{d\omega_m}{dt}
\]  

(C.4)

where, $i_m$ is the peak phase current, $P$ is the number of poles, $\omega_m$ is the mechanical speed, $T_l$ is the load torque, $J$ is the moment of inertia and $B_m$ is the friction coefficient of the drive system. The current and voltage limits can be represented as given in (C.5):

\[
\begin{align*}
  V_{max} &\geq \left( V_d^2 + V_q^2 \right)^{1/2} \\
  V_{max} &\geq 2\pi f \left( \lambda_d^2 + \lambda_q^2 \right)^{1/2} \\
  I_{max} &\geq \sqrt{I_d^2 + I_q^2}
\end{align*}
\]  

(C.5)

where $V_{max}$ and $I_{max}$ are the maximum available peak phase voltage and current respectively and $f$ is electrical frequency.
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159
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