



CRANFIELD UNIVERSITY

HAYDER ABDULHASAN ABDULRAHEM

INVESTIGATION AND ANALYSIS OF TWO-LAYER SPOKE-TYPE
FERRITE INTERIOR PERMANENT MAGNET MACHINE

SCHOOL OF WATER, ENERGY AND ENVIRONMENT
Energy Engineering

PhD

Academic Year: 2015 - 2019

Supervisor: PROFESSOR PATRICK LUK

Co-Supervisor: Dr. John Economou

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This thesis is submitted in partial fulfilment of the requirements for
the degree of DOCTOR OF PHILOSOPHY

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ABSTRACT

A Novel two-layer spoke-type ferrite IPM design is presented based on a commercial induction motor as a low-cost high-performance alternative for new installation to meet efficiency requirements “greener” for various variable speed applications. The proposed design improves PM flux by flux-focusing techniques and maximises the reluctance torque with a two-layer structure. High torque density and efficiency are achieved with comparable performance to a rare-earth equivalent.

At first, two-Dimensional d-q frame analytical methods are developed based on the magnetic circuit models. Then, FEA models of the ferrite and induction motors are built and analysed. A comprehensive investigation is carried out focusing on the performances in terms of torque profiles, losses, efficiency and power factor. The influences of temperature variation and distribution on the performance of novel two-layer spoke-type interior permanent magnet machines under different operating conditions are analysed. The analysis is based on a developed thermal circuit model. For benchmarking, a standard industrial IM is used. “Comprehensive experimental tests are undertaken, and the results show the proposed ferrite IPM machine has distinctive advantages over the IM in terms of efficiency, power density and speed range. In addition, due to low cost ferrite materials, it has a comparable price advantage.

Finally, A prototype machine with the proposed design is manufactured, and both the two motors are tested under various operating conditions. The experimental results confirm the FEA simulations and validate superior performance of the proposed ferrite motor over its induction counterpart.

Keywords:

Ferrite, analytical model, induction machine, low cost, high efficiency.

ACKNOWLEDGEMENTS

I would like to express particular thanks to my supervisors, Professor Patrick Chikwong Luk. and Dr John Economou for their support encouragement and supervision over the whole study, which has been extremely helpful and essential. I would like to offer my special thanks to the Ministry of Higher Education and Scientific Research of Iraq for offering me the funding to run this research.

I would like to thank Dr Weizhong Fei and Dr Wenyi Liang for the valuable help and discussions during the research. I would also like to thank Sam Skears for her valuable assistance.

I would like to give my special thanks to my wife, Alaa Manhal, for her unconditional support. My heartfelt appreciation for my children Kerrar. Dalia and Mohammed for their love, encouragement, and support. My deepest thanks and love to my parents, my sister and brothers who encouraged me to do this research. I owe every step forward in my life to them.

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LIST OF ABBREVIATIONS

AC	Alternating Current
AFPM	Axial Flux Permanent Magnet
BLDC	Brushless Direct Current
CPSR	Constant Power Speed Range
DC	Direct Current
DEA	Differential Evolution Algorithms
DOE	Design of Experiments
EMF	Electromotive Force
EU	European Union
FEA	Finite Element Analysis
FSPM	Flux Switching Permanent Magnet
IM	Induction Motor
IPM	Interior Permanent Magnet
MMF	Magnetomotive Force
MTPA	Maximum Torque-per-Ampere
NdFeB	Neodymium-Iron-Boron
PM	Permanent Magnet
PMASynRM	Permanent Magnet Assisted Synchronous Reluctance Machine
PMSM	Permanent Magnet Synchronous Motor
RSM	Response Surface Method
p.u.	Per Unit
SDM	Steepest Decent Method
SMC	Soft Magnetic Composite
SPM	Surface-mounted Permanent Magnet
SynRM	Synchronous Reluctance Machine
UK	United Kingdom
US	United States
2D	Two Dimension
3D	Three Dimension

NOMENCLATURE

A_{copper}	Cross-section area of the copper wire	mm ²
b_o	Stator Slot Opening	mm
B_{adv}	Flux Distribution Due to the d-axis the vth MMF	tesla [T]
B_{g1}	Flux Density in the Airgap of Core Area ①	T
B_{g2}	Flux Density in the Airgap of Core Area ②	T
B_{gduv}	Amplitude of the vth Component Decomposed from Bgdu	T
B_{gdv}	Amplitude of the vth Airgap Flux Distribution by d-axis Current	T
B_{gquv}	Amplitude of the vth Component Decomposed from Bgqu	T
B_{gqv}	Amplitude of the vth Airgap Flux Distribution by q-axis Current	T
B_i	Intrinsic Magnetization	T
B_m	Flux Density of the PM	T
B_r	Nominal Residual Flux Density	T
Br_{20}	Residual Flux Density at the Temperature of 20°C	T
Br_T	Residual Flux Density at the Temperature of Tc	T
B_s	Flux Density in the PM Slot Opening	T
D_m	Location of the Secondary Pole	mm
D_{ro}	Rotor Outer Diameter	mm
D_{so}	Stator Outer Diameter	mm
Emf	Back EMF	v
f	Operation Frequency	Hz
F_{ad}	d-axis MMF	v
g	Actual Airgap Length	mm
g_c	Equivalent Airgap Length	mm
H_{g1}	Flux Intensity in the Airgap of Core Area ①	A/m

H_{g2}	Flux Intensity in the Airgap of Core Area ②	A/m
H_{gd}	Flux Intensity in the Airgap Due to the d-axis MMF	A/m
H_{m1}	Flux Intensity of the Main PM	A/m
H_{m1d}	Flux Intensity Inside the Main PM slots by d-axis MMF	A/m
H_{m2}	Flux Intensity of the Secondary PM	A/m
H_{m2d}	Flux Intensity Inside the Secondary PM slots by d-axis MMF	A/m
H_{s1}	Flux Intensity in the Main PM Slot Opening	A/m
H_{s2}	Flux Intensity in the Secondary PM Slot Opening	A/m
I	Current flowing in the copper coils	A (Amper)
i_a	Current of Phase a	A
i_b	Current of Phase b	A
i_c	Current of Phase c	A
i_d	d-axis Current	A
I_m	Amplitude of Armature Current	A
i_q	q-axis Current	A
I_{rated}	Rated Current	A
k_{br}	PM Temperature Coefficient	°C
k_c	Carter's Coefficient	-
k_c	Coefficient of eddy current loss	-
k_e	Coefficient of excessive loss	-
k_h	Coefficient of hysteresis loss	-
k_{rpac}	Increment factor	-
k_{rs1}	Rotor Main PM Slot Coefficient	-
k_{rs2}	Rotor Secondary PM Slot Coefficient	-
k_{wv}	Winding Factor for the uth Harmonic Component	-
L_1	Length of the Main PM Pole	mm
L_2	Length of the Secondary PM Pole	mm

L_{copper}	Length of the copper wire in each phase	mm
l_{ef}	Stack Length	mm
L_{md}	d-axis Magnetizing Inductance	henry (H)
L_{mq}	q-axis Magnetizing Inductance	H
L_{s1}	Depth of the Main PM Slot Opening	mm
L_{s2}	Depth of the Secondary PM Slot Opening	mm
M	PM Magnetization	weber (Wb)
m	Number of Phases	phase
N_a	Number of Turns in Series Per Phase	turns
N_{fa}	Winding Function of Phase a	-
N_{fb}	Winding Function of Phase b	-
N_{fc}	Winding Function of Phase c	-
N_t	Number of Stator Teeth	teeth
p	Number of Poles	phase
P_{cpl}	Copper loss	w
P_{cpl_ik}	Copper loss of the ith design at the kth generation	w
P_{crl}	Core loss	w
p_e	Excessive loss density	W/m ³
p_h	Hysteresis loss density	W/m ³
q	Number of Stator Slots	slots
R	Electric resistance of the copper coils	Ω
R_a	Phase AC Resistance of Armature Winding	Ω
R_{dc}	Phase DC Resistance of Armature Winding	Ω
R_{ro}	Rotor Outer Radius	mm
R_{si}	Stator Inner Radius	mm
T_c	Operational Temperature of PM	°C
T_{em}	Electromagnetic Torque	newton metre (N·m)
T_{pm}	PM Torque Component	N·m

T_{pmm}	Maximum PM Torque	N·m
T_r	Reluctance Torque Component	N·m
T_{rm}	Maximum Reluctance Torque	N·m
W_1	Width of the Main PM Pole	mm
W_2	Width of the Secondary PM Pole	mm
W_{s1}	Width of the Main PM Slot Opening	mm
W_{s2}	Width of the Secondary PM Slot Opening	mm
α_1	Mechanical Radians of Core Area ①	radians
α_2	Mechanical Radians of Core Area ②	radians
β_1	Mechanical Radians of Half of the Main PM Slot Opening	radians
β_2	Mechanical Radians of the Secondary PM Slot Opening	radians
δ	Skin Depth of Skin Effect	mm
ΔT	Change of Temperature	°C
$\Delta \rho$	Change of The Resistivity	Ωm
θ	Mechanical Position in the Stator Frame	
μ_0	Permeability of Vacuum	henry per mete
μ_m	Relative Permeability of PM	-
μ_r	Relative Permeability of Conductors	-
π	The Ratio of a Circle's Circumference to the Diameter	-
ρ_c	Conductor Resistivity	Ωm
τ_p	Pole Pitch	slots
τ_s	Stator Tooth Pitch	degree
ϕ_{bri}	Flux through the Saturation Bridge	Wb
ϕ_{g1}	Flux through Core Area ①	wb
ϕ_{g2}	Flux through Core Area ②	wb
ϕ_{m1}	Flux through the Main PM	wb
ϕ_{m2}	Flux through the Secondary PM	wb
ϕ_{ml1}	Flux through the Main PM Slot	wb
ϕ_{ml2}	Flux Through the Secondary PM Slot	wb

Ψ_f	Open-Circuit PM Flux Linkage	wb
ω	Electric Angular Velocity	rpm

1 INTRODUCTION

1.1 Research Description

Electrification brings considerable environmental benefits by shifting people's dependence on fossil fuel sources, and in many cases, electrification increases energy efficiency. Most of the electric industrial equipment and domestic appliances that have actuating parts require one or more electric motors, such as washing machines, fans and blowers, electric lawn mowers, air compressors, water pumps, woodworking machinery, and so on. There is a tremendous market for electric machines. Electric machines consume 35-40% of the electricity generated worldwide. Especially in the industrial sector, electric motor systems account for 70% of all industrial electric energy [1]. Electric machines with higher efficiency can not only cut down energy bills but also reduce greenhouse gas emissions. Due to the tremendous advantages of the ambient, researchers from both academia and industry are seeking for the viable solution of electrical drives with high power density and efficiency. Studies showed that potentially 25-30% of energy saving could be achieved in electric machines if higher efficiency motors were used in the past decade [2]. As mandatory efficiency standards are put into effect in US and European countries, there are more critical needs of improving motor efficiency. As is known, the induction machines (IMs) are commonly used in both industry and household devices due to the robustness in their structure, the maturity of manufacturing techniques and low cost, taking up 90% of the overall machine installation. However, IM's relatively low power factor and low efficiency still represent disadvantages. Due to the requirement of reactive power for excitation and considerable copper loss in the rotor cage bars/windings, it is difficult to further improve the power factor and efficiency for IMs [3].

With the increase of use the high energy permanent magnet (PM) materials, permanent magnet synchronous machines (PMSMs) have become the most attractive and promising machine type in many applications with advantages of high-power density, high efficiency and controllability. Since excitation current is not required using pre-magnetized PMs as field excitation, the copper loss and

accordingly overall loss can be reduced, leading to higher power factor and efficiency. Statistical data shows that a vast amount of energy can be saved by replacing old low efficient induction machines [4]. Compared to IMs, 3-4% improvement in efficiency could be achieved for the PM machines, and the cost difference due to the PM materials could be paid back through operating cost savings in less than 2.5 years at a 50-percent duty cycle [5]. What is more, with the development of high energy PM materials, torque density can be improved, and smaller and lighter electric motor is possible. Thus, PM machines with high energy rare-earth materials such as neodymium-iron-boron magnet (NdFeB) have been an appealing contender for various high-performance applications since its discovery, owing to the outstanding features in terms of efficiency and power density [6–10]. However, the supply chain and prices of rare earth materials are unstable and highly rely on China's export. The price history of neodymium material from 2009–2018 is depicted in Figure 1-1. Because of China's rare earth output restriction on 6th October 2008, the price of neodymium grows sharply from late 2009 to its peak in early 2011 and then decreased dramatically in the next two years. The rare-earth market is still volatile and highly relies on China's exportation. Therefore, there has been a surge of interests in seeking alternative solutions to rare-earth PM based electric machines. It is of strategic importance to develop alternative electric machines solutions with less or even no rare-earth PM materials for high-performance at low cost [11–13].

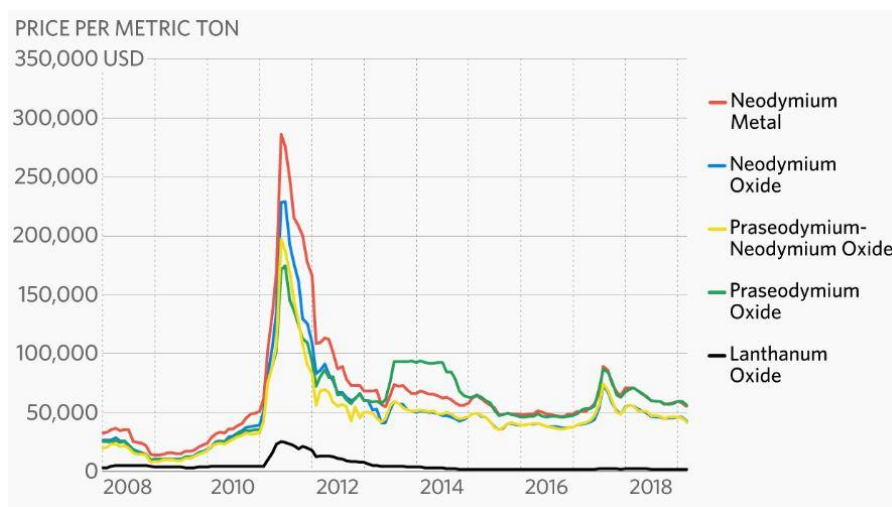


Figure 1-1 The price history of Neodymium material from 2008–2018[14].

Ferrite magnets, on the other hand, are abundant in supply and much cheaper than rare-earth. As one of the commonly used PM materials, ferrite PMs are considered as the most potential alternative for high-performance, low-cost applications. However, the residual flux density of ferrite magnets is only one-third of the NdFeB magnet, which raises a challenge for machines use ferrite PM to deliver required torque density. Thus, with well-designed geometric configurations ferrite PM machines can achieve comparable power density as rare-earth PM machines

1.2 Motivation

In the trend of global electrification, there is a large market of electric machines for electric industrial equipment and domestic applications. With concerns over the fuel depletion and environmental impact, energy efficiency has been given high priority in order to reduce greenhouse gas emissions. So, the electric machines with high efficiency are in demand. Thus, more and more strict efficiency mandatory standards are being introduced in China, the US and EU countries. Induction machines, as one of the most essential loads in both domestic and industrial applications are facing challenges to meet the efficiency requirements. PMSMs are showing remarkable performance for delivering high torque density and efficiency comparing to the IMs, which are the most promising way to improve the efficiency and fulfil the mandatory requirements. However, the high-performance PM machines are too reliant on expensive rare-earth materials at the moment, and thus the development of PMSM is hindered due to the concern over the volatility of the rare-earth material market. With the advantages of low price and abundance of supply, ferrite magnets are the most potential candidates for PMSMs to replace the IMs with similar material cost and higher efficiency. Despite the low residual flux density, ideas of flux focusing technology and reluctance torque harnessing provide inspiring thoughts on the ferrite machine design. Considering the huge number of variable speeds IMs need to be installed in industry and domestic applications, like air conditions and wash machines, there is a vast prospect in developing ferrite PM machines.

1.3 Research Aim and objectives

The research aims to develop novel two-layer ferrite machines with low cost, high torque density and efficiency, which offer improved performance and competitive cost to IMs. The objectives of this thesis can be summarised and listed as follows:

1. To carry out an e the literature review of the current developments and state of ferrite PM machines, evaluate the pros and cons of different ferrite machine structures and the discussion is concluded with the selection of the possible solution for the research project.
2. To develop an effective analytical model for the preliminary design of the proposed two-layer ferrite IPM machines. The new configuration base on one-layer configuration by adding the second layer to increase the reluctance torque and reduce harmonics in air gap flux
3. To present the electromagnetic design of two-layer ferrite IPM machines and determine the design solution performance for the project
4. To develop 2D FEA and 3D using ansys maxwel and jmag softwares model for the proposed machine and induction machine, to investigate the influence of rotor design on the machine performance such as power loss and thermal analysis.
5. To benchmark a commercial induction machine, and verify the performance of the proposed ferrite PM machine by comparing the performances of ferrite IPM and commercial IM theoretically and computationally.
6. To build and test the prototype machine, with comparisons of the commercial induction machine to validate the proposed ferrite design.

1.4 Major Research Contributions

The contribution of this research project can be summarised as follows:

- Optimum design of a novel two-layer structure for low energy ferrite PM machines is proposed and developed including high harmonics. With excellent flux focusing effect and improved rotor saliency, higher torque density and efficiency is achieved with a low-cost ferrite PM material.

- The systematic design process is presented and the influence of the key design parameters over the machine performance is revealed.
- Iron loss analysis of two-layer spoke type ferrite IPM machine is carried out considering with the effect of PWM harmonics
- A prototype machine with a two-layer configuration is developed and tested under various operating conditions; with the testing results of the commercial induction machine as a benchmark, the advantageous performance of the proposed design is validated.

1.5 Thesis Organisation

The thesis is organized as follows:

Chapter 2 reviews of publications relevant to the design and application of different rotational electric machines and the current state of the ferrite PM machines, and the most potential solution for the research project is noted.

Chapter 3 proposes the novel design of a ferrite two-layer interior permanent magnet (IPM). Firstly, the development of analytical models for two-layer spoke type IPM machines and the influence of the key design parameters are comprehensively investigated. Then, the optimal design is chosen for the prototype machine.

Chapter 4 analyses the loss characteristics. The heat sources are investigated and identified, and then the thermal effect of the losses and thermal performances are simulated under both rated and overloading conditions, so that the robustness of the design is verified thermally.

Chapter 5 Compares the performances of the two-layer ferrite machine and a commercial induction machine and confirms the advantages of the proposed ferrite design by testing the prototype machine under various operating conditions.

Chapter 6 discusses and analyses the experimental results of the prototype. With the commercial induction machine tested as a benchmark, the advantageous

performance of the proposed ferrite IPM machine over the induction machine is validated experimentally .

Chapter 7 summarizes the outcomes of the thesis and discusses possible future work.

1.6 Publications

1. Abdulrahem, Hayder; Luk, Patrick; Mohamed, Alsaady" Core Loss Analysis of a Ferrite Interior Permanent Magnet Motor" Cranfield Science for a Circular Economy Conference on, Cranfield University, 16 June 2017 (published).
2. Abdulrahem, Hayder A.; Luk, Patrick; Economou, John "Thermal Analysis and Temperature Distribution of Two-Layers Spoke-Type Ferrite Interior Permanent Magnet Machine" The 1st International Conference on Electromechanical Engineering & its Applications (ICEMEA-2019), Baghdad, Iraq on 15-16 December 2019. (accepted)
3. Abdulrahem, Hayder A.; Luk, Patrick "Inductance Calculation and Torque Separation of a Unique Multi-Layer Spoke-type Interior Permanent Magnet Motor using the Frozen Permeability Method" The 1st International Conference on Electromechanical Engineering & its Applications (ICEMEA-2019), Baghdad, Iraq on 15-16 December 2019. (accepted)
4. Abdulrahem, Hayder A.; Luk, Patrick; Economou John "Review of recent developments in ferrite permanent machines in EV and high-performance applications" 2019. (submitted to *eTransportation*)
5. Abdulrahem, Hayder A.; Luk, Patrick; Economou John "Performance Comparison of Two-Layer Spoke-Type Ferrite Interior Permanent Magnet Machine with Induction Machine". (submitted to *IEEE Transactions*)
6. Abdulrahem, Hayder A.; Luk, Patrick; "Demagnetization mitigation of Two-Layer Spoke Type Ferrite Interior Permanent Magnet Machines". (submitted to *IEEE Transactions*)

7. Abdulrahem, Hayder A.; Luk, Patrick; "Benchmarking a Novel Ferrite Interior Permanent Magnet (IPM) Decarbonisation of Transport through Electrification Summit, Cardiff University, 20th March 2019
8. Mohammed K. Al-Saadi, Patrick C. K. Luk, Hayder A. Abdulrahem "Impacts of Storage Battery on the Economic Operation of the Microgrids with considering the battery degradation cos" Cranfield Science for a Circular Economy Conference on, Cranfield University, 16 June 2017 (published).

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2 LITERATURE REVIEW ON FERRITE PM MACHINES

2.1 Background

Nowadays, with the massive use of fossils fuels, there would be predictable consequences for carbon emissions because of greenhouse gas and fuel depletion. These gas emissions lead to global warming and air pollution which bring about adverse influences on both the natural environment and human lives. With the growing concerns of the energy and environment issues, the mandatory efficiency standards in US and European countries, make more critical needs of improving motor efficiency, The induction machines (IMs) are the most widely used machine type in the industry due to the robust structure, sophisticated manufacturing techniques and low cost, taking up 90% of the overall industrial installation[1]. However, IMs have the disadvantages of low power factor and low efficiency. Due to the requirement of reactive power for excitation and considerable copper loss in the rotor cage bars/windings, it is difficult to improve the power factor and efficiency for IMs further. Statistical data shows that an enormous amount of energy can be saved by replacing old low efficient induction machines. This chapter is based on an exhaustive review of publications relevant to the design and application of different rotational electric machines. Firstly, four main electric machine categories, DC machine, induction machine, reluctance machine and permanent magnet machines are reviewed. For each category, different machine topologies and structures are summarized and compared. In the next section, an intensive review of recent developments in permanent ferrite machines in high performance applications has been conducted.

2.2 Main Categories of Electric machines

2.2.1 DC machines

DC motors have been utilised in various applications for over one hundred years now. They were originally chosen because of their inherent compatibility with DC power supplies and ease of controlling. DC motors may have series or separately-excited fields.

Series DC machine is the simplest type. The armature and field winding are connected in series and the excitation is proportional to the load current. So, the control system is quite simple. This type of series motor can develop a high starting torque and approximately constant mechanical power output

In the separately-excited motor, the field excitation is obtained using a power supply independent from that of the armature circuit. Since the flux density in the airgap can be adjusted by the independent field excitation, and excitation field and armature field are decoupled. These separately-excited DC machines have good controllability and simple controller structure. They may be controlled to produce high torque at low speed range, and yet fully utilise its rated power at high speeds. Also, it provides many other advantages such as better flux weakening ability, wide speed range, smooth field current and less pulsating torque. BOSE in [2] proposed a control system for separately-excited dc motor. A thyristor chopper was used to control the rotor armature during the low-speed range, while a high-frequency transistor was used to regulate the field in the high-speed range. The advantages of both series and shunt motors are achieved. It retains high starting torque and increases the operating speed range greatly with field weakening control. Additionally, optimal maximum torque can be achieved by increasing the current in armature and field coils simultaneously to improve the acceleration ability in the high-speed range.

To achieve more efficient DC machines, permanent magnets are used to eliminate the loss of excitation coils. P. Wolm et al. in [3] proposed an analysis of a PMDC motor model for application in feedback design for electric powered mobility vehicles. It proposed a method to increase the accuracy of PMDC motor modelling for a target EV application, with an enhanced model for DC motor by including an additional parameter of Coulomb friction torque.

ZOU Yuan et al. in [4] presented a new type of PMDC motor, with an augment-winding and controller to adjust the magnetic. The system is shown in Figure 2-1. The drive system was used in the electric transit bus. The mathematical model of the power-train is developed in the acceleration phase and high-power density

and efficiency were obtained. By controlling the occupancy ratio of IGBT, power density and efficiency is improved and strong acceleration capability is achieved.

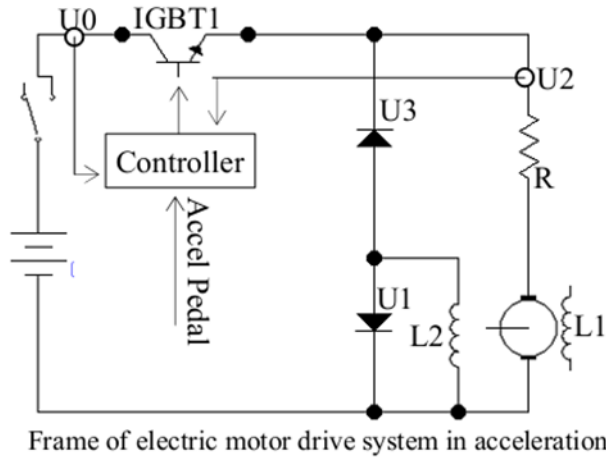


Figure 2-1 The control system of the new PMDC motor [4]

With the development of materials, new topologies and structure of DC machines are investigated by the researcher. B. Chalmers and E. Spooner in [5] presented a large superconducting DC drives with separate torque reaction windings, shown in Figure 2-2. A slotless armature DC motor with multiplex winding was proposed for higher power density comparing to the conventional slotted motor. The primary benefit of a slotless motor design is that the lack of teeth in the lamination eliminates cogging torque and results in a motor with very smooth running characteristics. Moreover, Slotless structure of the armature reduced the motor weight and eased commutation, while multiplex superconducting field winding obtained much higher torque output.

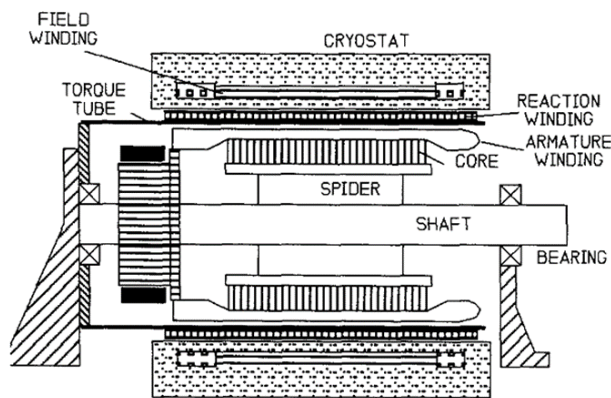


Figure 2-2 Superconducting Motor Structure [5]

P. Campbell, in [6] investigated the principles and structure of axial flux permanent-magnet DC machines. Depending upon the design adopted, these advantages can be high efficiency, high power/weight ratio, low inertia, high speed, and the physical shape of the machine. Distribution of magnetic field in the airgap is discussed and investigated with greater accuracy.

S. M. A. Sharkh et al. in [7] presented an axial field permanent magnet dc motor with powder iron armature. Axial flux permanent magnet (AFPM) DC motor with double air-gap was investigated in this paper. A prototype machine is shown in Figure 2-3. The two prototypes were of the same structure except for the material of armature teeth, one using powder iron and the other lamination steel. Results indicated that since the permeability and saturation of powder iron were lower, the one using lamination steel had a superior performance with higher back EMF, flux distribution, output torque and efficiency. In order to match the same performance, the one using powder iron needed to increase magnet thickness by about 50%. However, one disadvantage of steel laminated stator teeth was its labor-intensive construction. Nowadays DC machines are not as popular as AC machines. But it can still find some DC motors used in some applications very sensitive to the cost.

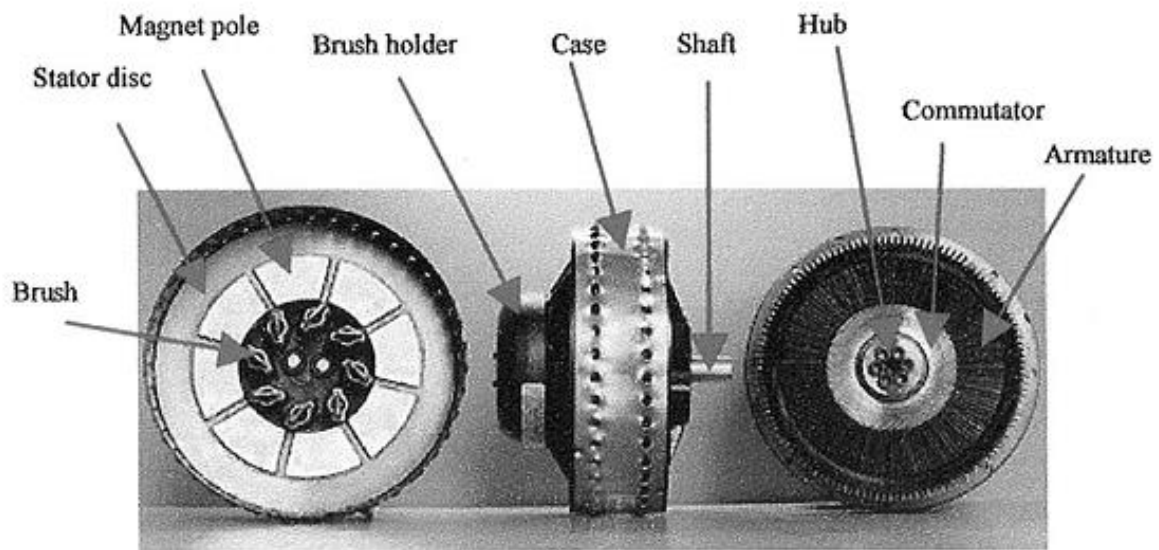


Figure 2-3A view of AFPM DC motor [7]

Tao Zhao et al. in [8] presented the simulation and analysis of DC motor driver system for the hybrid electric vehicle. Model of PWM converter with different switch devices for DC motor control system was developed based on MATLAB/Simulink with self-tuning modifying the algorithm for quicker system response. Pan, Singh et al. in [9] presented a power management method of fuel cell and battery fed DC motor drive for electric vehicle application. The fuel cell and battery are connected to the DC motor in parallel through DC-DC convertor, and the battery provides power during starting and transient conditions, while power supply relies on fuel cell during the dynamic condition.

Most presented papers of DC motor propulsion systems are concentrated on the control system and strategies, such as power management, energy regeneration and speed control. Clearly the development of power electronics has progressed apace since then and DC traction motors have become somewhat of a side-show in modern traction systems. With the development of the motor and drive system, the AC motor becomes more attractive solutions.

2.2.2 . Induction machine

Induction machine (IM) is the most widely used motor type in today's industry. IM has the characteristics of simple structure, robust, reliability and low maintenance

requirements, and can operate in harsh environments various applications. Considering that the exploiting of rare-earth magnet material is a great damage to the environment, and world-supply might be restricted in the future, IMs do have advantages over costs of material. Moreover, the mature manufacturing technology and infrastructure make IM a favourable motor even in high performance applications.

IM offers excellent peak torque capability, good flux-weakening characteristic and extent operating speed range. Since no magnet is used in IM, there is no risk of demagnetisation. These characteristics make the induction motor drive a suitable candidate for wide-speed range operation applications because the flux of the IM can be easily controlled and weakened. Properly designed IMs, such as spindle motor, are able to operate at a field weakened range of up to 5 times of its base speed [10].

However, induction machines are characterized by a lower efficiency when compared with the permanent magnet motors. So many researchers were trying some ways to maximise the efficiency in their works. Iron core and copper losses are frequency-dependent, and the influence of harmonics are analysed to find a way of reducing losses so that the overall efficiency is improved. [11] described a procedure to generate the flux maps of induction motor based on the equivalent circuit model. With the flux maps, the motor can be controlled to provide optimum flux at different torque and speed operating points. Hence high overall efficiency is achieved in the whole speed range. [12] presented a response surface method to optimise the slot shape to obtain maximum torque and efficiency in whole-speed range. [13] used two speed transmissions to couple with the driving motor so that it can operate at higher efficient points at whole speed range due to the different ratio of the transmissions. Many control strategies were also investigated to achieve maximum efficiency for IM [14]-[19].

In order to further improve the operation characteristics of IMs, leakage reactance needs to be restricted [20]-[21]. Open rotor slots structure, larger airgap width and unskewed open slots can reduce the leakage reactance. But certain harmonic effect should be taken into consideration.

With the development of material and motor technologies, new material and motor structure are applied in the design of induction motors. M. Morimoto used soft magnetic composite (SMC) to manufacture IM stator core in [22] and [23], and lower iron loss was obtained comparing to the laminated motor. And a divided core structure was presented in [24] for the assembly of the stator by tooth and yoke modules using SMC. Also, three-dimensional (3D) shape tooth module was developed to extend the pole surface area, as shown in Figure 2-4. And the efficiency and the maximum output power of the motor were increased.



Figure 2-4 3D shape tooth [24]

Weili Li and Junci Cao presented a new type of induction machine with compound cage rotor (IMCCR) in [20] and [25]. Rotor bars of the motor were re-designed base on squirrel cage rotor. The upper section of the bar adopted alloy material, which is made from materials conduct both magnetism and electricity such as pure iron and copper. The lower section was cast aluminium. The rotor bar structure is illustrated in Figure 2-5. At the same time, the operating current has to be small at normal operating. So the high starting torque and low starting current have to be obtained, but not take low efficiency and power factor as the price. The starting performance of normal induction machine and operating performance of induction machine with solid rotor can not satisfy the actual request. And this structure gave the motor a smaller starting current and higher starting torque. Axial Flux Induction Motor (AFIM) is also attracting considerable attention as a direct-drive wheel motor for EV application [26]. F. Caricchi et al.

presented a 20 kW AFIM with a single stator and double squirrel-cage rotors in [27].

The two rotors of the motor were coupled the two wheels of the EV, through independent shafts. This motor behaved as both an engine and a differential of a traditional vehicle. The superconductor is also used in IM. In [28] the authors proposed a squirrel-cage induction motor with superconductor as the secondary rotor conductor, and higher efficiency, output power, and starting torque could

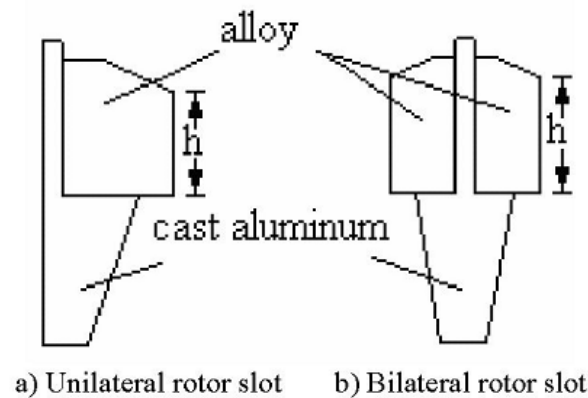


Figure 2-5 Rotor bar structure of IMCCR [32]

be achieved To extend the speed range, pole-amplitude modulation (PAM) technique was used in early times to obtain two-speed windings for induction motors [29]. For high performance drive systems, the concept of combining motor and generator into a single machine is favoured considering the power rating, efficiency and compact structure. So for this integrated-starter-generator (ISG) application, extended speed and a torque range of driving machine are desirable [30]. Recently pole-phase modulation (PPM) techniques are presented to provide induction machines with a suitable solution to extend its speed and torque capabilities as ISG [30]-[32]. Design methods of conventional winding machine, toroidally winding machine and dual-rotor toroidally wound machine were presented in [33]. Machines with the conventional winding and dual-rotor toroidal winding showed a high efficiency, while the conventional winding machine had a simple structure with reasonable performance, which could be attractive for utility. A uniform model for the PPM IM with convectional winding, considering pole-changing was built for design, analysis and control in [34].

Many researchers have attempted to address the trade-offs of fractional-slot concentrated windings in IMs for higher slot fill factor, larger power density, better efficiency and flux-weakening capability, with lower manufacturing cost [35]-[36], though this winding type is more popular in PM machines. In [35], the results showed that the performance of IM with traditional distributed lap windings was superior to that with fractional-slot concentrated windings, and fractional-slot concentrated winding IM tend to have higher torque ripple. As indicated in the analysis and experimental results, the 1/2 slot/pole/phase motor reduced coil number and simplified the end turn structure. Additionally, it had a promising potential in torque production. The weak point was that significant additional loss of rotor bar deriving from the 2nd and 4th mmf harmonics. J. Le Besnerais et al. [48] also revealed the appearance of some odd spatial order vibration waves due to fractional slotting, and some slotting lines and PWM lines combinations could bring about a lot of noises because of an elliptical mode natural frequency close to the switching frequency.

Many researchers have realized that there would be a significant increase in the efficiency of the induction motor by substituting the aluminium squirrel-cage rotor bars with copper [3743]. The electrical conductivity of copper is superior to aluminium. Thus, the rotor loss will be greatly reduced. According to the analysis and laboratory tests, the overall loss could be reduced by 14 to 23 percent in copper-rotor machines [3738]. Also, copper rotor motors often have a relatively higher breakdown torque when comparing to its aluminium counterpart [43]. But due to the technical problems in copper die casting, aluminium rotor is still widely used in small power IMs. However, for large power motors and some smaller special purpose motors, copper bar rotor structure is quite attractive for its high efficiency [38], [42].

In all, the induction machine has great advantages of simple construction, robust structure, low maintenance requirement and mature manufacturing techniques as well as excellent performances. But the biggest problem of the induction machine is its efficiency. Due to the rotor bar loss, it is not as competitive as its permanent magnet counterpart.

2.2.3 Reluctance Machine

2.2.3.1 Switch Reluctance Machine

Switched reluctance machine (SRM) is drawing people's attention as a potential low-cost candidate in the industry. The rotor structure of the SRM is simple and robust since no winding or permanent magnet exists in the rotor. Due to the inherent simplicity of machine structure, SRM has the advantages of low cost, high reliability and robustness, which make it very suitable and attractive in high-speed and harsh environment applications. Moreover, SRM provides competitive characteristics such as high power density, high instant power, high torque, fast dynamic response, and good fault tolerance [44-48]. However, SRM has significant disadvantages of large torque ripple, vibration and acoustic noise, which restrains it getting wider popularity in industry applications.

Extensive work has been carried out regarding the influence of the geometrical parameters on the performance of SRM.

The number of phases and the number of poles per phase have a nonlinear effect on SRM's performance. Normally, SRM has a higher stator pole number than the rotor. In [49] V. Petrus et al. analysed stator/rotor different pole number combinations with the same main dimensions such as outer diameter, stack length, and airgap length. The authors revealed that with the increase of phase number, the average electromagnetic torque was increased and the torque ripple was reduced. But the overall efficiency of higher phase number SRM was lower because the iron losses are seriously enlarged due to the largely changing magnetic field in the iron core. Also, higher phase number means more power electronic units and more complicated converters. Yuichi Takano et al. in [49] and [50] also concluded that torque was increased with the increase of pole number. In addition, Takano et al. proposed the use of Super Core “ Super Core is a registered trademark or trademark of JFE Steel Corporation in the United States ... JNEX-Core is the highest-grade non-oriented magnetic steel” with a thickness of 0.1mm laminations for their 18/12 structure machine, and with an increased taper angle of stator pole and super core material, competitive torque and efficiency with respect to the IPMSM was achieved for identical rating and dimensions.

Though conventional SRM structures have higher stator pole number than that of the rotor, a novel configuration with greater number of rotor poles than stator poles has been investigated recently [5158]. C. Desai et al. [5254] proposed a 6/10 three phase configuration and compared it with conventional 6/4 three phase structure. These two motor models were of the same main dimensions. The novel 6/10 SRM had many improvements in performance, such as higher average torque and lower torque ripple. Additionally, the stator and rotor pole width decreased and cross-sectional area of stator became smaller, and thus the area for armature winding was enlarged with a reduction of lamination mass. A higher continuous current was allowed and the power density could be increased due to the increase of the slot area. But the core losses of 6/10 could be higher than those of 6/4 because of the higher switching frequency. B. Bilgin et al. [5557] pointed out that 6/10 SRM would not bring any complexity in the control of the machine. And the machine presented with greater stator pole arc angle than rotor was capable of starting at any position. Moreover, 6/10 SRM had a better fault tolerant capability even when one entire phase was lost. S. Smaka et al. [58] investigated an 8/14 SRM and compared with a conventional 8/6 SRM. The static analysis showed that the 8/14 configuration had better torque density and lower copper losses.

To improve the output of SRM, B. C. Mecrow et al. in [59] proposed a segmented rotor switched reluctance machine (SSRM) structure as shown in Figure 2-6. the length of magnetic flux path is greatly reduced compared to conventional SRM, and thus greater torque density could be achieved. B. C. They later in [59] presented an improved SSRM with single tooth wound windings, as shown in Figure 2-7. Current in a single phase excited two adjacent slots, and up to 30% less copper was reduced because of shorter end-winding. Overall mean torque capability was increased more than 40% compared to conventional SRM. R. Vandana et al. in [60] proposed circular slots segmented switched reluctance machine (CSSSRM), as shown in Figure 2-8. The flux path was optimised, and active iron weight can be reduced. The electromagnetic torque was further improved compared to SSRM, which made it suitable for weight constrained applications. However, the mechanical robustness of the segmented rotor was

not compared to conventional SRM in these researches, as the robust of these configurations might be a weak point.

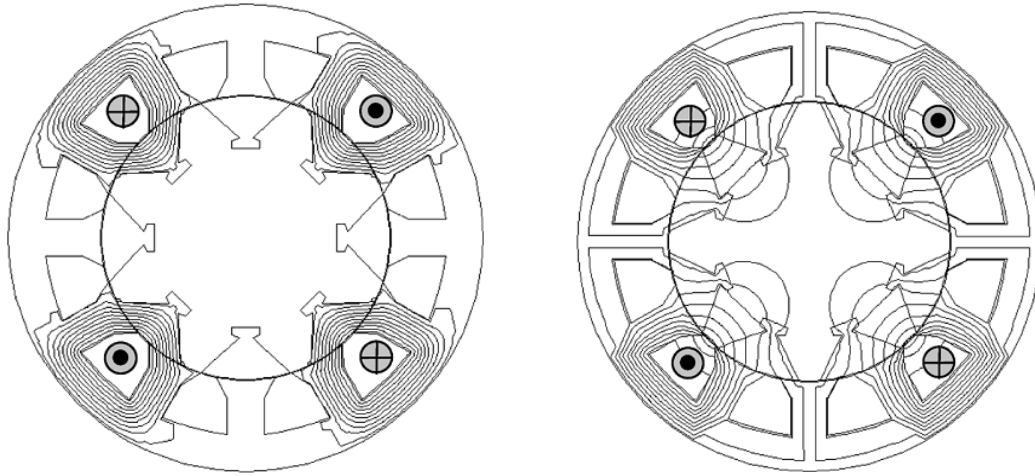


Figure 2-6 12/8 machine with one phase excited in the aligned and unaligned positions [59]

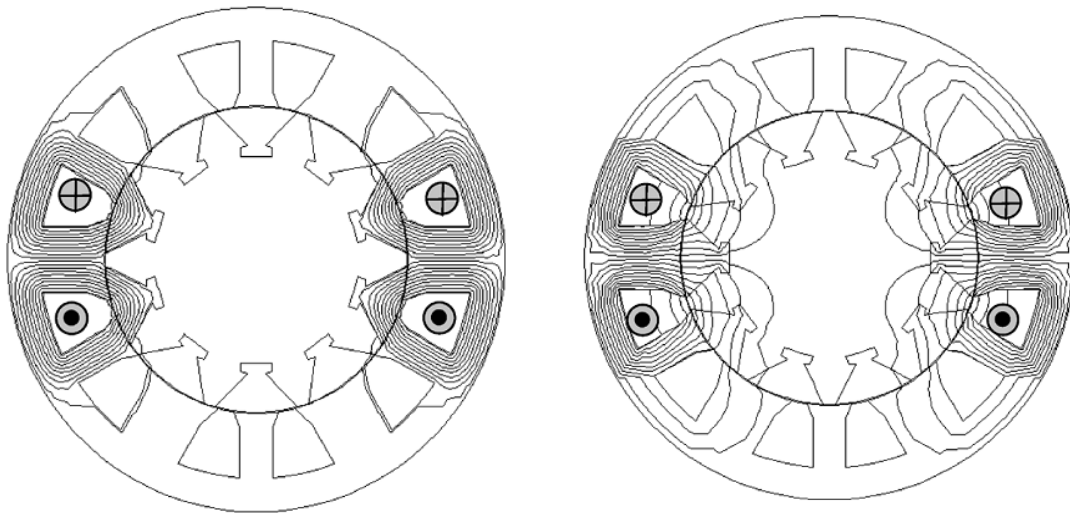


Figure 2-7 12/10 single tooth winding machine with in aligned and unaligned positions [47]

C. S. Kim et al. in [61] proposed designs with new π squire cores for SRMs. The flux path of π core SRM was shorter, but the stator structure was more complicated as shown in Figure 2-9. Experiments showed that these new SRM had an improved torque output, reduced torque ripple and higher efficiency. In all, SRM has many advantages of robustness, low cost, high torque, wide speed range and fast dynamic response, which makes it an attractive candidate. But the

SRM has major disadvantages of high torque ripple, noise and vibration. Though many design methods and topologies have been proposed to minimize the torque ripple and improve the performance of SRM, a lot of work still need to be done to make it more suitable for high performance applications.

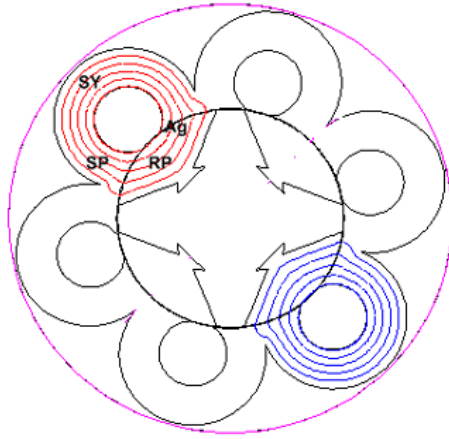


Figure 2-8 Structure of CSSSRM [60]

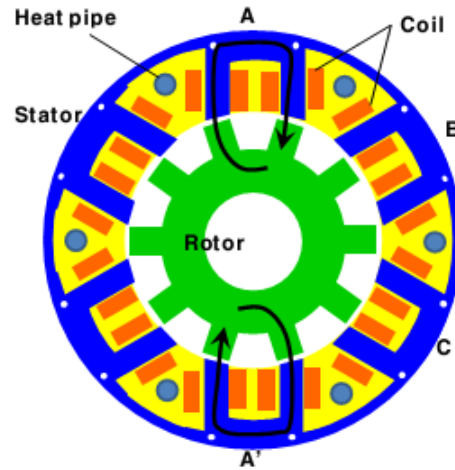


Figure 2-9 π Core SRM [61]

2.2.3.2 Synchronous Reluctance Machine

Synchronous reluctance machine (SynRM) has a simple structure, rugged characteristics and high efficiency. It does not have rotor winding or permanent magnet and operates at synchronous speed. SynRMs operates similar to synchronous machines, and the controllers of synchronous machines are compatible for SynRMs.

There are three main different types of SynRM with anisotropic rotor structures, namely simple salient pole (SP) rotor, axially laminated anisotropy (ALA) rotor and transversally laminated anisotropic (TLA) rotor, as shown in Figure 2-10. The salient pole (SP) rotor has the simplest and sturdiest construction, but the inductance differential is too small. Thus, the torque generated is too low under pure sine excitation [62]-[63]. Axially laminated anisotropic (ALA) rotor SynRM has layers of electromagnetic iron sheet in the rotor's axial direction to improve saliency ratio. It has a very high saliency ratio and a high difference between d- and q-axis inductances. However, due to the complexity of the laminated core

construction as well as problems in locking the core to the shaft, the manufacture and stiffness of such a machine is a big problem at present [6466].

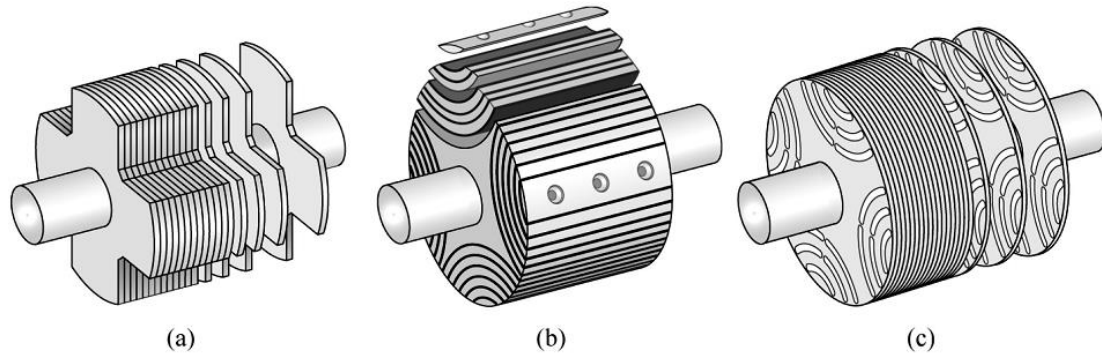


Figure 2-10 Three main different types of SynRM: (a) SP rotor, (b) (ALA) rotor, (c) (TLA) rotor [62]

I. Boldea in [64] presented a two pole ALA rotor SynRM. Analysis results suggested that a large number of slots per phase per pole should be used in the stator of an ALA rotor machine in order to obtain the expected high torque with saliency ratio of 16. Two-pole rotors offer the highest values of saliency ratio, but the core depth requirement restricts the space available for the shaft. B. J.Chalmers in [66] presented the design and performance of a four-pole SynRM, using thin rotor laminations, which achieved an unsaturated saliency ratio of 12.5 and a speed range of 2.25:1.

Laminations of transversally laminated anisotropic rotor cores are assembled in traditional ways and can be manufactured with existing equipment. There are multi-flux barriers in the rotor in order to obtain a high saliency ratio. Compared to the ALA rotor, TLA rotor is rugged and economical, easier to manufacture and more suitable for industry applications.

The rotor barrier number and shape can influence the performance significantly, and many research works have concentrated on the optimisations of stator and rotor geometries for SynRM.

In [6770] rotor optimum design algorithms were presented focusing on flux barrier width and numbers to reduce torque ripple and increase saliency ratio. M. Sanada in [28] presented asymmetric flux barrier arrangement to decrease torque ripple

without sacrificing the average torque. In [7172] stator slot and tooth were also taken into consideration for the improve the performance.

According to [7375], a proper quantity of permanent magnet was added into the SynRM rotor core to improve the operating performance of SynSM. In this case, the machine is similar to an interior permanent-magnet (IPM) motor, but the amount of permanent magnet used is much less.

SynRM does not have a speed range as wide as SRM, and flux barrier rotor to some extent limited the maximum because of the mechanical strength of the rotor. But SynRM has its advantages, e.g. the drive of SynRM is the same as a standard 3-phase machine with standard inverter power blocks, and it has a smooth torque down to zero speed [76]. According to J. Malan and M.J. Kamper, though SynRM is not as popular as other motor types, it may offer an intermediate capability between induction and permanent magnet machine drives.

2.2.4 Permanent Magnet Machine

Permanent magnet (PM) machines are becoming more and more popular with the advent of high energy permanent magnets. With the use of PMs, electric machines offer many definite advantages. PM machines are more efficient since field excitation windings and are absent. Thus, the rotor excitation loss is eliminated. High energy PM materials make it possible to use smaller magnetic thickness to achieve high flux density in the airgap, resulting in smaller machine diameter and mass [7782]. With unique features of their inherently high efficiency and high-power density, PM machines are attractive candidates for various applications.

According to the position of PMs in the rotor [8182], PM machines can be divided into surface-mounted, surface-inset, interior-radial, and interior-circumferential topologies, as shown in Figure 2-11. For a simpler classification, rotor PM machine can be classified into two categories, Surface PM (SPM) in Figure 2-11 (a)-(b) and Interior PM (IPM) machines Figure 2-11(c)-(d).

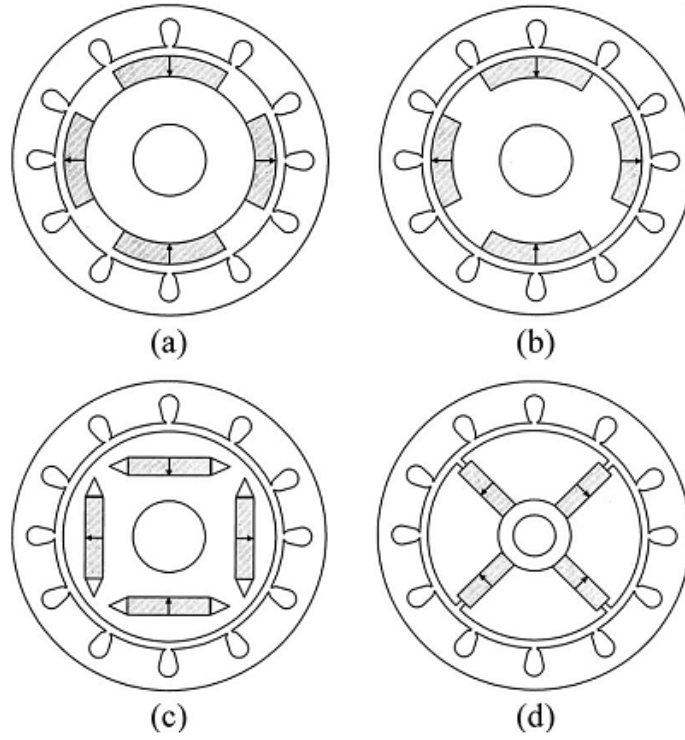


Figure 2-11. Rotor PM machine topologies: (a) Surface mounted; (b) Surface inset; (c) Interior radial; (d) Interior spoke [81].

2.2.4.1 SPM Machines

For the Surface-mounted PM (SPM) machine topology, the PMs are simply mounted on the rotor surface by using epoxy adhesives. Since the permeability of PMs is near to that of air, the effective airgap is the sum of the actual airgap length and the radial thickness of the PMs. Hence, the corresponding armature-reaction field is small, and the stator winding inductance is low. In addition, since the d-axis and q-axis stator winding inductances are nearly the same, its reluctance torque is almost zero. In order to extend the flux weakening ability and speed range, researchers have proposed many suitable design methods.

Concentrated, fractional-slot stator winding configurations were adopted in [83] and compared with both distributed windings. Machines with concentrated windings had a wider field-weakening region and lower winding losses. A.M. EL-Refaie et al. in [8485] presented a design approach for achieving optimal flux-weakening operation in SPM machines by properly designing the machine's

stator windings. The design increased the machine inductance, achieved optimal flux weakening, and delivered near-sinusoidal back-electromotiveforce (EMF) waveforms and low cogging torque. New machine topologies also have been proposed recently. G.B. Hamadou et al. in [86] presented a design of an integrated electromechanical structure for HEV application. The machine consisted of an inner SPM rotor, an outer SPM rotor, a stator sandwiched between the two rotors and a nonmagnetic material ring decoupling the outer and inner part for higher power density. J.A. Tapia et al. in [87] explored a magnetic structure of consequent-pole permanent magnet (CPPM) machine having inherent field weakening capability. The proposed motor structure had two components of the field flux. One was almost constant, produced by the rotor surfaced mounted PM, while the other is variable, produced by a field winding located in the centre of the stator, as shown in Figure 2-12. With this structure, air-gap flux can be controlled over a wide range without the risk of demagnetisation for PMs. Both SPM and field winding flux components for high torque density and wide speed range.

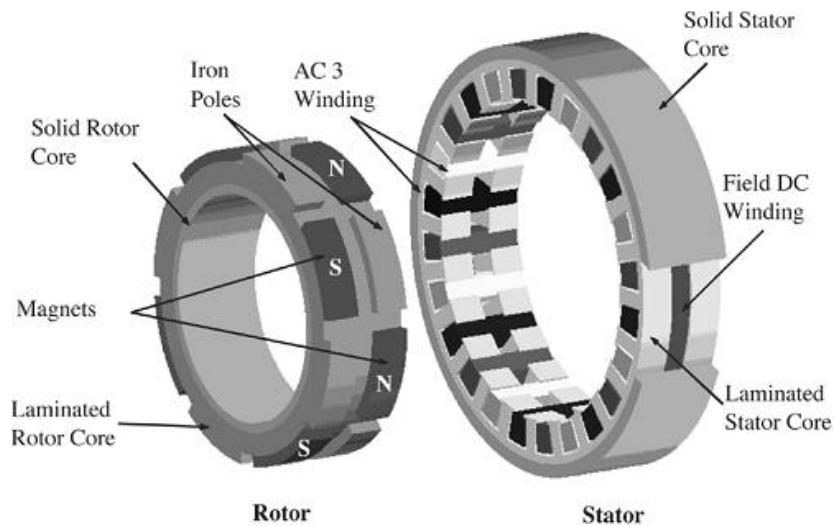


Figure 2-12 Structure of CPPM machine [87]

The existence of iron teeth in the rotor produces rotor anisotropy. Since the PMs are inset into the rotor surface, surface inset magnet machines have several advantages compared to SPM machines, such as higher rotor mechanical robustness, lower PM eddy current losses, higher q-axis inductance than d-axis,

and wider flux weakening region [8890]. However, compared to interior PM machines, the saliency ratio of surface inset PM is quite low, and thus, the contribution of the reluctance torque is limited and often not exploited [89].

SPMs are very difficult to meet the high requirements of the electric vehicle. Because, (a) the d-axis and q-axis stator winding inductances of such machines are the same, so they exhibit zero reluctance torque. (b) The magnets are exposed directly to the armature reaction field, and, hence, are susceptible to partially irreversible demagnetisation. (c) SPMs are also generally to have a relatively limited flux-weakening capability.

2.2.4.2 IPM Machines

An IPM machine with wide constant power speed ratio is considered to be ideal for many applications where energy saving is an important issue. Large saliency differential ratio of rotor exists which produces a reluctance torque. When added to the PM alignment torque, the torque output is increased. Thus, the power density of the machine can be improved. A wide speed range can also be achieved through field weakening due to the large d-axis inductance. Moreover, by burying PMs into the rotor lamination stack, a number of advantages can be created such as rotor mechanical robustness, reduced risk of demagnetisation from the stator winding currents and lower eddy currents on PMs. But there is a disadvantage of higher PM flux leakage for IPM rotors, which results in a reduction of the air-gap flux per pole and the PM torque.

Wide constant-power speed range (CPSR) is preferred for variable speed drive systems to increase efficiency and reduce controller capacity. S. Wu et al. in [91] compared four different rotor configurations, namely circumferential, radial, V-type and U-type configurations, using non-overlapping winding. In which each coil is wound around a single stator tooth. Compared with overlapping winding several advantages can be gained like. Shorter end turn winding, •easier manufacturability, •More fault tolerant and less probability of turn-to-turn short circuit fault.. Results showed that for non-overlapping winding, V-type and U-type rotor configurations exhibit better torque and power capabilities. The

investigations in [9296] covered that with proper design, complex PM configurations had better operation features such as higher flux density, higher torque output and most important, flux weakening capacity. In [95] L. R. Chong et al. claimed a 1kW IPM machine with a very wide 10:1 CPSR and an efficiency of 91.8% at maximum speed. Aimeng Wang et al. in [96] proposed $\overline{W}$ shape PM configuration with excellent flux-weakening performance and high efficiency over a wide speed range (above 95% over most of the operating speed range).

Multi-layer IPM machine reported in [97]-[100] with even wider CPSR and higher reluctance torque; however, the construction of such machines is more complex. The CPSR is extended by enhancing the saliency ratio to exploit the reluctance torque during flux weakening. However, due to q-axis saturation, a constant ratio of L_q to L_d cannot be maintained, leading to a reduction of CPSR.

A circumferentially magnetised (spoke-type) segmented rotor IPM machine is described in [101], but there are problems of high cogging torque and non-sinusoidal back electromotive force (EMF). X. Wang et al. in [102] proposed a spoke-type IPM with auxiliary poles to enhance flux concentration.

Fractional-slot concentrated winding has also been used for IPM machines to achieve higher magnetic saliency and reluctance torque as well as wide CPSR [103]-[105]. It was concluded in that concentrated winding may not be as effective in the IPM machine to achieve wide constant power operation as in the SPM machine.

Flux-switching permanent magnet (FSPM) machines have double salient structures, with both coils and PMs in the stator. This type of machines has simple and robust rotors, which is suitable for high speed applications. However, the stators are complicated and difficult for heat management. The structure consumes a large amount of PM material and has serious shortcomings of vibration and acoustic noise.

2.2.5 Comparison

DC motors are less reliable and require frequent maintenance, and the commutator limits the power range of DC machines. Therefore, DC motors are

mainly used in EVs with low power ratings and low operation requirements for simple and cheap construction. Induction motors are not as efficient as PM machines due to rotor bar losses. Moreover, the mature nature of this technology makes it unlikely to achieve a breakthrough for further improvements in efficiency and power density. Reluctance machines have better over-load capability and wider speed range in constant-power operation, which are potential candidates. But the low power factor and vibration problems make it difficult to suit high performance applications. Permanent magnet machines have the highest power density, efficiency and power factor, which is inherently the most attractive options. But magnets, especially rare-earth ones, are expensive, and the CPSR is relatively low.

2.3 Literature Review on Ferrite PM Machines

PMSMs are becoming the most attractive solutions for many high performance applications, including traction motors, wind generators and propulsion motor for aircrafts, because of their high power density, excellent controllability and high efficiency [106]. During the past decade, most of the high performance PMSMs have used rare earth magnets to provide high-torque density and efficiency. Nevertheless, the fluctuation in the prices and limited availability of the rare earth materials is the main concern for the researchers, which would definitely hinder further developments. After the difficult struggle in 2011 when the price of rare earth material hiked over 20 times, PM machine manufacturers started the search of an alternative method to substitute rare-earth for the PM machines.

Ferrite magnet is an important type of PM materials, which is drawing the attentions of researcher and designers for high performance PM motors. As a matter of fact, ferrite PMs were first commercialised in the 1960s, and are still the most widely used magnets today based on the consumption mass. However, the residual flux density and energy product of ferrite PMs are the much lower compared with its rare earth counterpart, as illustrated in Figure 2-13. The typical remanence of ferrite magnets is in the vicinity of 0.4T, which is only one third that of NdFeB. Thus, the machine characteristics would be greatly reduced simply by replacing rare-earth with ferrite PM or increasing the amount of ferrite PM in the

original design based on conventional PM machine configurations. Moreover, the weakness of ferrite magnets has been the low magnetic strength. Compared with rare-earth PMs, ferrite PM materials have much lower residual flux density and coercive force (200 kA/m), and are more vulnerable to the armature reactive field in terms of demagnetization. However, the incomparable advantages of unlimited availability of the material resources, low cost, stability to corrosion and high temperature, and very high electrical resistivity, ferrite magnets are represented as the most promising candidate for high performance machines with low-cost PM materials [107–109]. To satisfy the requirements of excellent efficiency, high power density and wide speed range, alternative configurations for ferrite machine designs are necessary to be investigated to maximise the torque capability of ferrite PM machines.

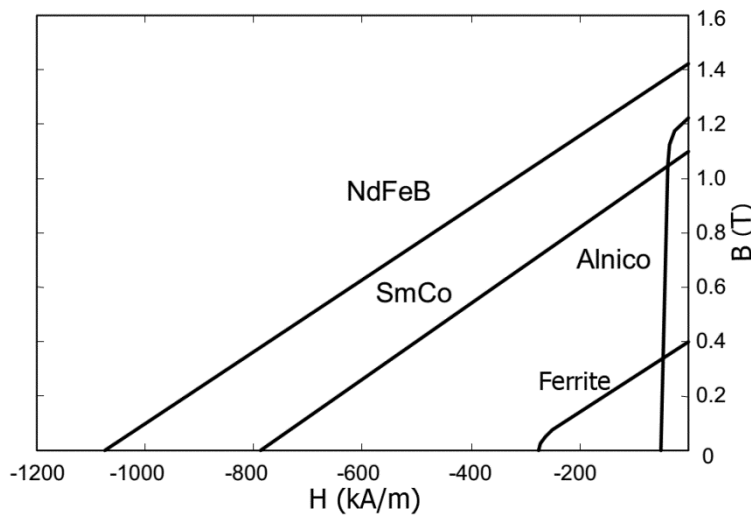


Figure 2-13 Comparison of BH curves among different PM materials.

Generally, there are two kind of electromagnetic torque components generated in PMSMs: the torque component resulting from the interactions between the armature windings field and the PM field, which is called PM torque, and the component created by the variation of inductance due to the rotor saliency and position, which is named as reluctance torque. It is obvious that ferrite PM machines would deliver much lower PM torque component as a result of much lower PM strength. To compensate the lower PM torque, special attention should be paid to the machine configurations to maximise reluctance torque component.

Ferrite PM materials have been used in the design of PM machines for over half a century, very soon after their commercialisation [110–115]. Researchers have investigated and compared various machine configurations with different PM materials in [116–119]. As a focus of this thesis, only developments of ferrite PM machines are reviewed for comparison and novel ideas of design technologies. Based on the existing literature, ferrite machine solutions can be divided into four categories:

- Conventional PM machines, which are of similar structure as the rare-earth machine with larger amount of ferrite PM;
- PM assisted synchronous reluctance machines (PMASynRMs), whose structures originate from synchronous reluctance machines (SynRMs) with PM in the flux barriers;
- Spoke-Type PM machines, using flux focusing techniques to achieve similar airgap flux density and torque density as rare-earth machines;
- Other types, which are not widely studied or not practical for mass production at the moment.

2.3.1 Conventional PM Machines with Ferrite Materials

Conventional structure PM machines are similar in terms of geometrical configurations as rare-earth ones, only by replacing rare earth PM with ferrite materials. Because of the low remaining flux density of ferrite PMs, a usually larger amount of PM was required to improve the airgap flux density. Even so, the power density ferrite machines are still very low comparing to its rare-earth counterparts. As a result, ferrite machines with conventional structures were mostly used for low-cost and low-power applications such as electrically assisted bicycles [120], pumps [121] and fans [122].

S. Pal in [123] investigated the different performances of SPM machines with exactly the same machine geometry but different PM materials of ferrite, SmCo and NdFeB respectively. The ferrite machine could only achieve 36% torque of SmCo and 31% that of NdFeB. P. Sekerak in [124] optimised ferrite SPM design

and obtained a final design with a slightly larger size but only 60% torque of rare-earth one. To increase the airgap flux density for SPM ferrite machines, the easiest way is to increase the size of the machine to accommodate larger amount of PMs. Nevertheless, the drawback of much larger volume-imposed limitations on the rotor design, and made it impossible for high performance machines with strict size and mass requirements.

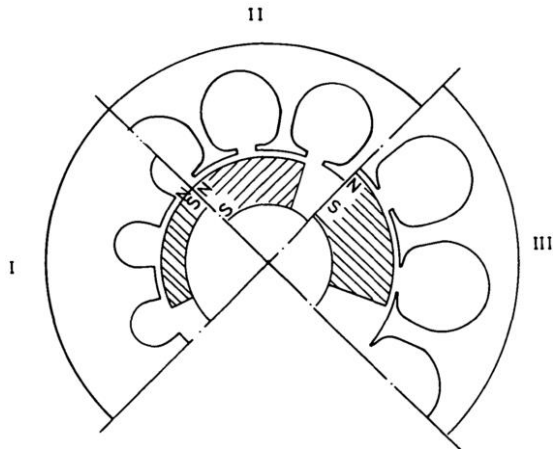


Figure 2-14 Design consideration with more PM material, thinner stator yoke and larger size [124].

Despite larger amount of ferrite being using in SPM machine, the airgap flux density still couldn't achieve that of rare-earth counterparts [125]. Thus, the stator yoke thickness can be reduced and larger diameters were necessary in order to place more armature conductors to increase the electric load and power output [126], as shown in Figure 2-14. In this case, the copper loss would be increased. Meanwhile, the machine exhibited over 10% lower efficiency under the same current density. K. Kim et al. in [127] compared rare-earth and ferrite SPM machines with the same size, and the ferrite model had over 5 times higher copper loss due to larger amount of copper used. In [128], copper loss of ferrite generator was almost 3 times as that of NdFeB with the same outer diameter, and thus ferrite machine suffered 7% lower efficiency. And with 30% larger diameter, the efficiency of ferrite motor was still 1% lower compared to NdFeB motor in [129]. R. Gupta et al. in [122] proposed a ferrite SPM machine with radial- and Halbach-magnetized magnets to increase the flux density and improve the efficiency, but the extra cost was brought in due to the Halbach-magnetization.

D. Woo et al. and H. Kim et al. in [130,131] proposed another way to increase the flux density in armature flux linkage by applying the structure of longer rotor than the stator. With this structure, the leakage flux loss would be offset at the end of the lamination cores, and thus the performance of the machine can be improved at the expense of more PM material and larger mass.

Outer-rotor structure, with a larger airgap diameter, was able to increase the torque and thus can improve the performance of the machine. I. Petrov and J. Pyrhonen in [132-133] presented outer-rotor low cost ferrite SPM machines of 4.7 kW and 50 kW for automotive applications. Although high efficiency of 93.4% was achieved, the torque density is relatively low. R. Qu and T.A proposed a dual-rotor SPM machine with toroidal windings, Lipo in [134] to increase torque density and reduce end-windings. But the flux density in the airgap was still too low to achieved acceptable power density, despite the extremely complicated structure.

With PMs buried inside the rotor, the magnetic airgap length can be reduced and higher flux density can be achieved with IPM machines. H. Kim et al. in [135] concentrated on the magnet pole shape optimisation to increase airgap flux density and reduce the cogging torque. Y. Im in [136] focused on the shape of rotor rib using response surface methodology to improve the performance of ferrite IPM. However, the improvements are still far from the requirements for higher torque density. B.N. Chaudhari et al. in [137,138] presented an IPM design using combined circumferentially and radially magnetised PMs to increase the flux density and rotor saliency, as shown in Figure 2-15. However, since the magnet poles were not sinusoidally distributed in the rotor, there would be a high content of harmonics components in the flux distribution.

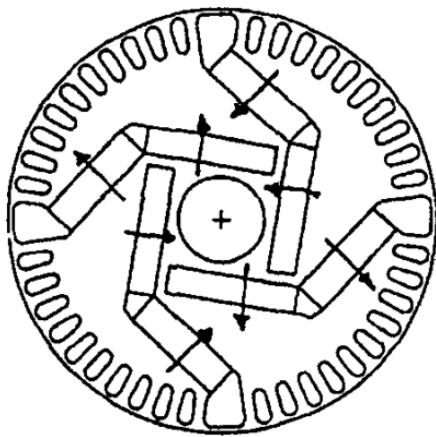


Figure 2-15 IPM with combined circumferentially and radially magnetized PMs [139].

Although the power density also could be improved by using more PMs, the amount of PMs has restrained the available space inside the rotor, and it is almost impossible to accomplish the design goal for high power range with sizing constraint [140]. To deliver similar torque with rare earth machine, the ferrite one had 47% higher volume to accommodate enough PM material [141].

To conclude, the low remaining flux density of ferrite materials and the large magnetic airgap length, SPM machines are not able to achieve the required torque density and efficiency. Though the airgap flux density can be improved by IPM structures, it is not likely for conventional IPMs to fulfill the requirements of high torque density and efficiency simply by increasing the amount of PM since the limitation of available space inside the rotor.

2.3.2 PM-Assisted Synchronous Reluctance Machines

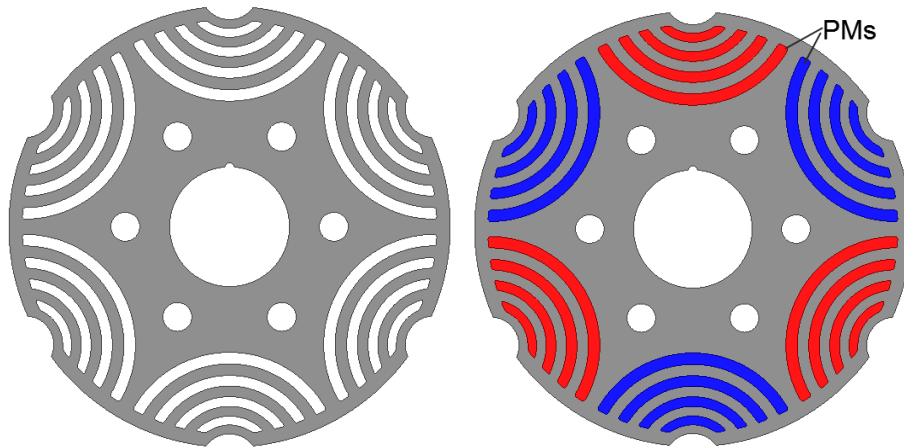


Figure 2-16 Configurations of SynRM and PMASynRM rotor.

PMASynRMs are derived directly from SynRMs, as illustrated in Figure 2-16. By adding the proper amount of PM materials into flux barriers in the SynRM rotor lamination, the power density, power factor and efficiency of the PMASynRM can be improved, which makes it a potential solution for ferrite IPM design [142–144]. And with more ferrite PM added to the original SynRM lamination, higher power factor, as well as efficiency, could be achieved [145,147]. E. Armando in [148] pointed out that with even a small amount of PM, not only the torque output but also the constant power speed range could be improved greatly compared with SynRM design for washing machines. However, it is not always the case that more PM results in higher torque. The main component of the torque for PMASynRMs is reluctance torque, and the rotor with too much PM would affect the thickness of barrier ribs and thus the saliency. D. Prieto et al. in [149] investigated the influence of the thickness of inserted PM poles and suggested more PM material could improve power factor, but also may cause a significant decrease in torque output. M. Barcaro and N. Bianchi in [150] optimised the length and thickness of PMs and indicated with similar rotor structure, reduced PM thickness can still attain similar torque capability, but slightly lower power factor.

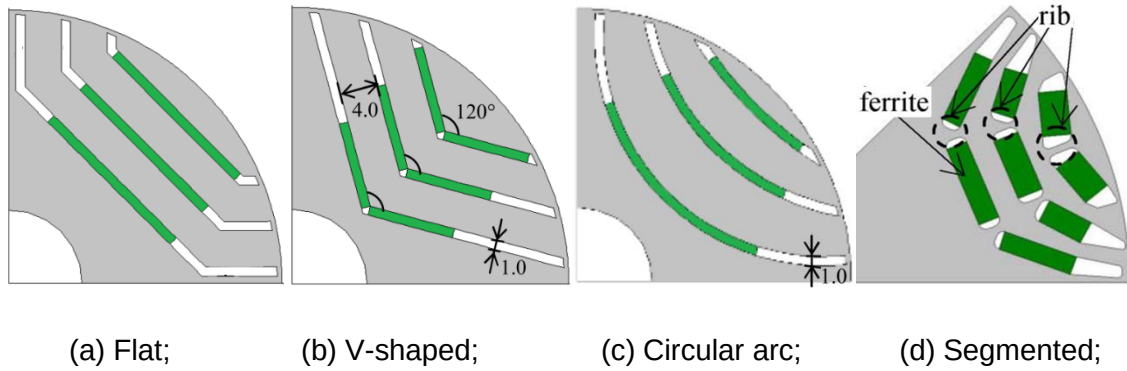


Figure 2-17 Rotor with different shapes flux barriers and PMs [151].

The shape of the PMs and flux barriers also showed great importance to the PMASynRM design. S. Musuroi et al. in [151] presented a two-layer V-shape ferrite PMASynRM since it has simple flux barriers and rectangular PM poles. K. Hayakawa et al. investigated the influence of flat, V-shaped and circular arc flux barriers, as shown in Figure 2-17. And results revealed that circular arc structure could substantially improve the maximum torque and power output with three flux barriers [152]. It should be noted that ferrite PM was brittle, especially when large thin arc shape was used. When taken massive production in consideration, long arc shape PMs were more expensive to produce and difficult to assemble. Thus, rectangular segmented PMs have proposed in place of arc shaped magnets for easy manufacturing and assembly [153,154]. Though there was about 4% decrease in the torque output, both the amount of ferrite and manufacturing cost was reduced.

PMASynRMs with three flux barriers were the most common configurations for the rotor design, with a compromise between complexity and saliency. However, designs with a higher number of flux barrier were also adopted in some research works. M. Paradkar et al. in [155] presented a design with 4 flux barriers and Y. Jeong et al. in [156] proposed a design with 5 flux barriers to gain a higher saliency. But the structures became too complicated. With the thickness of the PMs reduced, designs with more flux barriers made the PMs more fragile against mechanic stress and demagnetising armature field. Researchers also found that the thickness of saturation bridges, pitch and angle of flux barriers could affect the machine performance. Since the saturation bridges provide a flux path for PM

flux leakage and d-axis flux, both PM torque and reluctance torque could be reduced. H. Cai et al. proposed a design with all the bridges removed and yielded 8% increase of overall torque [157]. However, the lamination rib layers would be completely detached from one another, and the assembly of the rotor was not discussed, which would seriously affect the robustness of the rotor especially during high speed operation

To conclude, PMASynRM machines are descendent from SynRM and mainly rely on the reluctance torque component. Due to the inherent limitations of reluctance machines, PMASynRMs have inferior performances in terms of torque ripple, power factor and efficiency. A higher number of flux barriers can increase the rotor saliency and thus the power density, but it also complicates the rotor structure.

2.3.3 Spoke Type IPM Machines

The spoke type configuration, also known as flux squeeze configuration, has long PMs magnetised in the circumferential directions, as shown in Figure 2-18. This structure can provide a flux from the PMs to lamination poles to have an air-gap flux density nearly similar to that of the rare-earth machine. Thus, the spoke type structure also represents one of the most potential solutions for the ferrite PM machines with high performance.

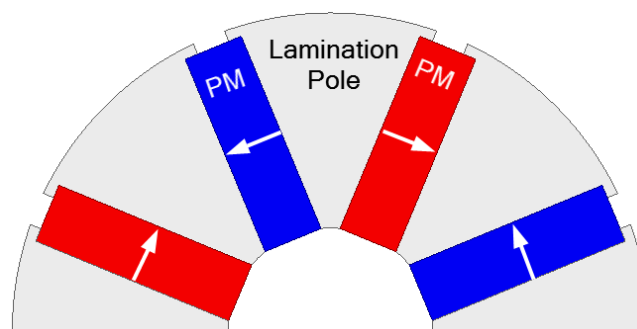


Figure 2-18 Rotor configuration of a typical spoke-type machine.

A. Isfanuti et al. in [158] compared surface NdFeB and spoke-type ferrite machines within the power range of 100-2000W, and results showed that same torque can be achieved with low cost, but higher mass and torque ripple. Eriksson and Bernhoff in [159] evaluated the spoke-type wind generators of the

same power range with ferrite and NdFeB PMs, and revealed that ferrite generators would be 50% heavier but with only 30% of PM cost. E. Spooner et al. proposed a modular design of ferrite spoke-type wind generators for easy assembly and low manufacturing costs [160]. For vertical wind turbines, low-cost ferrite rotor was preferred since there was little influence on the foundation structures. G. Dorrell in [161] presented a spoke-type design benchmarked with Prius rare-earth IPM, and the amount of ferrite PM was almost maximized in the rotor while only 2/3 torque was obtained even though larger machine diameter was adopted. Design, optimization and assembly of ferrite spoke type machines were presented in [162], and promising results were achieved for prototype machines. Although the torque density was much lower than rare-earth machines, it was still considered to be viable alternative solution in the sense of cost. To increase the torque density, F. Demmelmayr et al. in [163] proposed an outer-rotor spoke-type design. In order to achieve the same performance with rare-earth design, larger amount (5.7 times) of PM was used and thus outer diameter was enlarged by 26%. To compensate the low residual flux of ferrite and reduce the consumption of rare-earth, hybrid PM arrangements with both ferrite and rare-earth PMs were presented in [162]. Though the amount was reduced, rare-earth PM was still necessary for the design. Another issue should be paid special attention to be the cross- demagnetization between strong rare-earth and weak ferrite PMs.

To achieve similar airgap flux density as rare-earth machines, various designs are proposed to use more PM to enhance the PM exciting field. K. Kim et al. in [164] proposed a ferrite spoke-type design with large axial overhang and the air-gap flux density was increased by 33%. I.C. Chabu et al. obtained a 55% increase in the airgap flux density applying 68% longer rotor stacking than the stator [165]. W. Kakihara et al. in [166] also applied 6 mm rotor overhang and 6 mm PM overhang, the torque output was improved and PM demagnetisation rate was reduced. Large rotor overhang would certainly increase the amount of PM and rotor lamination, and the overhang PMs were not actually fully utilised since a considerable part of the flux did not go through the airgap directly but became flux leakage and wasted.

Compared with PMASynRM, spoke-type suffers higher risks of irreversible demagnetisation [167], and the investigation of irreversible demagnetisation is critical for the robustness of ferrite machines. Demagnetisation occurred in many of the papers reviewed above [109,168-170], but no solution was presented. Wider and thicker PM poles were used to enhance PM MMF [162], but this would affect the whole electromagnetic design of the machine, especially for those harnessing reluctance torque. B. Lee et al. in [171] proposed flux barriers in between the adjacent spoke type poles to reduce the demagnetising armature field, but PM flux would also be reduced. It was indicated in [172] that divided the main pole structure had a lower risk of demagnetisation than conventional one. Since the part of PMs near the airgap was facing a strong armature field, the PM edges closer to the stator suffered a higher risk of local demagnetisation. Thus, chamfered PM edges were applied in [173], and deeper burying of PMs were presented in [161,166,174]. Although the PM flux field was weakened, local demagnetisation on the PM edges could be greatly reduced. After all, demagnetisation was discussed during the design for certain cases, but no systematic research was published for ferrite PM machines yet.

Since the optimisation of the design is a very complicated work considering the flux distribution, torque generation, efficiency and so on, various multi-objective optimisation algorithms were used to accomplish the optimal design. K. Hwang et al. applied steepest decent method (SDM) to minimise the cogging torque and response surface method (RSM) to reduce harmonics in airgap flux distribution and cogging torque in [175] and [176] respectively. P. Zhang et al. in [177] used a combined design of experiments (DOE) and differential evolution algorithms (DEA) to optimise three key objectives of loss, cost and torque ripple. Although good results were achieved, the optimisations still required finite element analysis (FEA) to evaluate each design. Also, no clear relation between the performance and sizing parameters were derived and provided weak guidance on the design.

For more useful theoretical guidance during the preliminary design period, analytical methods were required for fast calculation. Approximate open-circuit flux density was obtained in [165,178,179] for conventional spoke type

configurations. PM flux distribution equations were derived for spoke type with auxiliary ring poles [180,181], wing-type [182], and wing-shaped [183]. However, the overall torque profiles, especially the reluctance torque component, could not be expressed only by the PM flux field, and thus the analytical was not complete. PM flux, together with dq-axis inductances, was derived for spoke type configuration in [184,185], showing the influence of different design parameters on the performances. But only the fundamental flux was considered, which could bring in relatively larger error.

To conclude, the ferrite spoke-type PM machines have the ability to achieve high torque density, efficiency and power factor with focusing flux structures. However, the reluctance torque is relatively low and further improvement needs to be explored. What is more, the rotor structures of spoke-type motors need to be designed very carefully due to the higher risks of irreversible demagnetizations.

2.3.4 Comparisons

Ferrite PMs were invented in the 1950s and has been used in conventional PM machines since then, but mainly in low-cost applications [186]. However, because of the low remaining flux density in the ferrite PMs, normal designs are very unlikely to realise the required efficiency and torque density for the high-performance requirements in various industrial applications.

Based on the extensive review, PMASynRM and spoke type IPMs are most attractive solutions with relatively mature technology to achieve the aim of rare-earth-free high efficient machine designs[187]. PMASynRM configurations descend from SynRMs with PM inserted inside the flux barriers to increase the torque production and power factor. Since the PMASynRM originates from reluctance machines, it has a high saliency ratio and depends largely on reluctance torque. Also, it can operate at wider speed range with better flux weakening ability. On the other hand, spoke-type IPMs can potentially achieve approximately similar open-circuit flux distribution with rare-earth machines using flux focusing structures. As a result, high PM torque and power factor can be obtained. Additional, reluctance torque can also be harnessed for high torque density. And attempts have been made to increase rotor saliency such as

introducing auxiliary poles and splitting main poles to further increase in torque density. Since the length of PM poles to pole pitch ratio is very critical to the flux focusing effect, relative high pole number structure is more preferred for spoke type machines. Compared with PMASynRMs, spoke type configurations show poorer flux weakening capability and thus smaller constant power speed range (CPSR), but in terms of overall efficiency, power factor and torque density, PMASynRMs are inferior to spoke type IPMs.

Other types of machines provide some insight for possible future developments for ferrite PM machines and are not practical for wide applications in the industry at the moment. FSPMs rely highly on strong PM excitation to obtain high torque density and are not quite suitable with low energy ferrite PM. Surface-mounted AFPMs will not be able to achieve relatively high PM field and hence sufficient torque density, while the technology of interior AFPMs is still immature concerning the manufacture and assembly for massive production. AFPM and claw pole machines have 3D flux distributions, and usually need to use SMC materials, which will increase the material cost. As for Vernier machines, they are mostly restrained in low speed high torque applications.

Considering that most of the low efficient IMs used in domestic and industrial applications are medium speed machines, the spoke type IPM machine with advantages of high torque density, power factor and efficiency show higher potential for high-performance requirements. To take full advantages of the reluctance torque of IPM machines, multiple layer concept of PMASynRM can be introduced into the spoke type design to increase the torque density further. Due to the relatively weak magnetisation of ferrite materials, demagnetisation should be paid special attention to so that the durability can be insured without degradation.

2.4 Conclusion

The demand for the highly efficient electric machine is ever increasing as both energy consumption and CO₂ emission can be greatly reduced. The use of DC machines in high speed large power applications is limited due to the commutators. The widely installed IMs are not able to fulfil the efficiency

requirements, and reluctance machines are still having serious issues on vibrations, acoustic noises and low power factor. PMSMs appear to be the best solutions for high performance requirements. To counter the high cost and uncertain supply of rare-earth magnets, literature shows that the low-cost ferrite magnets with abundant supplies are considered to be the most attractive solution for high performance and low cost.

In this chapter, an extensive review of major machine types are carried out to identify the most potential structure for high performance electric machines. Then current development of ferrite PM machines was investigated searching for low-cost high-efficiency candidates, to mitigate disadvantages of the high cost and unstable market of rare-earth magnets. The focus was concentrated on torque density and efficiency. According to the existing literature, ferrite spoke type and PMASynRM machines are the most promising candidates to replace currently widely used low efficient IMs. PMASynRMs rely mainly on reluctance torque and thus have better high-speed performances. Spoke type machines adapt the flux-squeeze structure for flux concentration to compensate for the low residual flux density of ferrite material and exhibit higher torque density, power factor and efficiency. Considering the operating conditions and requirements of various industrial applications, the spoke type IPM shows better overall performance. Meanwhile, the multi-flux-barrier concept of PMASynRMs can be introduced to spoke type machine design to increase the salient pole ratio for higher reluctance torque. With carefully designed rotor structures, there is still potential to improve the saliency of the spoke type configuration and thus the torque density.

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3 DESIGN OF TWO-LAYER SPOKE TYPE MACHINES

3.1 Introduction

The spoke-type configuration is considered as a potential solution for high-performance low-cost ferrite PM design. Due to the flux focusing techniques, it is possible to achieve high flux distribution with low strength ferrite PM materials. However, it is still difficult to achieve comparable performances to rare-earth machines because of the large disparity of magnetic strength between ferrite and rare-earth magnets. The two-layer configurations, which are developed on the base of the spoke-type structure as shown in Figure 3 1 , maximize the rotor saliency while retaining the flux focusing structure, and are able to synergize the torque output for compelling performances. The electromagnetic features of the machines are highly dependent on the key design parameters such as the number of PM layers, the size and location of the PMs. It would be of particular interest to reveal the impacts of these parameters over the PM and reluctance torque components, and provide an insightful understanding of their influences on the overall electromagnetic torque generation

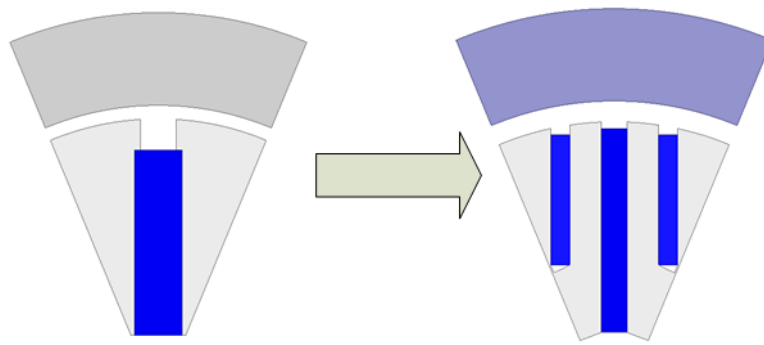


Figure 3-1 The two-layer configurations base of the spoke-type structure

A general way of developing analytical methods is based on Maxwell's equations by solving Laplace's or Poisson's equations [1-3]. Since IPM machines have their PM poles buried inside the rotor, the boundary conditions would be too complicated to solve the equations, especially for the reactive armature field. The open-circuit flux distribution and back EMF are derived by Laplace's equation in [1], but the armature inductance is not included. Since the inductance plays an

important role in IPM machines, PM exciting field and back EMF are not enough to determine the performance. Hence, this method is more popular with surface-mounted or surface-inset PM machines whose structures are much simpler [2-4].

Meanwhile, the equivalent magnetic circuit models were also widely used because of simplicity and fair accuracy [5]. The lumped parameter models based on a magnetic circuit can handle the saturation and nonlinear B-H curve by using FEA coefficients [6] or iterative calculations [7] for higher accuracy. However, FEA models are still necessary to extract saturation coefficients, which vary due to different saturation levels under various load conditions. The saturation coefficients could be different when some key design parameters change. Thus, a considerable amount of simulations are still necessary from FEA software. On the other hand, significant simplification can be made by neglecting saturation, and the analytical model could achieve clear relational understanding. Although error could be caused by making the assumptions, the analytical model should be fairly acceptable to provide theoretical guidance for preliminary design and optimization. Then further modifications can be carried out based on the preliminary designs and a significant amount of time and effort could be saved.

This chapter demonstrates the design procedure for a spoke-type two-layer machine configuration. First, analytical models based on the d-q frame are developed. In addition to fundamental components, higher order components are also considered for more accuracy prediction of the machine performances. Electromagnetic design and optimisations are carried out for the spoke type two-layer IPM machines. The torque density is maximised based on the developed analytical models and FEA simulations. The influence of the key design parameters is comprehensively investigated to provide helpful insight into the design of multi-layer IPM ferrite machines. Finally, the optimum design is chosen for the prototype machine.

3.2 Analytical Models

3.2.1 General Assumptions

The objective of the analytical model is to use a fast and straightforward way for the initial design. Therefore, some assumptions were made to simplify the calculations during the derivation of the analytical method.

The relative permeability of steel lamination material is higher than that of air thousands of times. Thus, the drop in magnetomotive force (MMF) in the lamination can be negligible compared with the drop-in air gap. Therefore, the permeability of the steel is considered to be infinite. So, all the flux lines go radially in the airgap and purely tangential to the eddy of lamination cores.

Moreover, the stator slot, the stator is assumed to be smooth because its slot usually is very small, Carter's coefficient presents the equivalent airgap length to the justification for the influence of stator slot openings. Carter's coefficient k_c can be obtained as follows [8],

$$k_c = \frac{\tau_s}{\tau_s + (b_0^2 / (b_0 + 5g))} \quad (3-1)$$

$$\tau_s = \frac{2\pi R_{si}}{N_t} \quad (3-2)$$

Where τ_s pitch of stator tooth, b_0 slot opening of the stator, g airgap length, N_t the number of teeth and R_{si} inner radius of the stator. Accordingly, the equivalent length of the is obtained as follows

$$g_c = k_c \cdot g \quad (3-3)$$

On the other hand, the rotor PM slot opening is usually much larger comparing to that of the stator slots; the flux distribution would be severely distorted. According to [9], the flux lines go through the airgap radially and reaches the slot edge tangentially in the rotor PM slot, as demonstrated in Figure 3-2(a). The flux lines will be arcs whose centres locate at the crossover point of the airgap edge and slot edge. Thus, the flux line distance within the slot area at x becomes $(g_c + \pi x / 2)$.

Consequently, the permeance along the airgap at a different position can be displayed in Figure 3-2(b). In equivalent magnetic theory, if excitation is kept constant, the flux density is proportional to permeance. Therefore, the flux density decreases significantly in the PM slot area and is even negligible if simpler models are required.

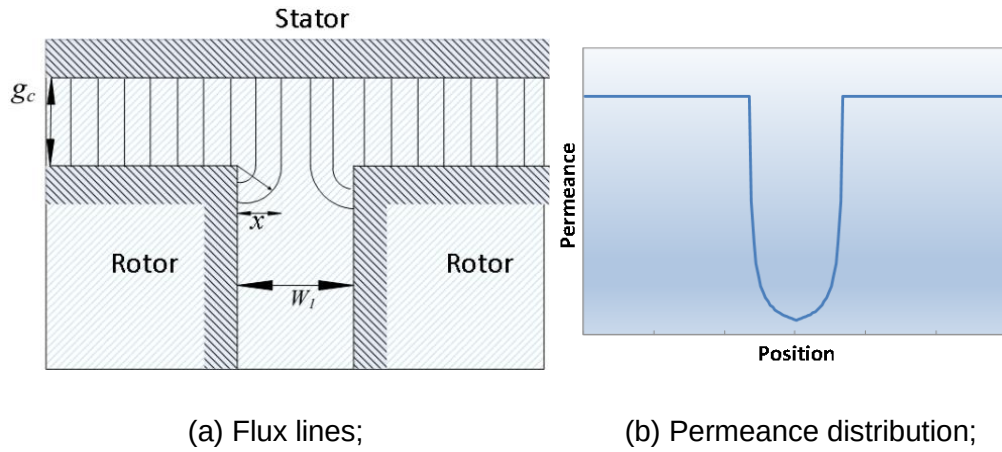


Figure 3-2 Flux lines and the permeance distribution in the airgap.

The armature current in the windings generate stator MMF and this current and windings distribution are the essentials to calculate the armature reaction. Only the fundamental component of the armature current is considered (purely sinusoidal). The initial phase angle is assuming zero for winding a , and the three-phase system is balanced. So, armature current can be written as

$$\begin{cases} i_a(t) = I_m \cos(\omega t) \\ i_b(t) = I_m \cos(\omega t - 2\pi/3) \\ i_c(t) = I_m \cos(\omega t + 2\pi/3) \end{cases} \quad (3-4)$$

where I_m is the magnitude of the armature current, ω is the electric angular velocity.

According to [10], the winding function is defined to describe any actual winding configurations. As the three phase windings are evenly distributed with a displacement of $2\pi/3$ radians among one another, the winding function for three phase windings can be expressed in Fourier series:

$$\begin{cases} N_{fa}(\Theta) = \sum_v \frac{2}{\pi v} N_p k_{wv} \cos v(\Theta) \\ N_{fb}(\Theta) = \sum_v \frac{2}{\pi v} N_p k_{wv} \cos v\left(\Theta - \frac{2}{3}\pi\right) \\ N_{fc}(\Theta) = \sum_v \frac{2}{\pi v} N_p k_{wv} \cos v\left(\Theta + \frac{2}{3}\pi\right) \end{cases} \quad (3-5)$$

where Θ is the mechanical position of the stator, N_p is the number of turns in series per phase, k_{wv} is winding factor for the v th MMF space harmonic component, and v odd harmonics =1, 3, ...

And the synthetic MMF produced by the armature windings is expressed by

$$\begin{aligned} F_m &= N_{fa}(\Theta) i_a(t) + N_{fb}(\Theta) i_b(t) + N_{fc}(\Theta) i_c(t) \\ &= \sum_{v \neq mk} \frac{m}{\pi v} N_p I_m k_{wv} \cos(v\Theta \mp \omega t) \end{aligned} \quad (3-6)$$

Where m is the number of phases, N_p is the number of turns per phase in series, and $k=1, 2, 3, \dots$

In general, the synchronous component is considered as the fundamental, and electric position is more convenient for the analysis of the electric machine. In a machine with p poles, there are $p/2$ electric cycles. Thus, the spatial harmonic order v defined in a multi-pole machine and the relation between the mechanical position and electric position are described by

$$\begin{cases} v = 2v / p \\ \Theta = 2\theta / p \end{cases} \quad (3-7)$$

As the triplen of spatial harmonics synthesised by armature windings is equal to zero in the three phase machines, Equation (3-6) can be written as

$$F_m = \sum_{v=\frac{6k-4}{p}} \frac{2m}{\pi v p} N_a I_m k_{wv} \cos(v\theta - \omega t) + \sum_{v=\frac{6k-2}{p}} \frac{2m}{\pi v p} N_a I_m k_{wv} \cos(v\theta + \omega t) \quad (3-8)$$

To verify the analytical method, FEA models are necessary to evaluate the accuracy of the developed models. Hence, examples are chosen for the

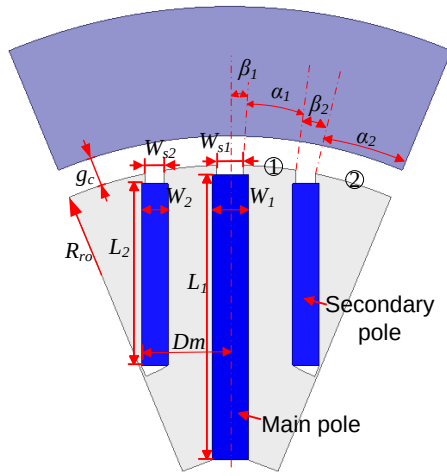
convenience of building FEA models. The general sizing parameters are given in Table 3-1. The dimensions and speed in this example are based mainly of the IM motor that the new machine will compare with.

Table 3-1 Common dimensional properties and specifications

Parameters	Value	Unit
Stator Outer Diameter (D_{so})	160	mm
Rotor Outer Diameter (D_{ro})	94.5	mm
Airgap Length (g)	0.25	mm
Stack Length (l_{ef})	90	mm
Base Rotating Speed	1500	rpm
Number of Poles (p)	8	
Number of Slots (q)	48	
PM Material	Y40	

The cross-section of the two-layer spoke type machine and the dimensional definitions for the analytical models are demonstrated in Figure 3-3. The size of the motor can be reduced to reduce the size of the problem using 2D FEA. The motor has 3 pairs of poles. Only one hight of the motor can be used. This is valid because the stator has 36 slots (6 is a divider of 36) and the three phase winding has also a periodicity of 60 degrees. Each rotor pole consists of three magnets, which are the main PM with circumferential magnetisation located in the middle, and two secondary magnets on both sides of the main PM with the same magnetisation direction. The neighbouring V-shaped poles are magnetised in opposite directions to form a flux focusing structure to increase the airgap flux density. Because secondary PM layers are introduced to the conventional spoke

type structure, the main flux is divided into two portions, as illustrated in Figure 3-3(b). The area ① represents for the rotor



(a) Machine model

(b) Dimensional parameters

Figure 3-3 Dimensions of the two-layer machine model.

lamination area in between the main pole and one of its neighbouring secondary poles, while area ② denotes the region between two adjacent V-shaped PMs. R_{ro} is the outer radius of the rotor, g_c is the length of the airgap, W_1 and L_1 are the width and length of the main PM poles, W_2 and L_2 are width and length of the secondary PM poles, W_{s1} , L_{s1} , W_{s2} and L_{s2} are the width and depth of the main and secondary PM slot opening respectively, D_m is the location of the secondary pole, β_1 is the mechanical radians of half of the main PM slot opening, and β_2 , α_1 and α_2 are the mechanical radians of the secondary PM slot openings, area ① and ② respectively. Compared to spoke-type, the two-layer rotor structure is more complicated. Because the rotor lamination area has been divided into two parts apart from the PM slot areas, more flux loops will be created. By proposing this novel rotor structure, more sinusoidal back-EMF waveform and higher rotor salient ratio are expected to improve the performance of ferrite PM machines.

3.2.1.1 Open-circuit flux and airgap flux distribution

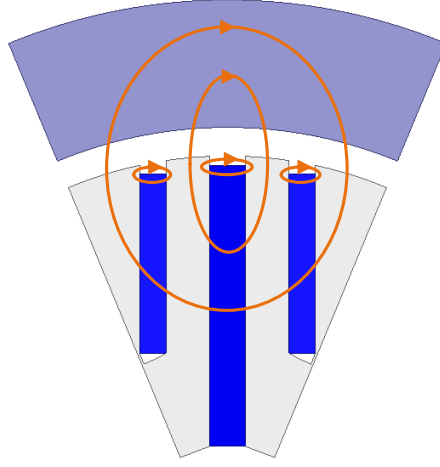


Figure 3-4 Flux loops excited by PMs of two-layer configurations.

The flux loops excited by PMs are depicted in Figure 3-4. The main poles have a major impact on the flux distributions because all flux goes through the main pole except the flux leakage of the secondary pole. According to Ampere's law and flux conversion law, the flux of the main and leakage loops must fulfill the following equations:

$$\begin{cases} H_{m1} \cdot W_1 + 2 \cdot H_{g1} \cdot g_c = 0 \\ H_{m1} \cdot W_1 + H_{s1} \cdot W_{s1} = 0 \\ H_{m1} \cdot W_1 + 2H_{m2} \cdot W_2 + 2H_{g2} \cdot g_c = 0 \\ H_{m2} \cdot W_2 + H_{s2} \cdot W_{s2} = 0 \end{cases} \quad (3-9)$$

$$\begin{cases} \phi_{m1} = \phi_{g1} + \phi_{g2} + \phi_{ml1} \\ \phi_{m2} = \phi_{g2} + \phi_{ml2} + \phi_{bri} \end{cases} \quad (3-10)$$

Where H_{m1} and H_{m2} are the flux intensity of the main and secondary PMs, H_{g1} and H_{g2} are the flux intensity in the airgap area ① and ②, H_{s1} and H_{s2} are the flux intensity inside the main and secondary PM slots respectively, and ϕ_{m1} , ϕ_{m2} , ϕ_{g1} , ϕ_{g2} , ϕ_{ml1} , ϕ_{ml2} and ϕ_{bri} are the flux through the main PM, secondary PM, area ①, area ②, main PM slot, secondary PM slot and the saturation bridge respectively. Considering $W_1=W_{s1}$ and $W_2=W_{s2}$, the airgap flux density of the area ① and ② in the airgap can be derived Appendix A:

$$B_{g1} = \frac{B_r [L_1 (\mu_m L_2 + L_{s2}) W_1 g_c + L_1 W_1 W_2 R_{ro} \alpha_2 - L_2 W_1 W_2 R_{ro} \alpha_2]}{(\mu_m L_2 + L_{s2}) W_1 g_c R_{ro} \alpha_2 + [(\mu_m L_2 + L_{s2}) g_c + W_2 R_{ro} \alpha_2] [2(\mu_m L_1 + L_{s1}) g_c + W_1 R_{ro} \alpha_1]} \quad (3-11)$$

$$B_{g2} = \frac{B_r \cdot [L_1 (\mu_m L_2 + L_{s2}) W_1 g_c + 2(\mu_m L_1 + L_{s1}) L_2 W_2 g_c + L_2 W_1 W_2 R_{ro} \alpha_1]}{(\mu_m L_2 + L_{s2}) W_1 g_c R_{ro} \alpha_2 + [(\mu_m L_2 + L_{s2}) g_c + W_2 R_{ro} \alpha_2] [2(\mu_m L_1 + L_{s1}) g_c + W_1 R_{ro} \alpha_1]} \quad (3-12)$$

Where B_r is nominal residual flux density.

If the flux in the PM slot areas should be taken into consideration, Equation (3-11) and (3-12) can be written as

$$B_{g1} = \frac{B_r L_1 W_1 (\mu_m L_2 g_c + 2L_{s2} g_c + W_2 R_{ro} k_{rs2} \alpha_2) - B_r L_2 W_1 W_2 R_{ro} k_{rs2} \alpha_2}{\mu_m L_2 W_1 g_c R_{ro} k_{rs2} \alpha_2 + (\mu_m g_c L_2 + g_c L_{s2} + W_2 R_{ro} k_{rs2} \alpha_2) (2\mu_m L_1 g_c + 2L_{s1} g_c + W_1 R_{ro} k_{rs1} \alpha_1)} \quad (3-13)$$

$$B_{g2} = \frac{B_r \cdot [\mu_m L_1 L_2 W_1 g_c + L_2 W_2 (2\mu_m g_c L_1 + 2g_c L_{s1} + W_1 R_{ro} k_{rs1} \alpha_1)]}{\mu_m L_2 W_1 g_c R_{ro} k_{rs2} \alpha_2 + (\mu_m L_2 g_c + L_{s2} g_c + W_2 R_{ro} k_{rs2} \alpha_2) (2\mu_m L_1 g_c + 2L_{s1} g_c + W_1 R_{ro} k_{rs1} \alpha_1)} \quad (3-14)$$

where k_{rs1} and k_{rs2} are the rotor slot coefficient for PM slots, and

$$k_{rs1} = 1 + \frac{2g}{\pi R_{ro} \alpha_1} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_1}{2g} \right) + \frac{2g}{\pi R_{ro} \alpha_1} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_2}{4g} \right) \quad (3-15)$$

$$k_{rs2} = 1 + \frac{4g}{\pi R_{ro} \alpha_2} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_2}{4g} \right) \quad (3-16)$$

$$\begin{cases} \beta_1 = \arcsin \frac{W_{s1}}{2R_{ro}}, & \beta_2 = \arcsin \frac{D_m}{R_{ro}} - \arcsin \frac{D_m - W_{s2}}{R_{ro}} \\ \alpha_1 = \arcsin \frac{D_m - W_{s2}}{R_{ro}} - \beta_1, & \alpha_2 = \frac{\pi}{p} - \arcsin \frac{D_m}{R_{ro}} \end{cases} \quad (3-17)$$

Based on the B_{g1} and B_{g2} , the waveform of the flux distribution in the airgap can be obtained. However, it is more convenient to express the flux distribution in the Fourier series to assess the fundamental component as well as higher order space harmonics. By applying a Fourier transform, the fundamental component and higher order harmonics can be achieved. If infinite reluctance in the PM slot area is assumed, the magnitude of the v th harmonic component is expressed by

$$B_{fgv} = \frac{4}{v\pi} B_{g2} \sin\left(v \frac{p\alpha_2}{2}\right) + \frac{4}{v\pi} B_{g1} \left[\sin\left(v \frac{\pi - p\beta_1}{2}\right) - \sin\left(vp \frac{\alpha_2 + \beta_2}{2}\right) \right] \quad (3-18)$$

With the consideration of the flux in the PM slot areas, the v th flux component is calculated by

$$B_{fgv} = \frac{4}{v\pi} B_{g2} \sin\left(\frac{vp\alpha_2}{2}\right) + \frac{4}{v\pi} B_{g1} \left[\sin\left(v \frac{\pi - p\beta_1}{2}\right) - \sin\left(vp \frac{\alpha_2 + \beta_2}{2}\right) \right] + \frac{4pg_c}{\pi^2 R_{ro}} \cdot \left\{ \begin{aligned} & B_{g1} \cos v \left(\frac{\pi}{2} - \frac{pg_c}{\pi R_{ro}} - \frac{p\beta_1}{2} \right) \left\{ \text{Ci} \left[v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\beta_1}{2} \right) \right] - \text{Ci} \left[v \left(\frac{pg_c}{\pi R_{ro}} \right) \right] \right\} \\ & - B_{g1} \sin v \left(\frac{\pi}{2} - \frac{pg_c}{\pi R_{ro}} - \frac{p\beta_1}{2} \right) \left\{ \text{Si} \left[v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\beta_1}{2} \right) \right] - \text{Si} \left[v \left(\frac{pg_c}{\pi R_{ro}} \right) \right] \right\} \\ & + \left[B_{g1} \cos v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\alpha_2}{2} + \frac{p\beta_2}{2} \right) \right] \left\{ \text{Ci} \left[v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\beta_2}{4} \right) \right] - \text{Ci} \left[v \left(\frac{pg_c}{\pi R_{ro}} \right) \right] \right\} \\ & + B_{g2} \cos v \left(\frac{p\alpha_2}{2} - \frac{pg_c}{\pi R_{ro}} \right) \left\{ \text{Ci} \left[v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\beta_2}{4} \right) \right] - \text{Ci} \left[v \left(\frac{pg_c}{\pi R_{ro}} \right) \right] \right\} \\ & + \left[B_{g1} \sin v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\alpha_2}{2} + \frac{p\beta_2}{2} \right) \right] \left\{ \text{Si} \left[v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\beta_2}{4} \right) \right] - \text{Si} \left[v \left(\frac{pg_c}{\pi R_{ro}} \right) \right] \right\} \\ & - B_{g2} \sin v \left(\frac{p\alpha_2}{2} - \frac{pg_c}{\pi R_{ro}} \right) \left\{ \text{Si} \left[v \left(\frac{pg_c}{\pi R_{ro}} + \frac{p\beta_2}{4} \right) \right] - \text{Si} \left[v \left(\frac{pg_c}{\pi R_{ro}} \right) \right] \right\} \end{aligned} \right\} \quad (3-19)$$

where

$$\text{Si}(x) = \int_0^x \frac{\sin t}{t} dt = \sum_{k=1}^{\infty} (-1)^{k-1} \frac{x^{2k-1}}{(2k-1)(2k-1)!} \quad (3-20)$$

$$\text{Ci}(x) = -\int_x^{\infty} \frac{\cos t}{t} dt = \gamma + \ln x + \sum_{k=1}^{\infty} \frac{(-x^2)^k}{(2k)(2k)!} \quad (3-21)$$

and γ is Euler–Mascheroni constant.

The flux distributions and spectra from analytical and FEA models are compared in Figure 3-5 with a smooth stator for model verification. The flux density in the core area is slightly higher for the analytical model because all the main flux is assumed to go through this area. High accuracy for the fundamental is predicted

with an error less than 1% even though infinite reluctance in the PM slot areas is presumed. However, higher errors are observed for the harmonics as the flux in the PM slot area may not be negligible for them. On the other hand, with a more complicated approximation of the flux distribution in the PM slot areas, higher accuracy can be achieved even for the harmonics by an improved analytical model.

Since stator slots are necessary to accommodate windings, the impact over flux distribution is also discussed. Figure 3-6 depicts the flux distributions with smooth and actual slotted stators. The flux distribution waveform can be further improved by considering the stator slots. But, with the introduction stator slotting effect, noticeable improvements are only observed for the tooth harmonics such as the 11th and 13th order in this case. In light of the complexity and minor improvement, the stator slotting effect is simplified with Carter's coefficient.

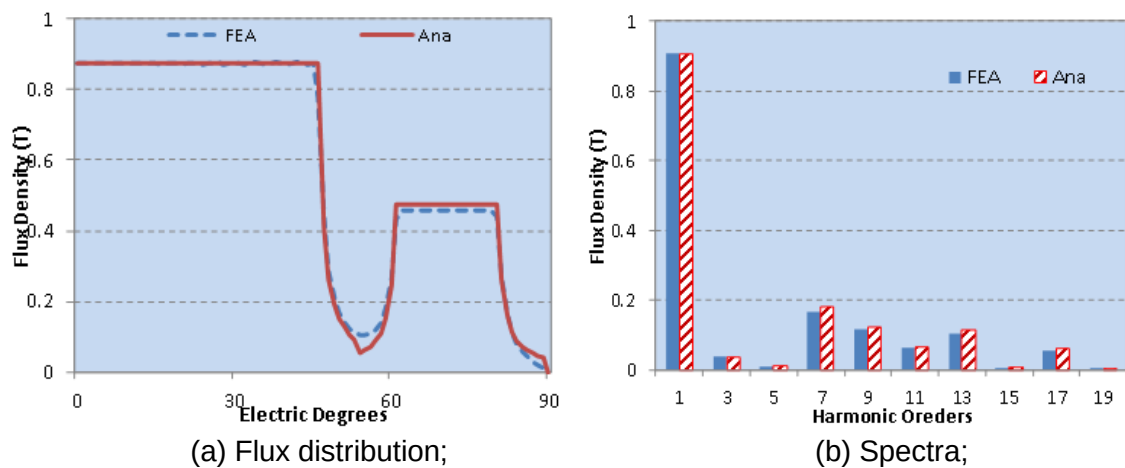


Figure 3-5 of PM flux distribution and spectra in the airgap with the smooth stator.

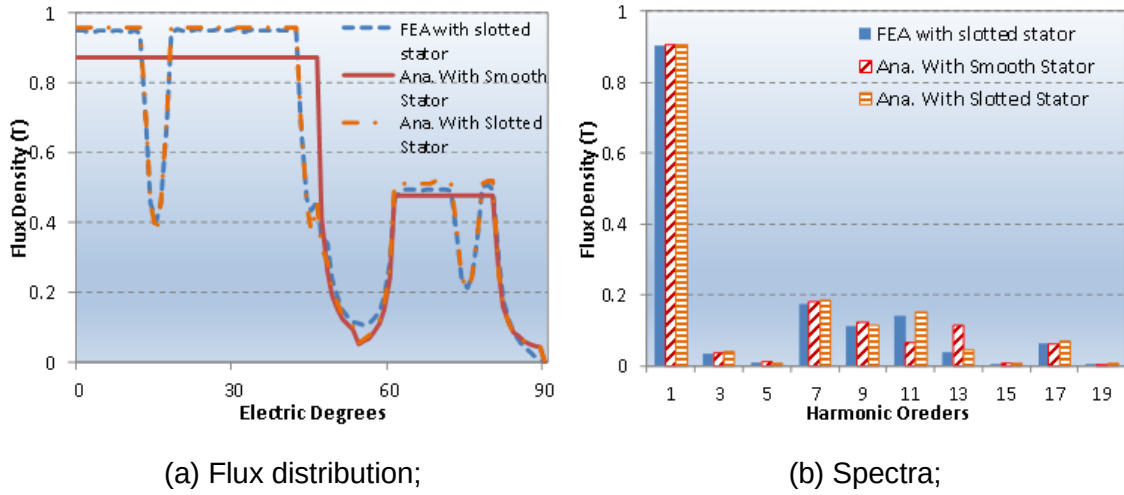


Figure 3-6 Comparison of PM flux distribution and spectra in the airgap with the slotted and smooth stator.

To get a clearer understanding of the relations of the key parameters and flux distribution, some approximation and simplification are made to obtain simpler equations. On the preconditions of much larger PM width than airgap length and zero flux in the PM slot area, Equation (3-18) can be simplified as:

$$B_{fgv} \approx \frac{4B_r}{v\pi} \left\{ \frac{L_1 - L_2}{R_{ro}\alpha_1} \left[\sin\left(v \frac{\pi - p\beta_1}{2}\right) - \sin\left(vp \frac{\alpha_2 + \beta_2}{2}\right) \right] + \frac{L_2}{R_{ro}\alpha_2} \sin\left(v \frac{p\alpha_2}{2}\right) \right\} \quad (3-22)$$

Each flux component can be regarded as an addition of two parts, the flux distribution in the core area ① and ②. The term $R_{ro}\alpha_2$ is the arc length of the core area ②. Hence, B_{g2} can be regarded as proportional to secondary PM length, which represents the secondary PM flux, and inversely proportional to area ② arc length, which represents the area. Since L_1 decides the overall PM flux and L_2 decides the PM flux in area, the flux in the core area ① is proportional to $(L_1 - L_2)$. Moreover, B_{g1} is approximated by the flux difference between the main and secondary poles divided by the arc length of the core area ①. Therefore, the flux distribution can be adjusted by changing the length (or location) of the secondary poles. A two-stepped waveform of flux distribution can be obtained in the airgap with less harmonic content.

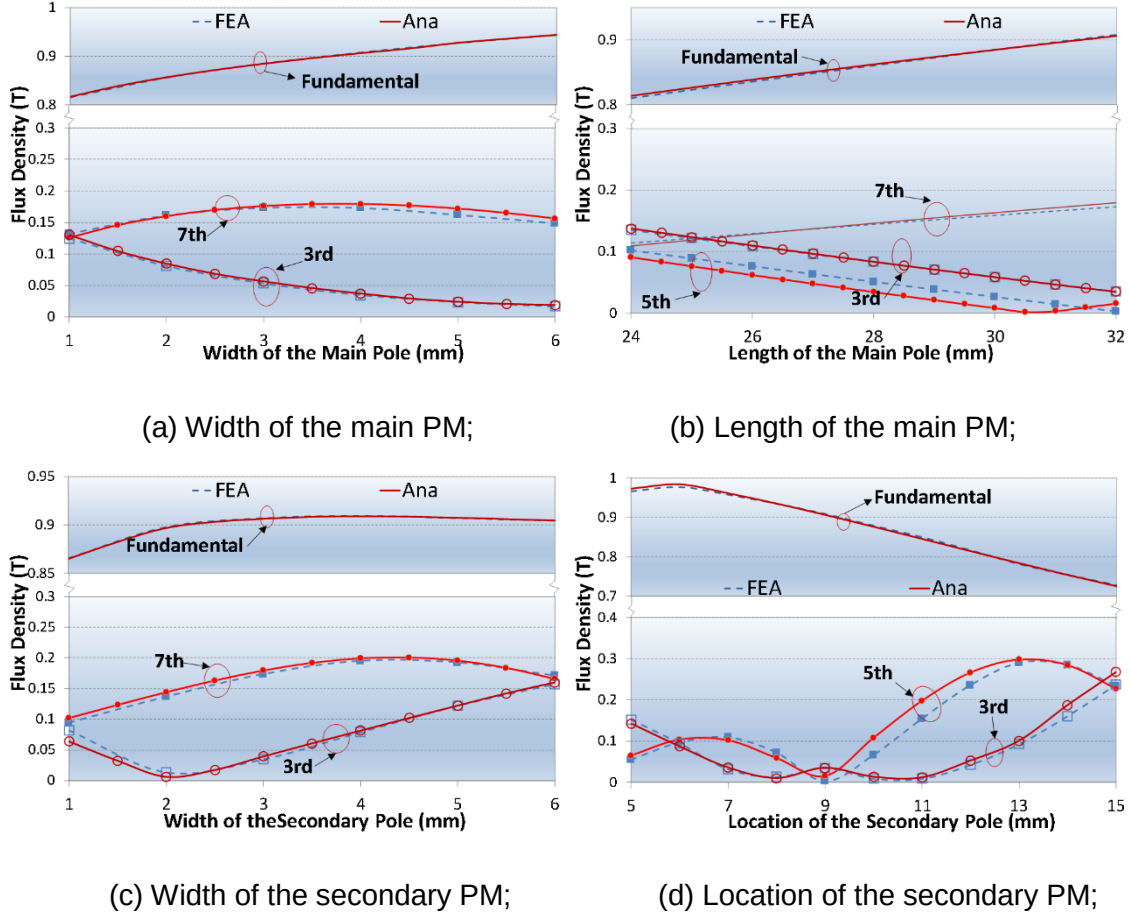


Figure 3-7 Influence of the size and location of the main and secondary PMs on flux distribution.

The fundamentals can be obtained by Equation (3-22) when $v=1$. Since all the sine terms are positive, the fundamental always increases with the width of the main PMs, as shown in Figure 3-7(a). That is because both the working points of the PMs and flux focus effect could be improved by thicker PM poles. More importantly, the PM width affects the PM slot opening and thus has considerable influence on the harmonics. The width of PMs should be chosen with the criteria of relatively high fundamental and low harmonics.

Equation (3-22) indicates that the length of the main PMs is proportional to the flux focusing effect and can increase the flux density effectively, as demonstrated in Figure 3-7(b). Because the two-layer structure is developed from the one-layer configuration, the main pole has a similar effect on flux focusing. As the sine terms could be positive or negative for different order harmonics, their amplitudes may increase or decrease with the main PM length. However, its impact on the

harmonics is small compared with other factors. Since the main PM length is the key factor to the flux focusing effect, longer main PMs are preferred as long as there is enough space inside the rotor, and the harmonic content should be adjusted by modifying other sizing parameters.

The influence of the secondary PMs is quite different from the main PMs. The secondary poles play a more active role in the flux allocation between the two core areas, as shown in Figure 3-7 (c) and (d). The width of the secondary PM does not have to be too thick, because its impact on intensifying the fundamental is limited. The prime consideration should be on the harmonic content when choosing the width of secondary PMs.

The most active and complicated factor for the airgap flux distribution has been the location of secondary PMs. By changing the location between the main and secondary poles, the arc length of the rotor lamination surface of the area ① and ② together with the length of secondary poles are all changed. That will affect not only the amount of flux through the area ② but also the shape of flux distribution waveforms. Therefore, both the fundamental and harmonics are greatly affected, as shown in Figure 3-7(d). As the secondary PM moves towards the main PM, D_m decreases and L_2 increases because the secondary pole can go deep inside the rotor. Therefore, the total amount of PM increases and so does the airgap flux density. That is the reason why the fundamental flux decreases with the decrease of D_m . But when the secondary PM is very close to the main pole, the increment of L_2 and flux density becomes very small. Also, due to the increase of harmonic content, the fundamental flux may decline. In all, the fundamental flux distribution increases with smaller D_m . And when the secondary pole touches the main PM, the rotor becomes a one-layer structure, which indicates that potentially higher flux density can be achieved for the one-layer configuration.

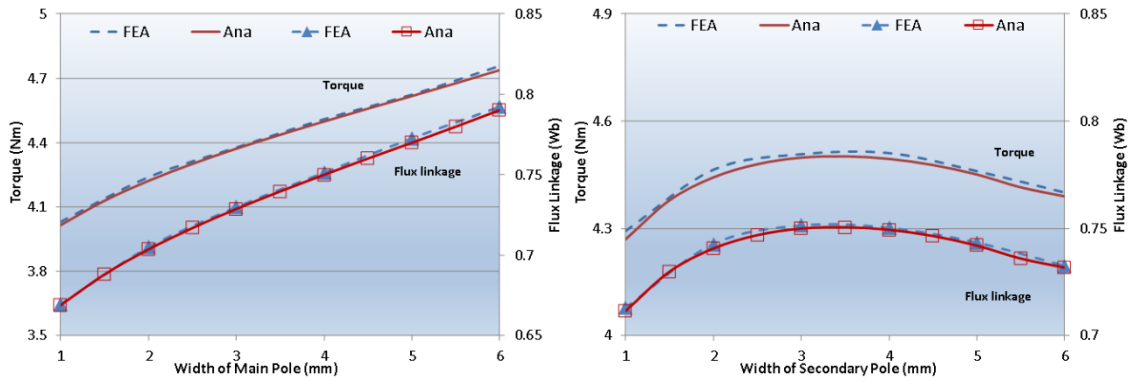
Meanwhile, the harmonic contents are much more sensitive to the location of secondary PM. Owing to the re-allocation of the PM flux between area ① and ②, the square flux waveform of the one-layer machine is modified into a two-stepped waveform for the two-layer structure with lower harmonic content. However, if

over adjusted, the flux distribution could be deteriorated with dramatically increased harmonics.

Based on the analytical models developed, the PM flux linkage and PM torque can be derived accordingly. Since the harmonic flux components with triplen orders in a three-phase machine are with the same phase angle in a balanced machine, it will not develop flux linkage in the d-q frame. Then the PM flux linkage and PM torque can be expressed as:

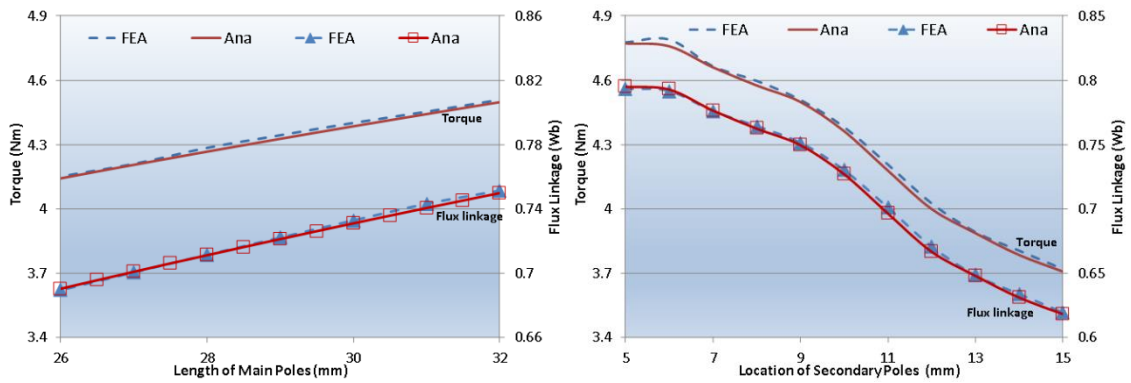
$$\Psi_f = \frac{4R_{ro}l_{ef}N_a}{p} \sum_{v=1,5,7...}^{6k+1} (-1)^{\frac{v-1}{2}} \frac{k_{wv}}{v} B_{fgv} \quad (3-23)$$

$$T_{pm} = \frac{mp}{4} \Psi_f i_q = mR_{ro}l_{ef}N_a i_q \sum_{v=1,5,7...}^{6k+1} (-1)^{\frac{v-1}{2}} \frac{k_{wv}}{v} B_{fgv} \quad (3-24)$$



(a) With different W_1 ;

(b) With different W_2 ;



(c) With different L_1 ;

(d) With different D_m ;

Figure 3-8 PM flux linkage and PM torque with difference PM poles.

The PM flux linkage and PM torque can be obtained by Equation (3-23) and (3-24), and the influence of different parameters on the performance is demonstrated in Figure 3-8. Obviously, both the PM flux linkage and torque share the same trends over different parameters with the fundamental flux

distribution, as normally over 97% of the flux linkage and torque are due to the fundamental flux. As expected, longer PMs can enhance the flux linkage and PM torque effectively and linear relationship is observed between the main PM length and PM flux linkage (torque). Thicker main PMs can also increase PM flux linkage and torque in terms of high working point and flux density, but not as effective as increasing PM length considering the amount of ferrite used. Since compositions of the fundamental and harmonic contents of the flux density are quite sensitive to the location of the secondary pole, PM flux linkage and torque are all affected greatly. Smaller D_m can increase the secondary pole length, and thus higher PM flux and PM torque are achieved. This indicates that conventional spoke-type configuration is able to achieve high PM flux and thus PM torque. But the harmonic content is relatively high, which would cause higher loss and torque ripple. With carefully designed secondary PM layer, the flux distribution waveform in the airgap can be improved with less harmonic content whereas the fundamental can be kept similar to the conventional spoke-type structure. But the secondary PM location also has a prominent influence on rotor saliency, which will affect the saliency of the rotor and be discussed in the following sections.

3.2.1.2 D-Axis Magnetizing Inductance

The magnetic flux loops excited by the d -axis armature current in one pole pitch area is depicted in Figure 3-9. The flux loops are similar to PM excited field except that there is no flux leakage in the rotor PM slots. By applying the Ampere's law and on the flux conversion law, the d -axis MMF and flux through different areas should satisfy the following equations:

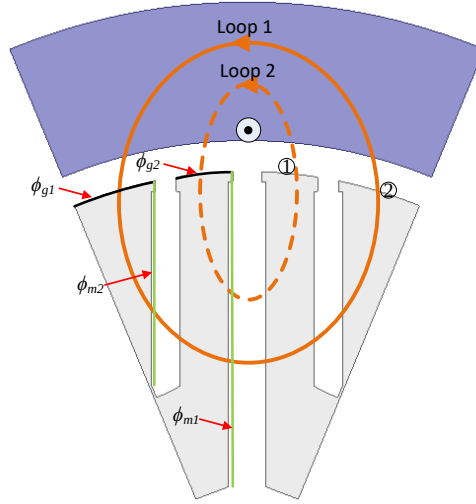


Figure 3-9 Flux loops for two-layer configuration under d-axis current

$$\begin{cases} 2F_{ad}(\theta) = H_{m1d}W_1 + 2H_{gd}(\theta)g_c & \frac{p\alpha_2 + p\beta_2}{2} \leq \theta \leq \frac{\pi}{2} - \frac{p\beta_1}{2} \\ 2F_{ad}(\theta) = H_{m1d}W_1 + 2H_{m2d}W_2 + 2H_{gd}(\theta)g_c & 0 \leq \theta \leq \frac{p\alpha_2}{2} \end{cases} \quad (3-25)$$

$$\begin{cases} \phi_{m1} = \phi_{g1} + \phi_{g2} \\ \phi_{m2} = \phi_{g2} \end{cases} \quad (3-26)$$

where H_{m1d} and H_{m2d} are the flux intensity inside the main and secondary PM slots, ϕ_{m1} and ϕ_{m2} are the flux going across the main and secondary PM slots respectively, $H_{gd}(\theta)$ is flux intensity in the airgap, ϕ_{g1} and ϕ_{g2} is the flux through the rotor core area ① and ② in half a pole pitch, respectively. By solving Equation (3-25) and (3-26), the flux intensity inside the main and secondary PM slots is derived by:

$$H_{m1d} = \frac{\frac{4}{p} \left\{ \sum_{v \neq mk} \frac{F_{adv}}{v} \left[\sin v \left(\frac{\pi}{2} - \frac{p\beta_1}{2} \right) - \sin v \left(\frac{p\alpha_2 + p\beta_2}{2} \right) \right] \right\}}{W_1\alpha_1 + \frac{2g_c(L_1 + L_{s1})}{R_{ro}} + \frac{W_1g_c(L_2 + L_{s2})\alpha_2}{W_2R_{ro}\alpha_2 + g_c(L_2 + L_{s2})}} \quad (3-27)$$

$$H_{m2d} = \frac{\frac{4}{p} \sum \frac{F_{adv}}{v} \left[\sin\left(v \frac{p\alpha_2}{2}\right) \right] - B_{m1d} W_1 \alpha_2}{2W_2 \alpha_2 R_{ro} + 2g_c (L_2 + L_{s2})} R_{ro} \quad (3-28)$$

According to the flux loops demonstrated in Figure 3-9, all main flux travels across the main PM slot, and the flux of the rotor core area ② passes through the secondary PM slot. Thus, two-stepped rotor MMF waveform is obtained by:

$$F_{rs}(\theta) = \begin{cases} H_{m1d} W_1 + 2H_{m2d} W_2 & 0 \leq \theta \leq \frac{p\alpha_2}{2} \\ H_{m1d} W_1 & \frac{p\alpha_2}{2} < \theta \leq \frac{\pi}{2} \end{cases} \quad (3-29)$$

To maintain the unified format for the expressions, the rotor MMF rotor is decomposed into Fourier series:

$$F_{rsv} = \frac{4}{\pi v} \left[H_{m1d} W_1 \sin\left(v \frac{\pi}{2}\right) + 2H_{m2d} W_2 \sin\left(v \frac{p\alpha_2}{2}\right) \right] \quad (3-30)$$

Then the flux density distribution excited by the v th MMF in the lamination pole area is calculated by:

$$B_{adv}(\theta) = \frac{2\mu_0 F_{adv} - \mu_0 F_{rsv}}{2g_c} \cos(v\theta) \quad (3-31)$$

Since the d -axis flux travels across the PM slots inside the rotor, the flux distribution excited by each MMF component will be severely distorted. Together with further deformation caused by PM slot opening in the airgap, $B_{adv}(\theta)$ is not the actual v th harmonic order of the airgap flux. Therefore, Fourier transformation is applied to each armature MMF harmonic field, and the v th harmonic component for the flux distribution is obtained by:

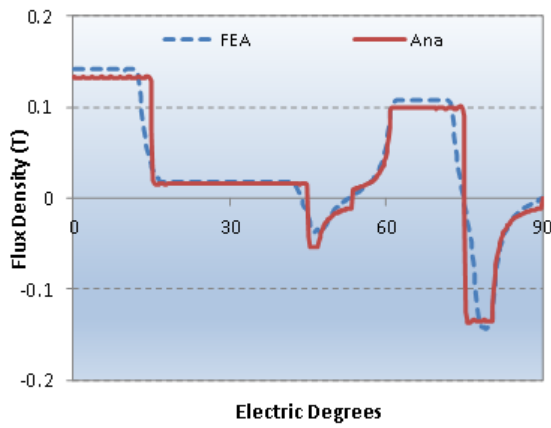
$$B_{gdv} = \sum_{u=1}^{2k+1} \left[\frac{1}{\pi} \int_0^{2\pi} B_{adu} \cos(u\theta) \cos(v\theta) d\theta \right] = \sum_{u=1}^{2k+1} B_{gduv} \quad (3-32)$$

With the flux in the PM slot areas assumed to be zero, the following is derived:

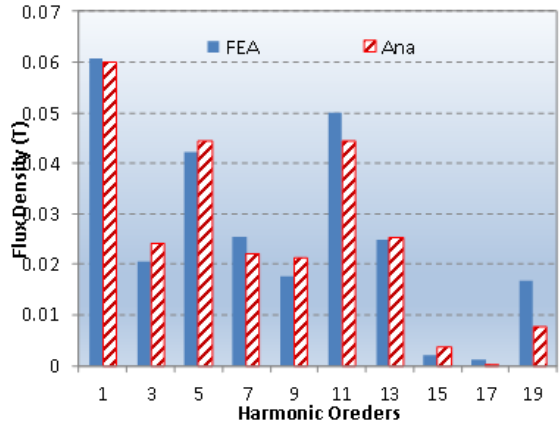
$$\begin{aligned}
B_{gduv}(\theta) &= \frac{4}{\pi} B_{adu} \left[\int_0^{\frac{p\alpha_2}{2}} \cos(u\theta) \cos(v\theta) d\theta + \int_{\frac{p(\alpha_2+\beta_2)}{2}}^{\frac{\pi-p\beta_1}{2}} \cos(u\theta) \cos(v\theta) d\theta \right] \\
&= \begin{cases} B_{adu} \left[1 - \frac{p(\beta_1 + \beta_2)}{\pi} + \frac{\sin v(\pi - p\beta_1) - \sin vp(\alpha_2 + \beta_2) + \sin vp(\alpha_2)}{v\pi} \right] & (u = v) \\ \frac{2}{\pi} B_{adu} \left[\frac{\sin \frac{(u+v)p\alpha_2}{2} + \sin \frac{(u+v)(\pi - p\beta_1)}{2} - \sin \frac{(u+v)p(\alpha_2 + \beta_2)}{2}}{u+v} + \frac{\sin \frac{(u-v)p\alpha_2}{2} + \sin \frac{(u-v)(\pi - p\beta_1)}{2} - \sin \frac{(u-v)p(\alpha_2 + \beta_2)}{2}}{u-v} \right] & (u \neq v) \end{cases} \quad (3-33)
\end{aligned}$$

Where $B_{gdv}(\theta)$ is the v th flux distribution in the airgap, $B_{gdu}(\theta)$ denotes flux distribution due to the u th MMF component, and $B_{gduv}(\theta)$ represents the v th order component of $B_{gdu}(\theta)$. Because of more PM slots for the two-layer structure, the expression will be much more complicated by considering PM slot flux.

The flux distributions and its spectra are illustrated in Figure 3-10. Despite zero flux in the PM slot areas are assumed, the analytical models show good agreement with FEA results. The improved analytical method exhibits higher accuracy in the PM slot areas by considering the flux through them. However, the model is too complicated because of the more intricate geometrical design of two-layer rotor. Usually, the width of PM slot openings are much smaller than the one-layer configuration; the influence on the airgap flux distribution is much lower. As shown in Figure 3-10(b), the spectra analysis indicates that acceptable accuracy can be achieved by the analytical model even with the flux in the PM slots neglected

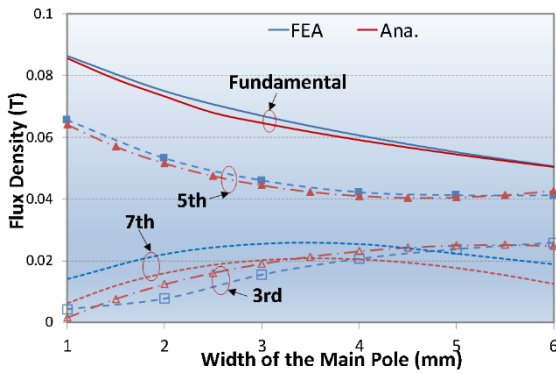


(a) Flux distribution;

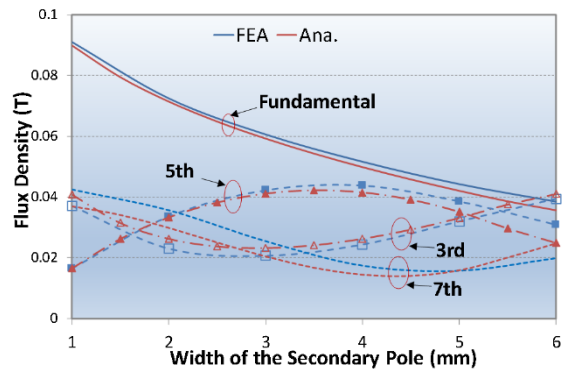


(b) Spectra

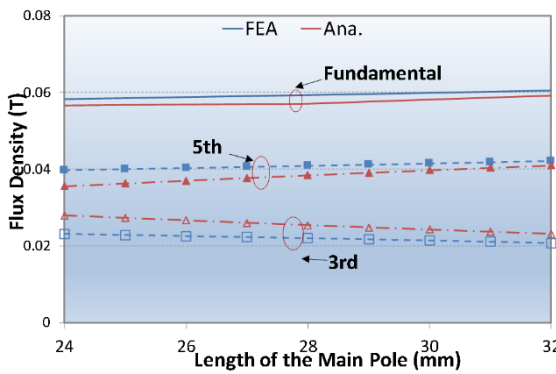
Figure 3-10 Flux distribution and spectra in the airgap excited by d-axis current.



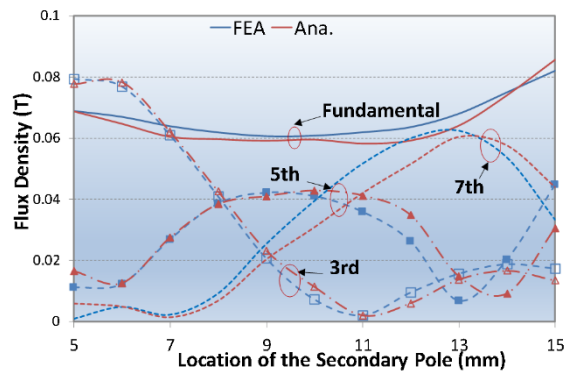
(a) With different W_1 ;



(b) With different W_2 ;



(c) With different L_1 ;



(d) With different D_m ;

Figure 3-11 Fundamental, 3rd and 5th component of flux distribution excited by d-axis current.

Figure 3-11 depicts the fundamental and some relatively lower order harmonic components against different PM parameters. Since the main MMF drop in the flux loops is on the PM slots, a larger PM slot width leads to a significant decrease in fundamental flux. The width of the PMs also exhibits considerable impact on the harmonic contents due to the distortion of the PM slots. The width of the secondary PM shows a higher influence than the main PM due to its more active role in the flux allocation between the two core areas. Since longer PMs would increase the permeance of the PM slots, a subtle increment in flux density can be observed. However, since the width mainly decides the reluctance of the PM slot, the length of the main PM slot has little influence over the flux distribution. The most influential parameter on the flux distribution has been the location of the secondary pole. As the flux waveform will be seriously shaped and the flux through the area ① and ② is regulated, both the fundamental and the harmonic components are significantly affected.

With flux distribution solved, the d -axis magnetising inductance is obtained by:

$$L_{md} = \frac{\Psi_{md}}{i_d} = \frac{4N_a R_{ro} I_{ef}}{p \cdot i_d} \sum_{v \neq m \cdot k} \frac{k_{wv}}{v} B_{gdv} = \sum_{v \neq m \cdot k} L_{mdv} \quad (3-34)$$

And the results of both FEA and analytical models are depicted and compared in Figure 3-12. As the fundamental flux contributes to the major part of the magnetising flux linkage, the similar influence of the design parameters is observed for the inductance. The d -axis magnetising inductance decreases dramatically with a larger PM width of the main and secondary PMs because of larger PM slot reluctance. The length of the main has a subtle impact on the flux distribution excited by armature MMF, and thus there is an only marginal increment in the inductance. In spite that the location of secondary poles has a high impact on the harmonic compositions of the flux distribution, no significant variation is seen for the inductance when D_m is not very large. As the secondary pole moves further away from the main pole, the arc length of the area ② and the length of secondary PMs reduce dramatically. Consequently, most of the main flux goes by Loop 1 through the area ② and only needs to travel across the main PM slot. The reluctance of the flux loops decreases and the inductance increases

significantly. It should be noted that the flux distribution due to the d -axis current contains a high amount of harmonics, which also generates flux linkage in the windings and contributes to inductance. It should be of interest to investigate the contributions by harmonics to make full use of them.

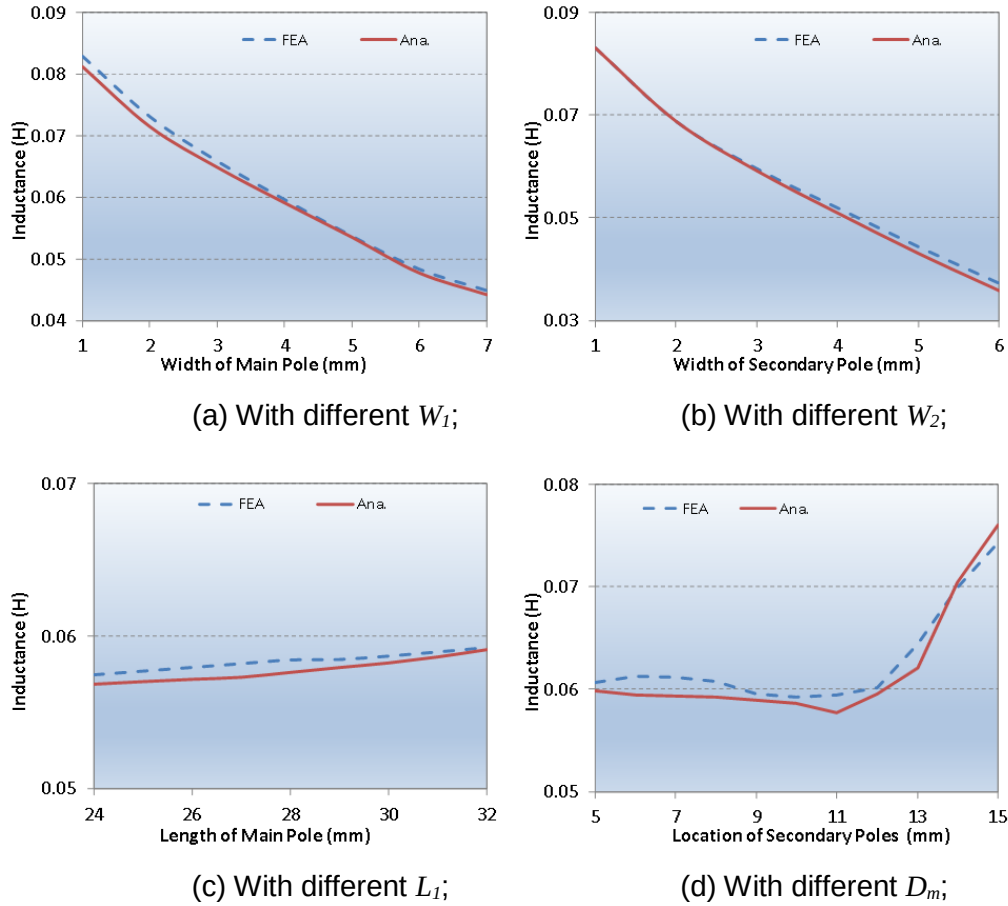


Figure 3-12 Comparison of d -axis magnetizing inductance against different rotor design parameters.

The contributions of flux harmonics to d -axis inductance against different PM parameters are illustrated in Figure 3-13. Due to the relatively large contents of 5th and 7th order harmonics, a considerable part of the magnetising inductance is due to the harmonic flux linkage. Up to 15% of L_{md} could be generated by

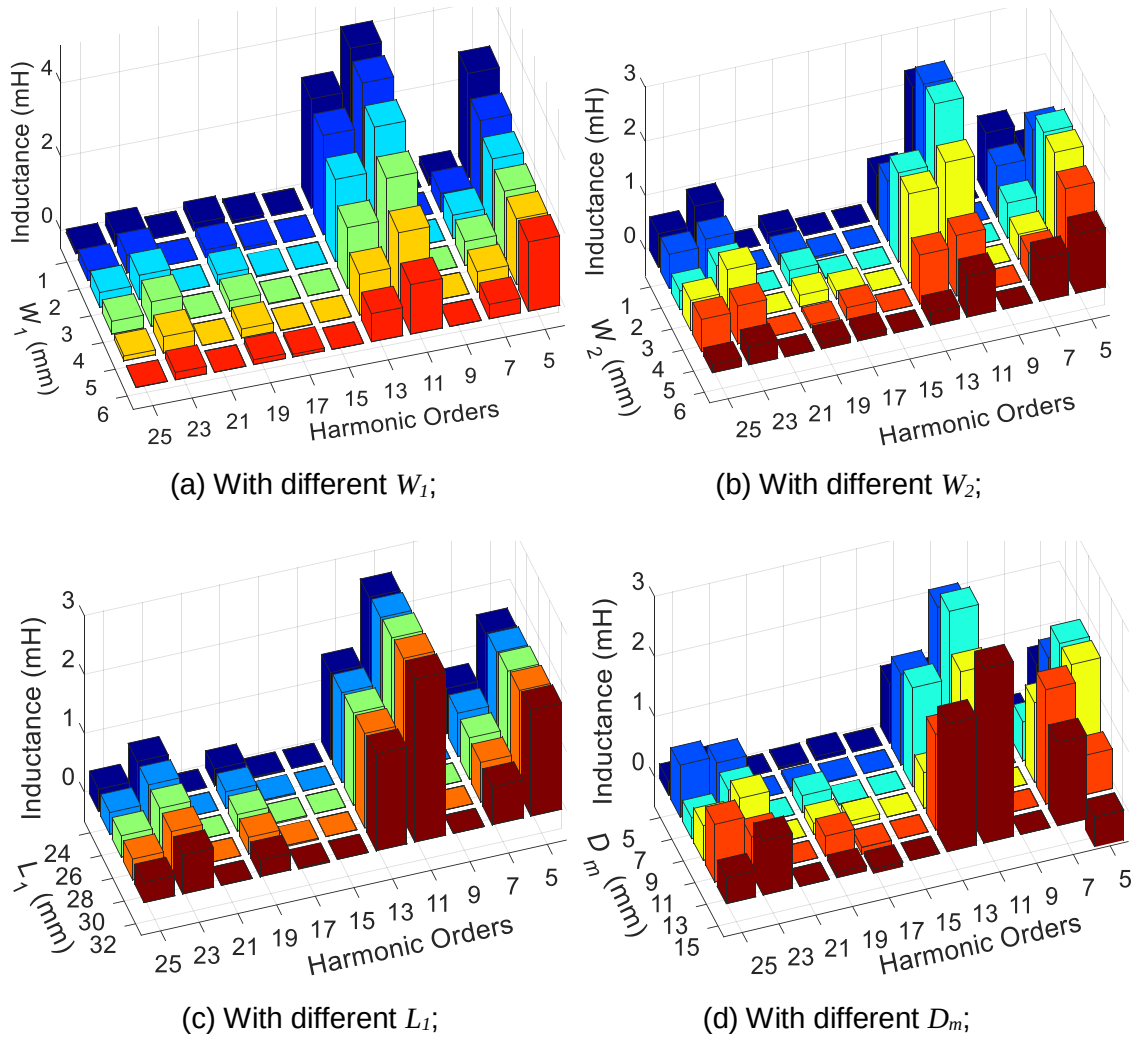


Figure 3-13 Contributions of flux harmonics to d -axis magnetizing inductance.

harmonics, especially from 5th, 7th and tooth harmonic orders such as 11th and 13th. Since all the flux goes through the main PM slot, the harmonics also decreases with W_1 . But the little impact is seen from L_1 . Because a large part of the flux needs to cross the secondary PM slots as well, harmonic contents reduced in general. On the other hand, the location of secondary poles has a huge effect on the harmonic content, and the inductance contribution of different harmonics varies enormously with D_m , which makes this parameter very important to regulate the harmonic contents of the flux distribution.

3.2.1.3 Q-Axis Magnetizing Inductance

The flux loops excited by q-axis current are depicted in Figure 3-14. The main flux does not pass through the PM slot, and the MMF drop occurs only in the

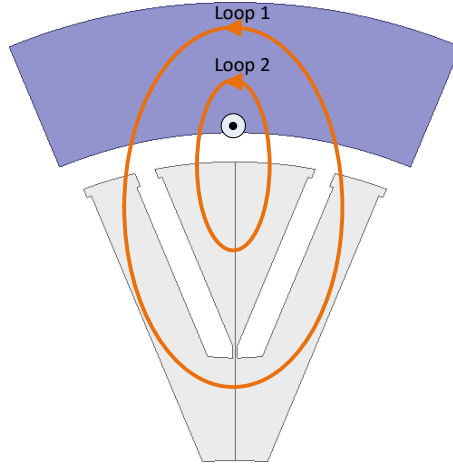


Figure 3-14 Flux line loops for two-layer configuration under q-axis current

airgap. Though two flux loops are displayed in Figure 3-14, their equivalent magnetic circuit models are exactly the same. The two flux loops are actually one, and the flux density in the core areas due to the v th order MMF can be expressed by:

$$B_{aqv}(\theta) = \frac{\mu_0 F_{aqv}(\theta)}{g_c} = \frac{2\mu_0 m N_a k_{wv}}{\pi v p g_c} i_q \cos(v\theta) \quad (3-35)$$

With the flux in the PM slot area neglected, the amplitude of the v th harmonic can be derived by Fourier transformation:

$$B_{gdv} = \sum_{u=1}^{2k+1} \left[\frac{1}{\pi} \int_0^{2\pi} B_{adu}(\theta) \cos(v\theta) d\theta \right] = \sum_{u=1}^{2k+1} B_{gduv} \quad (3-36)$$

$$B_{gquv} = \begin{cases} B_{aqu} \left[1 - \frac{p(\beta_1 + \beta_2)}{\pi} + \frac{\sin vp(\alpha_1 + \beta_1) - \sin(vp\beta_1) - \sin v(\pi - p\alpha_2)}{v\pi} \right] & (u = v) \\ \frac{2}{\pi} B_{aqu} \left[\frac{\sin \frac{(u+v)p(\alpha_1 + \beta_1)}{2} - \sin \frac{(u+v)p\beta_1}{2} - \sin \frac{(u+v)(\pi - p\alpha_2)}{2}}{u+v} \right. \\ \quad \left. + \frac{\sin \frac{(u-v)p(\alpha_1 + \beta_1)}{2} - \sin \frac{(u-v)p\beta_1}{2} - \sin \frac{(u-v)(\pi - p\alpha_2)}{2}}{u-v} \right] & (u \neq v) \end{cases} \quad (3-37)$$

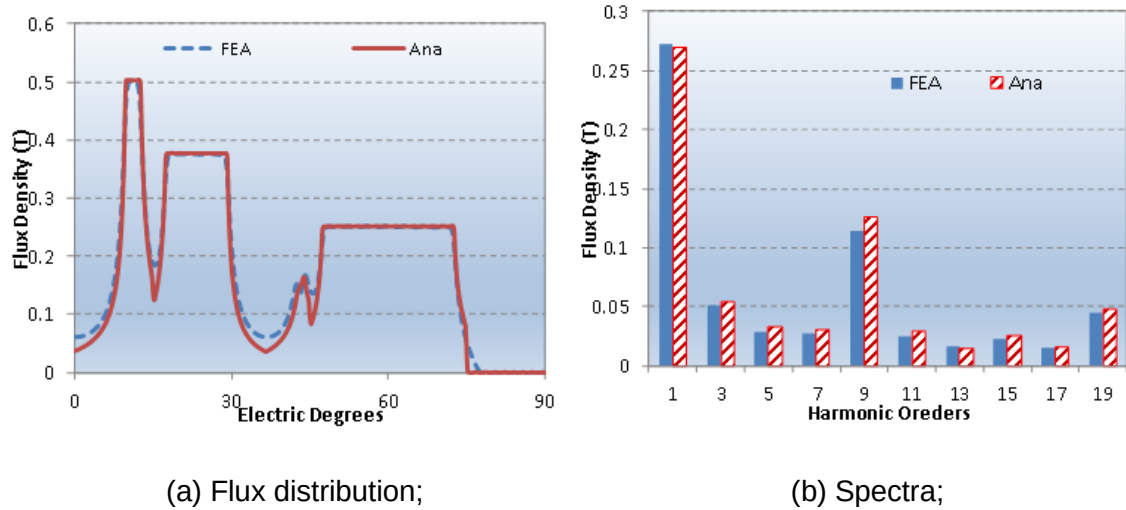


Figure 3-15 Flux distribution and spectra in the airgap excited by q-axis current.

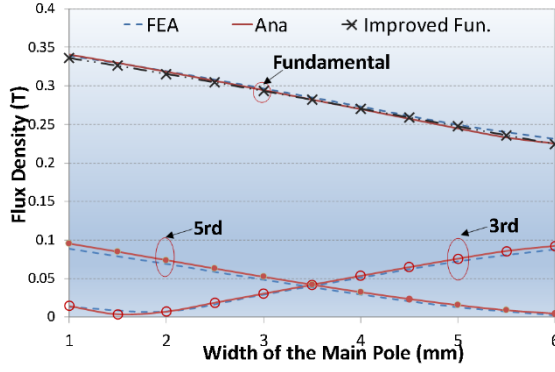
And if the slotting effect is to be considered, the flux distribution in the slot areas needs to be approximated as suggested in Figure 3-2, which would associate with a complicated calculation of sine and cosine integral functions for the main and secondary PM slot openings.

Flux distribution and spectra in the airgap excited by q-axis current for a typical two-layer model are depicted in Figure 3-15. The analytical methods can well follow the flux waveform by FEA. However, the spectra of the airgap flux distribution indicate that noticeable error on the fundamental is caused due to the ignored flux in the PM slot areas. Since the maximum value of armature MMF appears in the main PM slot area, the average flux density is still over 10% that of the maximum value in the core area and it is not negligible. Additionally, this part of flux has higher contributions to the fundamental according to the definition of Fourier transform with an electric position close to zero. As a result, the fundamental component derived from the analytical method is over 10% lower. Because the fundamental is the major source of armature flux linkage, the error would result in a considerable error for the prediction of flux linkage and inductance. On the other hand, a fair agreement can be achieved by the analytical model when the harmonic order is lower than ten. With an improved analytical model, a more accurate result can be achieved for both the fundamental and higher order harmonics at the expense of a much more complicated model.

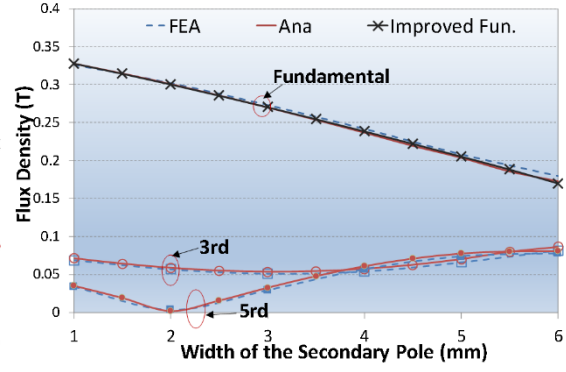
Despite the very high accuracy accomplished by the improved analytical model, the equations are too complicated and it is difficult to obtain clear relations between the machine performance and the key design parameters. In order to achieve acceptable accuracy for the analytical method, the flux through PM slot areas has to be taken into account. Since the fundamental is extremely critical for the machine performance prediction and is not accurately estimated in the analytical model, it is of great importance to find a simpler way to get an accurate estimation.

The fundamental content of the airgap flux is derived by summation of the first order decomposed from all the flux components excited by various armature MMF harmonics. The maximum flux density excited by the ν th MMF is inversely proportional to ν and would be further reduced due to short pitch windings according to Equation (3-35). By performing the Fourier transform, the decomposed fundamental from flux field excited by the ν th MMF harmonic will be further reduced to at least $1/\nu$ according to Equation (3-37). Thus, the fundamental content due to higher order stator MMF excitation is very small, and the flux distribution induced by the fundamental MMF is of sufficient accuracy to be included for the analytical method. As the flux in the PM slots areas cannot be ignored, the flux excited by fundamental MMF in these regions should also be added as compensation to the fundamental. Considering that rotor radius is much larger than the width of the main and secondary PM slots, an approximation of the fundamental flux component for the analytical method can be revised as:

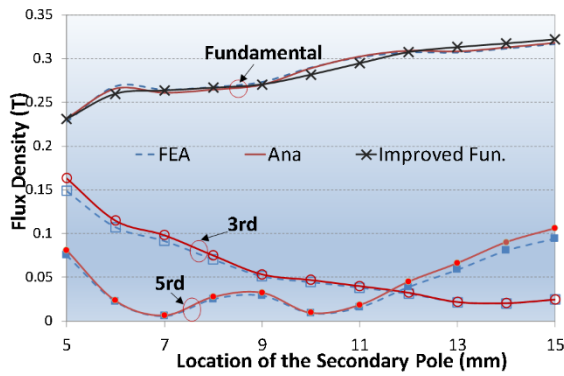
$$B_{gq1} \approx \frac{2\mu_0 m N_a k_{w1}}{\pi p g_c} i_q \left\{ \begin{aligned} & 1 - \frac{p(\beta_1 + \beta_2)}{\pi} + \frac{\sin p(\alpha_1 + \beta_1) - \sin(p\beta_1) - \sin(\pi - p\alpha_2)}{\pi} \\ & + \frac{4pg_c}{\pi^2 R_{ro}} \left[\left(1 - \frac{p^2 \beta_1^2}{4} \right) \ln \left(1 + \frac{\pi R_{ro} \beta_1}{2g_c} \right) + \ln \left(1 + \frac{\pi R_{ro} \beta_2}{4g_c} \right) \right] \\ & + \cos \left(p \frac{2\alpha_1 + 2\beta_1 + \beta_2}{2} \right) \ln \left(1 + \frac{\pi R_{ro} \beta_2}{4g_c} \right) \end{aligned} \right\} \quad (3-38)$$



(a) With different W_1 ;



(b) With different W_2 ;



(c) With different D_m ;

Figure 3-16 Fundamental, 3rd and 5th components of flux distribution excited by of q-axis armature current.

The results of the flux distribution by FEA and different analytical methods are demonstrated in Figure 3-16. With the compensation of the PM slot flux, great improvement is achieved for the fundamentals with a relatively simple model. The two-layer structure exhibits much lower harmonic contents. Since the higher order harmonics are not as a critical and good agreement is accomplished, the analytical model would be used for the prediction of the harmonic components.

As indicated in Equation (3-38), the fundamental flux decreases with the width of the main and secondary PM slots. That is because wider PM slots enlarge the slot area and reduce the equivalent airgap permeance. On the other hand, the location of the secondary poles appears to have much less influence on the fundamental. As secondary poles moving away from the main pole, the flux density field in the secondary PM slot area becomes weaker due to the lower

stator MMF, and the fundamental is less affected. However, since the airgap flux distribution waveform can be flexibly modulated by changing the location of the secondary pole, the harmonic contents are significantly affected, which should be made full use of during the machine design process to achieve required performances. The length of the main pole does not have any influence over the q-axis flux distribution in the ideal model.

For the sake of simplicity, the magnetizing inductance is calculated using the improved fundamental flux component, and the result is compared with FEA in Figure 3-17. The results show a similar, varying trend with the fundamental flux component, and sufficient accuracy is achieved. Due to the neglecting of the flux in the PM slot areas for the higher harmonics, the error would be higher with PM slot width increases. Since the harmonic content is greatly affected by the secondary PM location, there is a slightly higher variation. In all, much simpler model and fairly good accuracy can be achieved by the approximated model, and the error ratio is within 3%. The contributions of various harmonics to the magnetising inductance are demonstrated in Figure 3-18, and the higher order harmonics show an insignificant impact on the inductance due to the low harmonic contents. The fundamental flux contributes over 95% of the flux linkage and magnetising inductance, and the highest contribution from a single harmonic component is less than 3%.

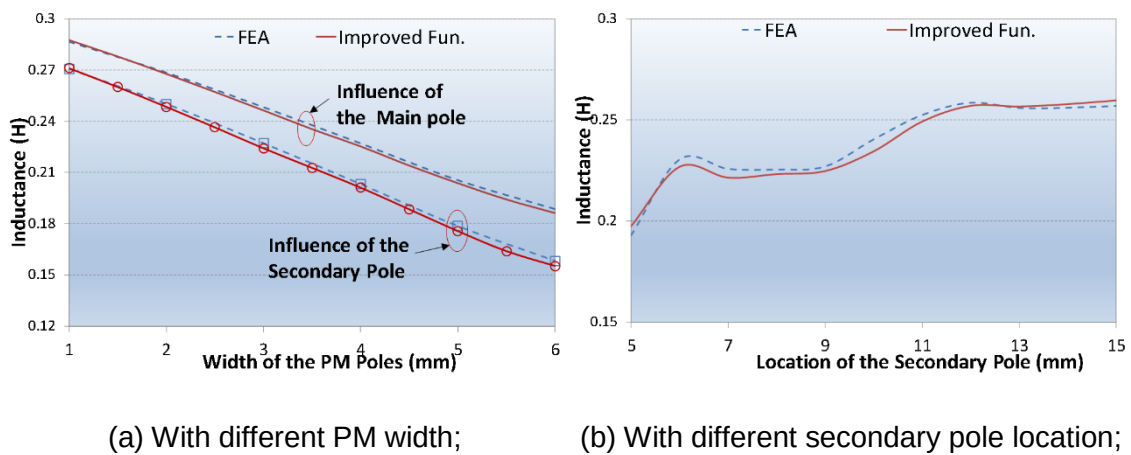


Figure 3-17 Comparison of q-axis magnetizing inductance of revised

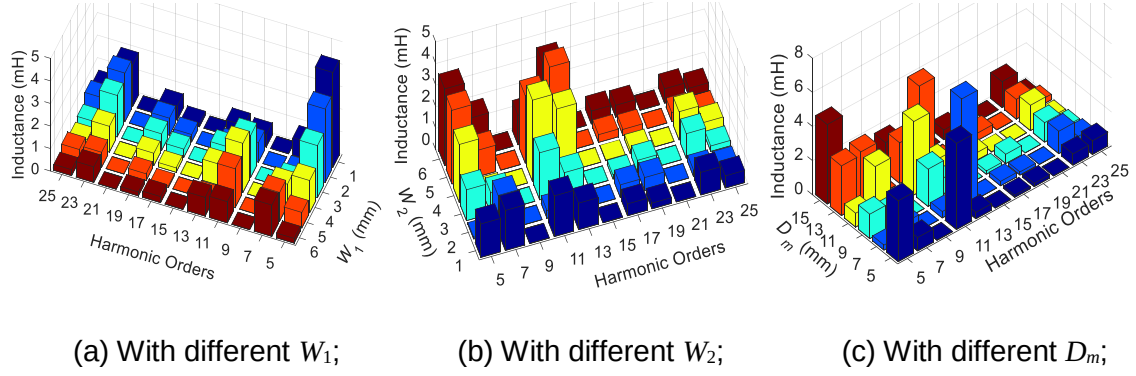


Figure 3-18 Contributions of flux harmonics to q-axis magnetizing inductance.

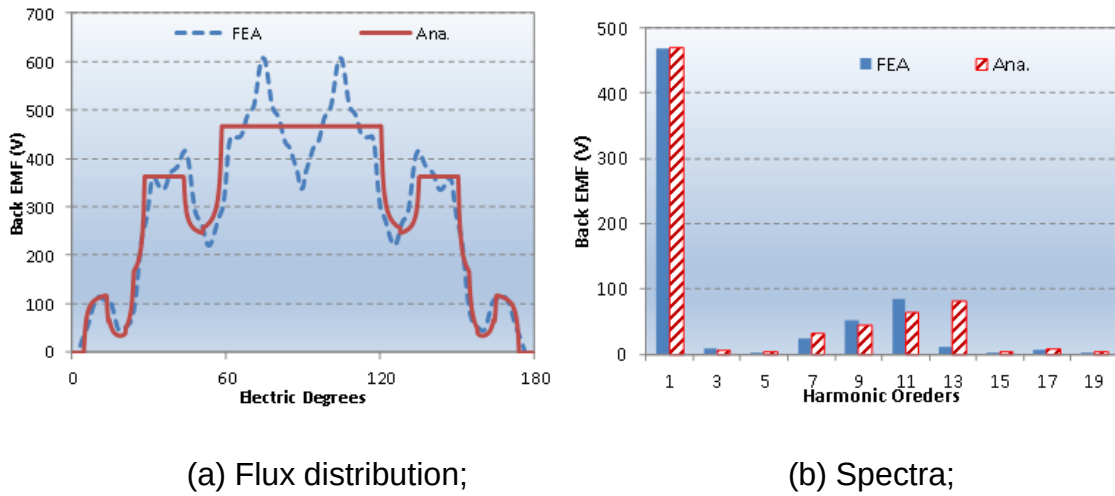


Figure 3-19 Back EMF and spectra of a typical two-layer structure.

3.2.1.4 Performance based on the analytical method

Based on the analytical models built, back electromotive force (EMF) can be obtained by

$$E_{mf}(t) = \frac{d\Psi_f}{dt} = \frac{4N_a R_{ro} l_{ef} \omega}{p} \sum_{v=2k-1} B_{fgv} k_{wv} \sin(v\omega t) \quad (3-39)$$

The back EMF waveform and spectra are derived as illustrated in Figure 3-19. Good capture of the shape of the waveform is obtained, and the spectra analysis indicates good accuracy of the developed analytical method. However, the back EMF waveform of the analytical method is not exactly the same with FEA and deviations are seen. The reason is that the back EMF obtained by FEA is calculated with a spinning rotor, while the analytical model is based on a static

model. Since the relative location of the stator and rotor slots changes when rotating, the tooth harmonics are different from those at a standstill.

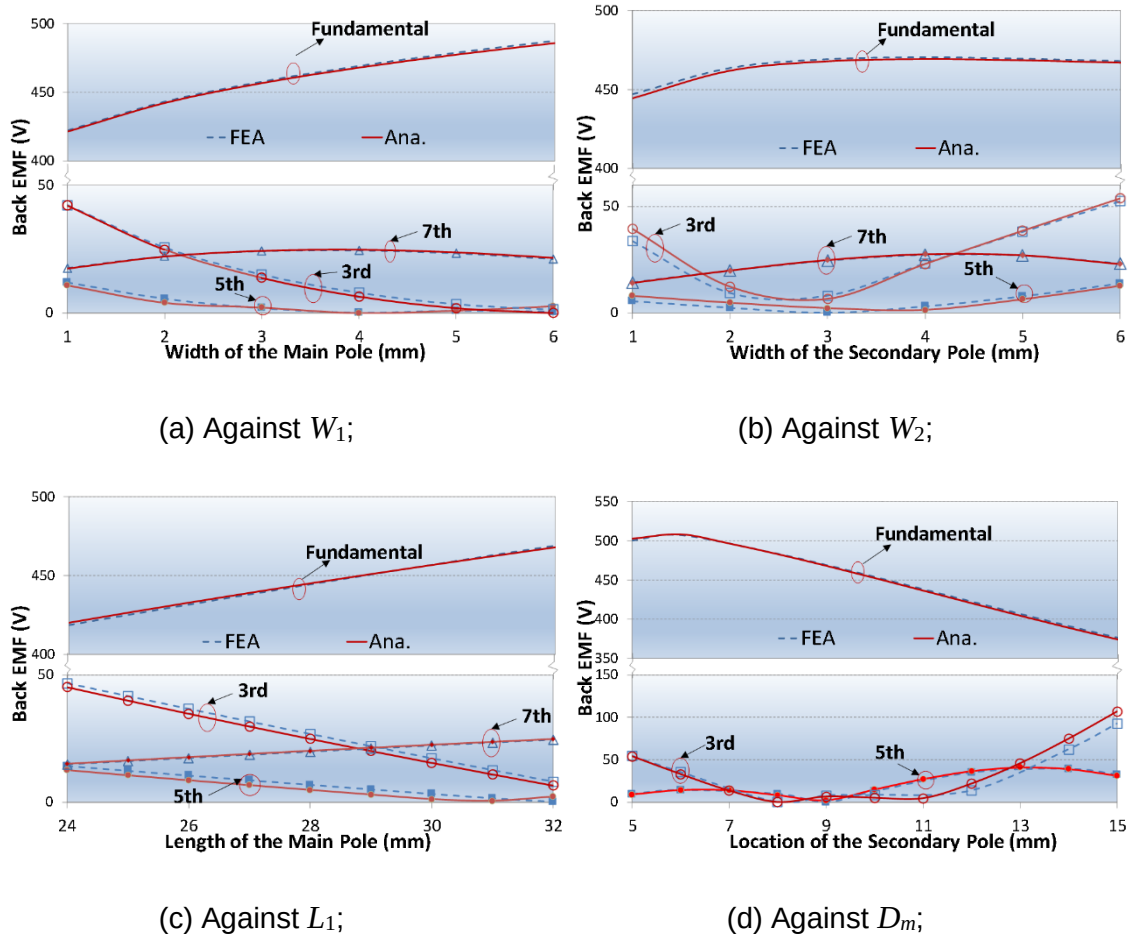
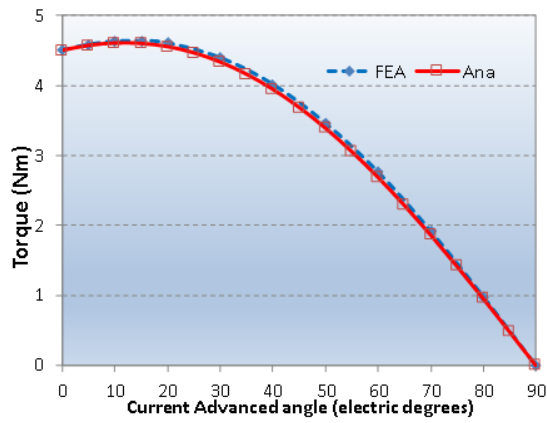
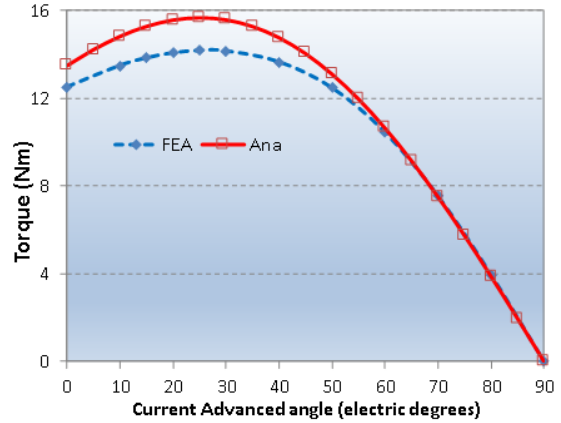


Figure 3-20 Fundamental, 3rd and 5th components of Back EMF with various rotor parameters.

Since the winding back EMF is induced by the flux field excited by the PMs, the influence of the design parameters is similar to that of open-circuit airgap flux distribution. Short-pitched windings reduce the harmonic components, which are not the tooth harmonics. Hence, the harmonic contents in the back EMF are much lower than the airgap flux distribution. As the major flux supplier, the size of the main PM has a higher effect on the fundamental of back EMF, as shown in Figure 3-20(a) and (c). Since the working point of the PM increases with the width and the flux focusing effect is enhanced with length, wider and longer main PMs can intensify the PM flux density. Since the secondary pole regulates the flux waveform, and both the fundamental and harmonics are

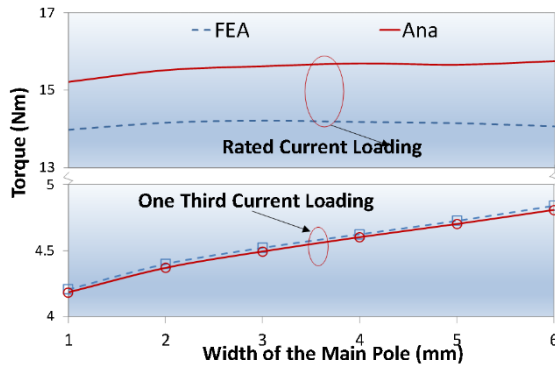


(a) Under 1/3 of rated current;

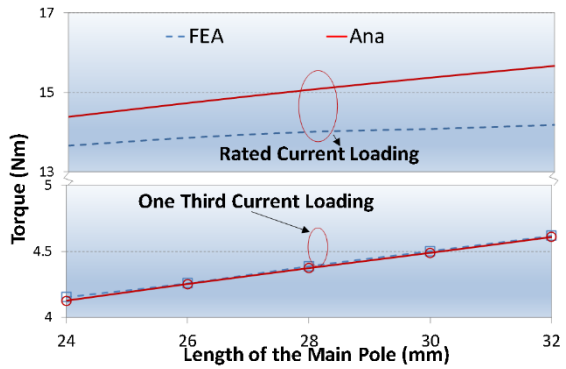


(b) Under rated current;

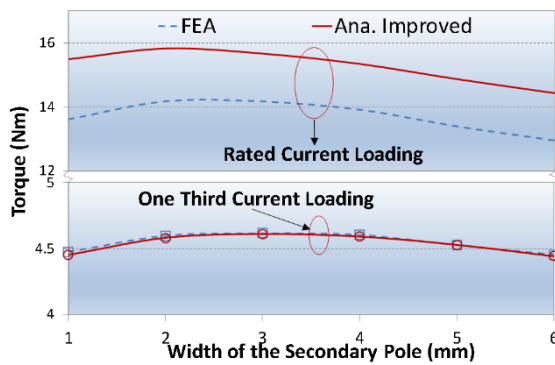
Figure 3-21 Comparison of torque from FEA and analytical models.



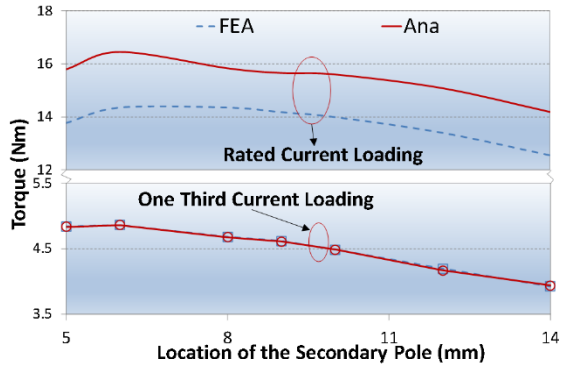
(a) Against W_1 ;



(b) Against L_1 ;



(c) Against W_2 ;



(d) Against D_m ;

Figure 3-22 Torque output with different PM parameters under light and rated load conditions.

greatly influenced. With the flexible design of the secondary pole, low harmonic contents can be achieved, which could potentially reduce loss and torque ripple.

According to [10], the total electromagnetic torque can be computed by:

$$T_{em} = \frac{mp}{4} [\Psi_f i_q + (L_d - L_q) i_d i_q] \quad (3-40)$$

where i_d and i_q is the d-axis and q-axis current respectively.

The electromagnetic torque profiles of FEA and analytical methods are compared in Figure 3-21 under light load and rated load conditions. When under light load conditions, no saturation occurs in the lamination cores, and the torque profile can be predicted accurately by the analytical methods. With flux in the PM slots ignored, the analytical model shows slightly higher torque at a low current angle because of slightly higher PM flux derived. But because the PM slots are relatively small and have a low impact on the flux distribution, very good agreement is achieved. On the other hand, the improved analytical method almost coincides with the FEA result owing to the more complicated model at light load. With the increase of machines loading, the error will increase due to the saturation. Under heavy load conditions, severe saturation appears in the machine, and error can go up to 12% for the maximum torque output shown in Figure 3-21 (b). With a higher current angle, a larger demagnetizing current is injected and saturation is alleviated. Thus, the torque predicted by analytical models coincides with FEA again. Although the error percentage increases with saturation, the trend of torque profile and the optimal current angle can be well predicted, which can be very helpful for the preliminary design.

The maximum torque against different PM parameters under 1/3 and rated current are illustrated in Figure 3-22. Under light load conditions, the PM torque consists of the majority of the overall torque and can be effectively enhanced by wider and longer main PMs. Therefore, the output torque increases with the main PM width and length. Under rated loading, reluctance torque consists of a considerable part of the overall torque, and wider main PM does not always mean higher torque because it also leads to higher saturation and decrease in

reluctance torque. Consequently, the overall torque is increased very little by wider main PMs when W_1 is relatively small. But the decrease in torque can be observed when W_1 is too large, as shown in Figure 3-22(a). Thus, the optimal composition of PM and reluctance torque components should be compromised with suitable W_1 to achieve higher torque density. Although the length of the main PM has little impact on the armature inductance, the torque increases with L_1 due to the increased PM torque. That is why longer main PMs are always preferred so long as there is enough space to accommodate the PMs.

As discussed previously, the width of the secondary pole has a very small effect on the enhancement of the PM flux field but a higher influence on the armature inductance. Proper W_2 can increase the PM torque slightly due to the higher PM working point and better rotor saliency. But as W_2 becomes wider, a significant decrease in L_{mq} occurs without noticeable enhancement in PM flux, which leads to the decrease in reluctance torque and thus the overall torque. The impact is more complicated for D_m because both the PM flux and rotor saliency can be greatly affected. The length of secondary PM poles is increased as the decrease of D_m , which results in much higher PM field and thus PM torque. As a result, an increase of maximum torque is observed when secondary poles are closer to the main poles at light current loading. Meanwhile, armature inductance reduces with the decrease of D_m , and the highest difference between d- and q-axis inductances is achieved with optimal D_m . And because of the improvement in reluctance torque, the overall torque increases D_m at first despite smaller PM torque. As the secondary pole moves further away, both PM and reluctance torques decrease, resulting in a decrease of overall torque.

3.3 DESIGN OF TWO-LAYER SPOKE-TYPE MACHINES

In this section, electromagnetic design and optimizations are carried out for the spoke-type two-layer IPM machines. The torque density is maximized based on the developed analytical models and FEA simulations. The influence of the key design parameters is comprehensively investigated to provide helpful insight into the design of two-layer IPM ferrite machines. Finally, the optimal design is chosen for the prototype machine.

3.3.1 Airgap Flux Distribution

In a two-layer machine, all the open-circuit main flux starts from the main PM poles and the size of the main poles have a direct impact on the average flux density in the airgap. On the other hand, the secondary poles regulate the flux between the core area ① and ②, and mainly affect the shape of the airgap flux distribution. According to Equation (3-22), the open-circuit fundamental flux component can be very roughly estimated by

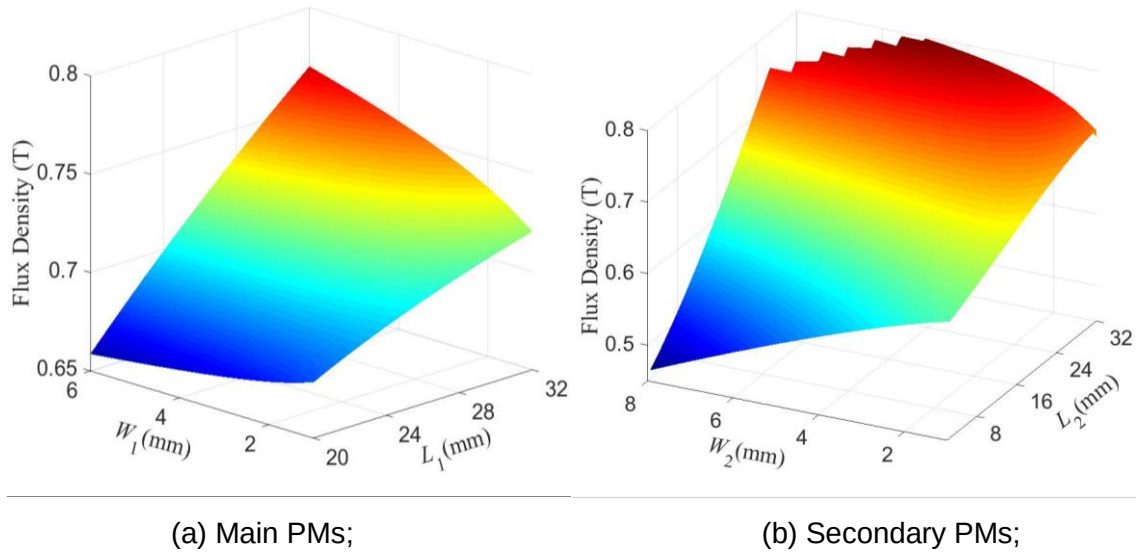


Figure 3-23 Magnitude of fundamental flux density against PM size

$$B_{fg1} \approx \frac{4B_r}{\pi} \left\{ \frac{L_1 - L_2}{R_{ro}\alpha_1} \left[\sin\left(p \frac{\alpha_1 + \alpha_2 + \beta_2}{2}\right) - \sin\left(p \frac{\alpha_2 + \beta_2}{2}\right) \right] + \frac{L_2}{R_{ro}\alpha_2} \cdot \sin\left(\frac{p\alpha_2}{2}\right) \right\} \quad (3-41)$$

Figure 3-23 depicts the influence of the size of main and secondary PM poles on fundamental flux density in the airgap. The result indicates that the length of the PM poles has a higher impact on the flux density. By contrast, the width of PMs shows smaller influence because the PM working point and the flux focusing ratio can be slightly improved. The length of PMs can boost the flux focusing effect and similar to the one-layer structure, the longer they are, the higher the flux density will be. For the main pole, longer PMs are preferred to maximise the flux focusing effect since L_1 is proportional to the overall PM flux. The length of the secondary poles can also increase the fundamental flux since more PM is used and the flux density in the area ② is intensified. It should be noted that the length

of secondary PMs increases as they move towards to the main pole. And when the maximum length is reached, the model becomes a conventional spoke-type machine. In other words, higher open-circuit flux density can be achieved by the spoke-type configuration. Nevertheless, the waveform of airgap flux distribution can be improved by the secondary PM layers, and more sinusoidal flux distribution can be achieved with lower harmonic contents. Also, more importantly, the rotor salient ratio is expected to be improved by the proper design of the second PM layers.

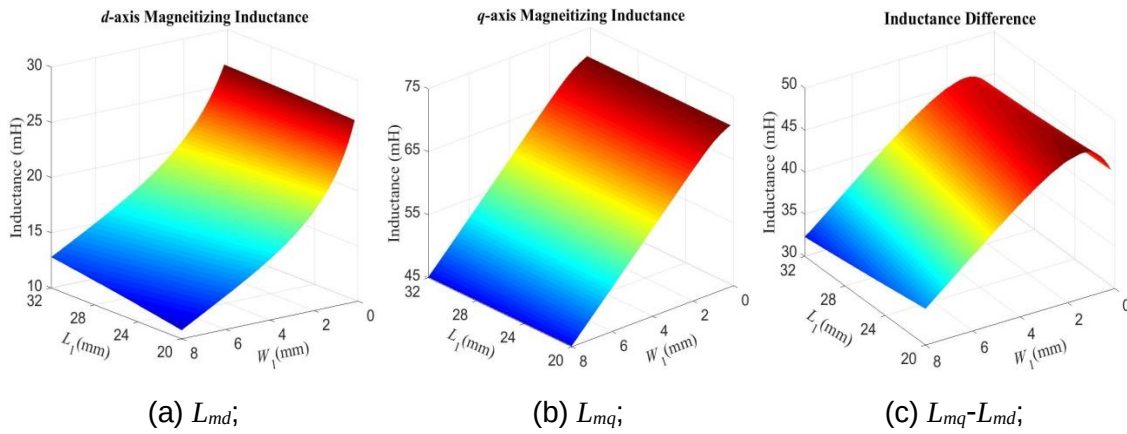


Figure 3-24 Influence of the main PM size on magnetizing inductance

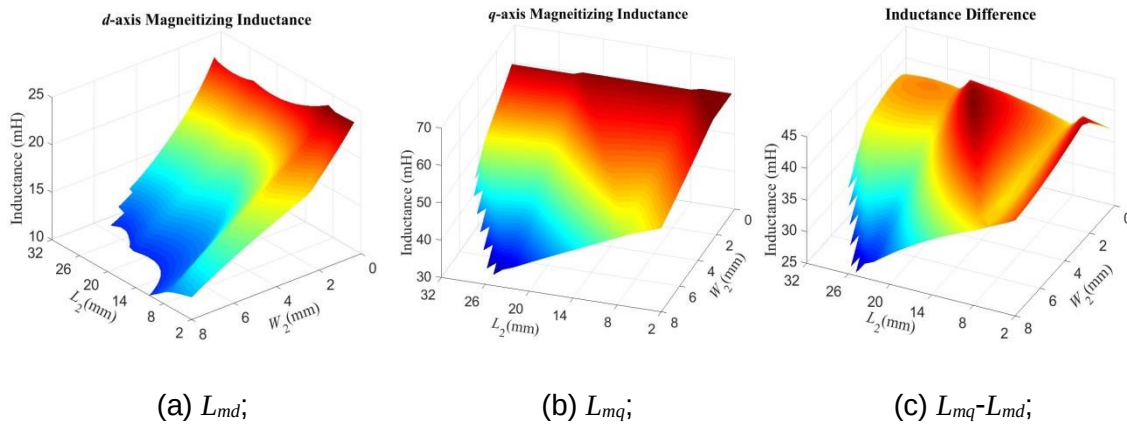


Figure 3-25 Influence of the secondary PMs on magnetizing inductance.

3.3.2 Magnetizing Inductance

The d- and q-axis inductance together with their inductance difference against the width and length of the main PM are demonstrated in Figure 3-24. Since the main PM poles function similarly to the spoke-type poles of the one-layer

structure, the influences are more or less the same. Both the d - and q -axis inductances decrease dramatically with the main PM width as the reluctance of the equivalent magnetic circuit is increased, while the PM length poses no noticeable impact. When the main PM width is small, L_{md} drops faster than L_{mq} and rotor saliency is enhanced. But both L_{md} and L_{mq} drop severely with large PM width, resulting in deterioration in saliency.

As a unique design of the multi-layer configuration, the secondary PM layers play a very important role in regulating the flux between the two core areas. Since the magnetic paths of flux can be altered by the secondary layers, both the d - and q -axis inductances are greatly affected. As shown in Figure 3-25, the influences imposed by the secondary layers are different on the d - and q -axis inductances. Hence, with an elaborate design of the secondary poles, the rotor saliency can be improved to boost the reluctance torque.

The impact of the secondary PM width is easy to understand. Because the PM slot area is increased, both L_{md} and L_{mq} decrease with wider PM slots. Besides, since part of the d -axis flux travels through the PM slot, L_{md} is more seriously influenced by the PM width. On the other hand, the influence of the location (length) of secondary PMs is more complicated. As shown in Figure 3-26, profound changes occur in the machine flux distribution due to the variation of the location of secondary PM layers. The maximum length for the secondary PMs is obtained when they touch the main PM, and the model becomes a conventional spoke-type configuration. The main and secondary slots merge together and only one flux loop for the d -axis reactive field is formed. All the flux has to cross the large main PM slot, and the reluctance of the magnetic circuit reaches its maximum. As the secondary poles move away from the main pole, Loop 2 emerges and part of the reactive flux goes through only the main PM slot through core area ①. As a result, the overall reluctance is reduced, and the reactive field and L_{md} increase accordingly. With the secondary poles moving further away, more flux goes by Loop 2 due to the much lower reluctance and L_{md} rises dramatically. When the secondary poles move too far away from the main PM,

the flux in Loop1 is negligible, and the maximum value of L_{md} is reached, as shown in Figure 3-26(c).

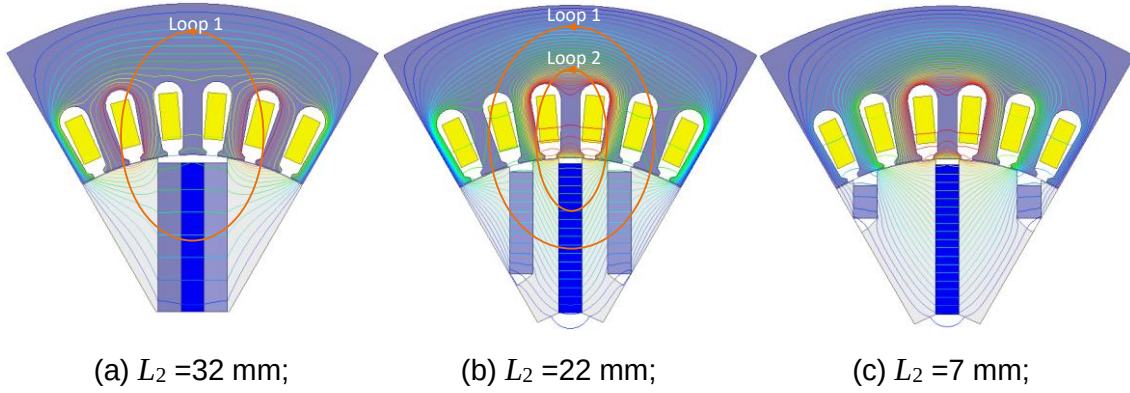


Figure 3-26 D-axis reactive field distribution and loops with different L_2 .

As for the q-axis reactive field, since the flux does not go through the PM slots, the PM slots only cause the change of equivalent airgap length in the slot area and thus less effect is expected. By ignoring the flux in the PM slot area and higher order space harmonics, roughly estimation of L_{mq} can be obtained with further simplification from Equation (3-38):

$$L_{mq} \approx \frac{8\mu_0 m N_a^2 R_{ro} l_{ef} k_{w1}^2}{\pi p^2 g_c} \left[1 - \frac{p(\beta_1 + \beta_2)}{\pi} - \frac{\sin(p\beta_1)}{\pi} - \frac{2\sin p(\beta_2/2)\cos p(\alpha_1 + \beta_1 + \beta_2/2)}{\pi} \right] \quad (3-42)$$

If the size of main PMs and the width of secondary PMs are fixed, the radian of area ① (α_1) becomes the only variable in Equation (3-42) and the effect of the location of secondary PMs can be revealed. As they move away from the main PM, L_{mq} rises with the increase of α_1 . However, due to the impact of the spatial harmonics from the armature MMF and large PM slots, the value of L_{mq} fluctuates while going up, as illustrated in Figure 3-25 (b). By taking advantage of the different influences on the d- and q-axis flux distribution, the rotor saliency can be further enhanced by the secondary layers, as depicted in Figure 3-25(c).

3.3.3 Torque

As indicated in Equation (3-40), the total electromagnetic torque in PMSMs can be segregated into two parts, which are the PM torque T_{pm} and the reluctance torque T_r . And they can be calculated by

$$\begin{cases} T_{pm} = \frac{mp}{4} \Psi_f i_q \\ T_r = \frac{mp}{4} (L_d - L_q) i_d i_q \end{cases} \quad (3-43)$$

The phasor diagram of a PMSM is demonstrated in **Error! Reference source not found.**, and according to the relations of the current and voltage vectors, the two torque components, as well as the total torque, can be written as:

$$\begin{cases} T_{em} = T_{pmm} \cos \phi + T_{rm} \sin 2\phi \\ T_{pmm} = \frac{mp}{4} \Psi_f I_m \\ T_{rm} = \frac{mp}{8} (L_q - L_d) I_m^2 \end{cases} \quad (3-44)$$

Where I_m is the amplitude of armature current, ϕ is the electric angle of back EMF vector and armature current, or the advanced current angle, T_{pm} and T_r the PM and reluctance torque component, T_{pmm} and T_{rm} are the maximum PM torque and reluctance torque that can be achieved for a PMSM. The parameters of the PM poles have quite different impacts on the open-circuit flux distribution and rotor saliency. Thus, the optimal design process is necessary to achieve better performance. The two-layer structure is more complicated, and more sizing parameters are related to optimisation. As the parameters are all interdependent due to geometrical restrains, the variables should be carefully chosen to make the optimisation easier and more understandable. The length of the main PMs is set to the maximum value, as it can boost the open-circuit PM flux field effectively while having little impact on rotor saliency.

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Practically, people are more concerned about the amount of magnet material consumed. With the amount of PM fixed, only two variables can determine the rotor geometry. For a more intuitional view of how the machine performance is affected by the rotor design, the width of the main PM W_1 and location of the secondary pole D_m are chosen as the major parameters for the investigation.

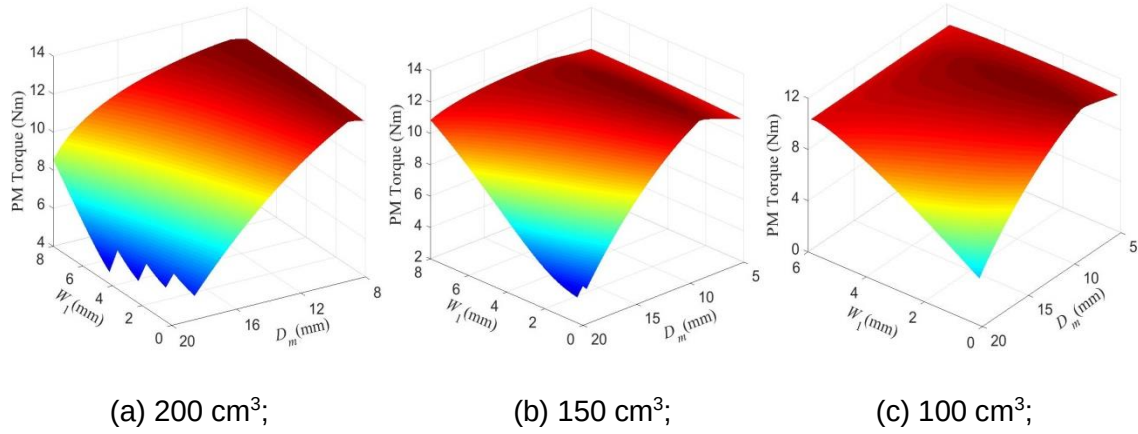


Figure 3-27 Maximum PM torque against W_1 and D_m at different amount of ferrite materials.

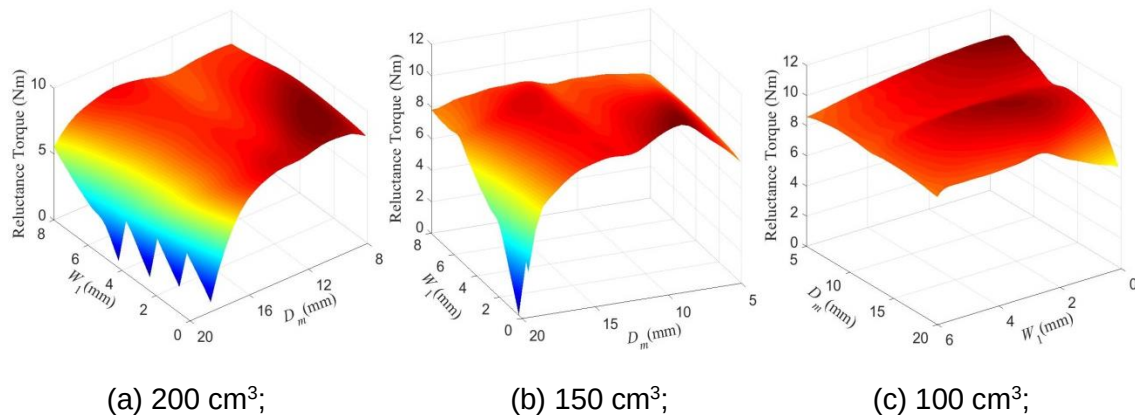


Figure 3-28 Maximum reluctance torque against W_1 and D_m at different amount of ferrite materials.

With the amount of magnet material fixed, maximum PM torque is achieved when the distance between the main and secondary PM poles is minimum, as depicted in Figure 3-27. In other words, the one-layer configurations offer better flux focusing effect and thus higher PM torque, as expected. And with a larger amount of PMs, higher PM torque can be generated, though the increment becomes smaller and smaller. On the other hand, the maximum reluctance torque

decreases with the PM amount as shown in Figure 3-28. That's because the larger PM slots occupy more space with less lamination core left, and both the d- and q-axis armature fields are weakened. Actually, only a very low amount of PMs is required to achieve the maximum saliency. It is also noteworthy that the reluctance torque tends to increase with thinner main PM slots, which indicates V-shaped structures have a higher salient ratio than the two-layer configurations.

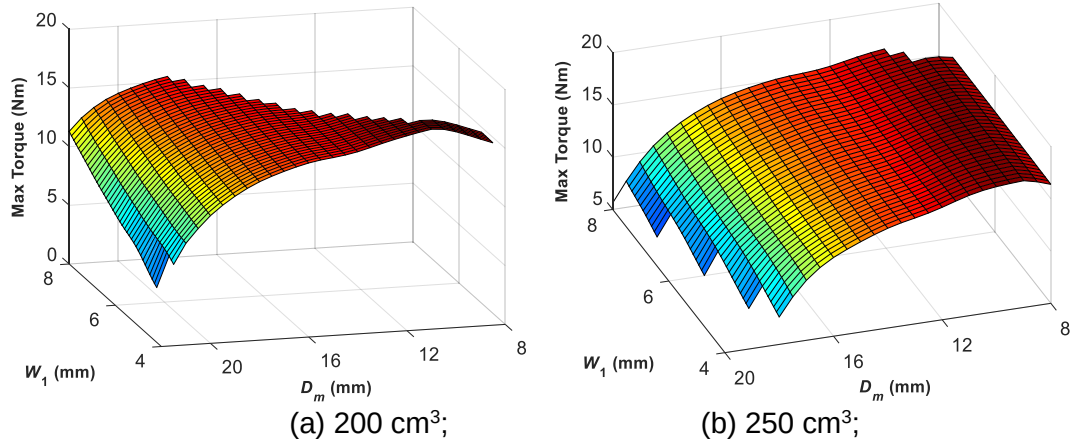


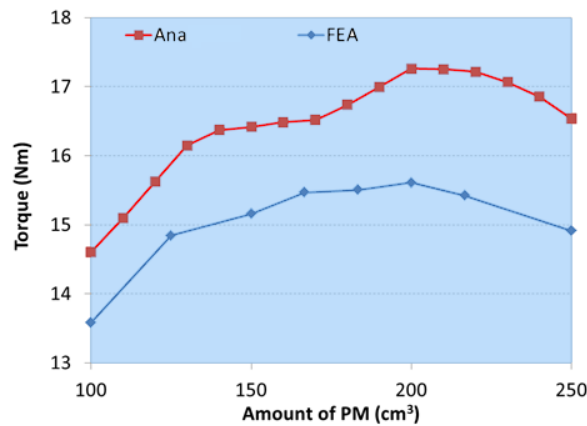
Figure 3-29 Maximum torque against W_1 and D_m at different amount of ferrite.

To achieve an optimal torque density, a compromise is necessary between the two torque components. Figure 3-29 depicts the maximum overall torque that can be delivered by different machine models. Owing to the fragile nature of ferrite PM, the designs with PM poles thinner than 4mm are considered as invalid. The results indicate that the width of PMs has a much lower impact on the torque output, and the torque capability is mainly influenced by the location of the secondary poles when the total amount of magnets is fixed. The maximum value

increases slowly with D_m at first because of increased rotor salient ratio,

Figure 3-30 Maximum torque achieved against the amounts of ferrite.

and then decreases dramatically as the secondary poles moving further away because of both PM and reluctance torque drop significantly. Since the amount of PM material has inverse effects on the PM and reluctance torque components, a larger amount of magnet does not always mean higher overall torque. More PM can increase PM torque, but it may also deteriorate the rotor saliency. There



would be a trade-off between them to achieve maximum value. The overall torque capability against PM volume is illustrated in Figure 3-30. The analytical method shows that the torque increases fast when the PM volume is low because a higher increment in the PM torque is obtained. With the increase of PM volume, it becomes less effective to enhance PM torque by using more magnets, and the negative effect on rotor saliency worsens more seriously. Thus, the torque declines after reaching the maximum value, and the maximum point is the goal of optimisation. FEA simulations verify the influence of PM volume and the variation trend is well predicted by the analytical method despite the difference caused by saturation.

The number of poles has a huge impact on the performance of the two-layer configurations. The flux focusing effect can also be enhanced by a larger number of poles. Meanwhile, the winding inductances decrease greatly causing severely degraded reluctance torque. As reluctance torque plays an important role in the overall torque, the two-layer configuration is not very suitable for machines with

a very high number of poles. With a very low number of poles, the advantage of the flux focusing effect would also disappear, so, the poles of the machine will be between 4-8 poles. To maximise the torque density, compromises should be made between the two torque components. Without considering the saturation, the result from analytical methods implies that the dramatic decrease in reluctance torque diminishes the overall torque as the number of poles increases. But higher saturation occurs in the stator yoke with lower pole number, resulting in much lower actual torque than the ideal models, as shown in Table 3-2. Although the FEA result indicates a slightly higher torque for the 8-pole model, the core loss is increased by 33% compared to the 6-pole design because operating frequency is proportional to the number of poles at the same rotating speed, iron loss increases exponentially.

Table 3-2 Maximum torque with different number of poles

Number of Poles	4	6	8
Max. Torque by Analytical (Nm)	18.70	17.26	16.74
Max. Torque by FEA (Nm)	14.31	15.57	15.63
Core Loss by FEA (W)	36.27	43.32	57.82

3.3.4 Preliminary Design for Two-Layer Structure

As a benchmarking IM is used for comparison, the stator and housing adopts exactly the same from the commercial IM. Therefore, stator dimensions as well as airgap length and rotor outer diameter are fixed. Optimisation is only performance on the rotor structure, and this project mainly focused on the two-layer design of the rotor structure.

To confirm the performance of the analytical design, FEA optimisations are also carried out for final verification. For the ease of manufacture, all the PMs are set to have the same width. Due to the influence of the saturation, both the actual PM and reluctance torques will be lower than the analytical model.

Table 3-3 Optimal parameters for the two-layer model by analytical and FEA methods

Parameters	Ana.	FEA	Unit
Number of Poles (p)	6	6	
Length of Main PMs (L_1)	31	31	mm
Width of Main PMs (W_1)	4.5	5	mm
Length of Secondary PMs (L_2)	25	21.5	mm
Width of Secondary PMs (W_2)	4.5	5	mm
Location of Secondary Pole (D_m)	11.5	13	mm
Volume of Ferrite	197	200	cm ³

According to Equation (3-44), maximum PM torque is achieved at a zero current angle, while the reluctance reaches the peak at 45 electric degrees. The maximum electromagnetic torque is obtained between 0-45 degrees, and negative d-axis current will be injected to the windings, which will also provide a flux weakening field and alleviate the saturation. Hence in practical, PM torque is more affected by saturation, and the optimal design will have a higher reluctance torque than the analytical design. The optimal parameters achieved by the analytical and FEA methods are listed in Table 3-3. FEA optimization with the consideration of spatial flux harmonics is run as a first validation stage to the , analytical model It can be seen that the analytical method is able to achieve a very close design with the FEA, though saturation is not considered. This confirms that the analytical method can locate the optimal design solution region, and then FEA modelling is required for verification and minor adjustment to the analytical design to achieve optimal design. This would significantly reduce the time required for doing optimisation.

3.4 Machine Performance

3.4.1 Open-Circuit Condition

The PM flux distribution, armature winding inductance and cogging torque have a great influence on the performances of the machines. Although all these parameters are strongly affected by the loading conditions, preliminary prediction can still be made based on the open-circuit results. The PM torque component is

largely decided by the airgap flux density and back EMF. The harmonics in flux distribution and back EMF would contribute to the torque ripple.

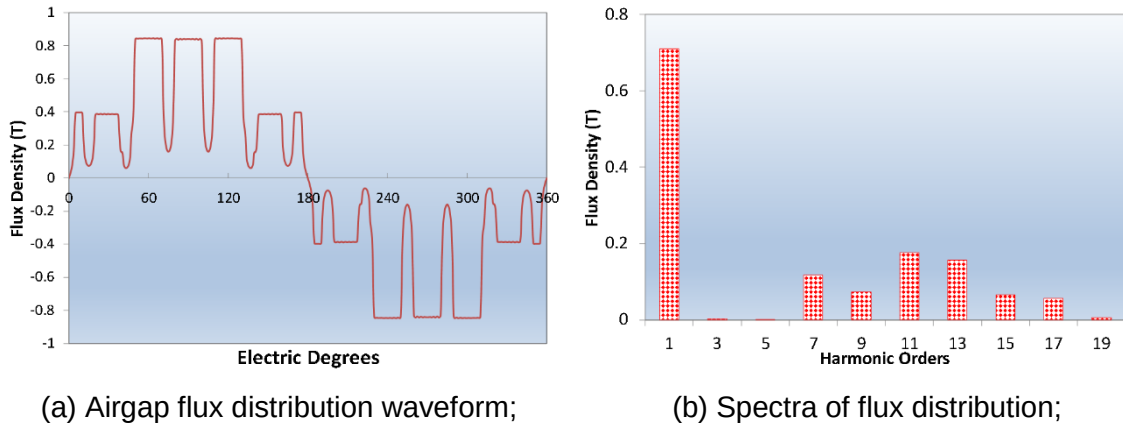


Figure 3-31 Open-circuit airgap flux distribution waveform and spectra.

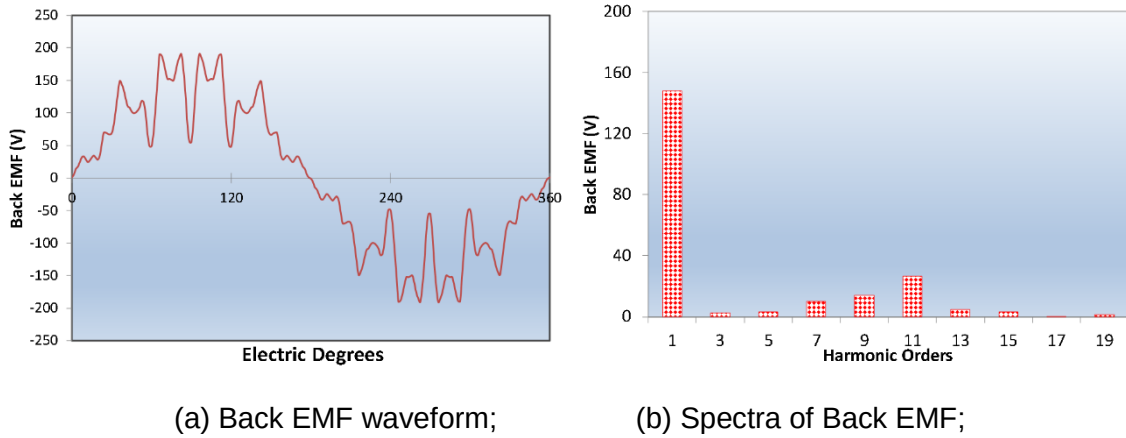


Figure 3-32 Open-circuit back EMF.

Magnetostatics FEA models are first built and the airgap flux distribution is derived in Figure 3-31. The airgap flux distribution waveform is improved because of the flux regulating the effect of the secondary layer. The third and fifth order harmonics are almost eliminated by using the two-layer configuration, and more sinusoidal waveform is obtained. Figure 3 32 depicts the corresponding back EMF waveforms and spectra.

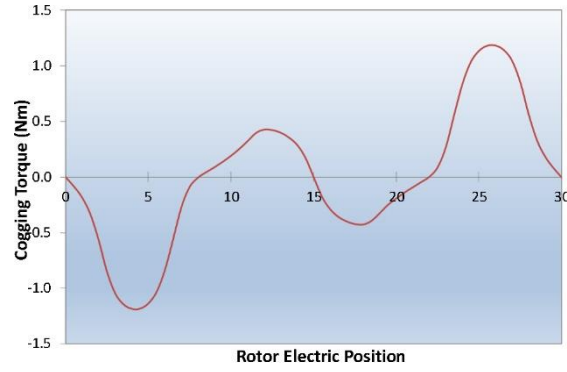
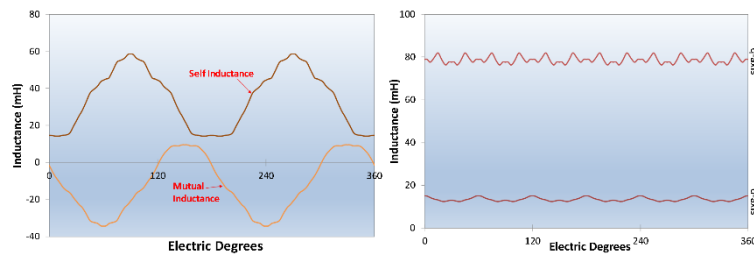


Figure 3-33 Cogging torque.

Similar to the airgap flux density, the machine exhibits low harmonic contents, which implies small cogging torque and core loss. The inductance of armature windings indicates the rotor saliency for PMSMs, which plays an important role, especially for ferrite machines as reluctance torque is crucial to the overall torque output. The armature self and mutual inductances under open-circuit conditions are depicted in Figure 3-34(a) and (b), respectively. The inductance waveforms exhibit large variations with rotor position. In order to manifest the impacts of those differences, the d- and q-axis inductances are derived accordingly by Park Transformation based on self and mutual inductances, as illustrated in Figure 3-34(c). The average values of d- and q-axis inductance show marginally high value and much high reluctance torque can be expected. It is also noteworthy that the low ripple of L_d and L_q waveforms is observed, which reflects low torque ripple.



(a) Self-inductance; ; (b) dq-axis inductance ;

Figure 3-34 Armature self and mutual inductance.

3.4.2 On-Load Conditions

Under various loading conditions, the induced voltage, inductance and even the cogging torque could change dramatically due to the armature reaction and

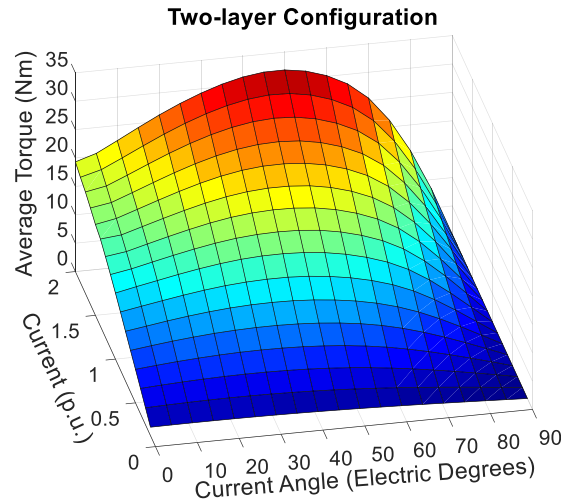


Figure 3-35 Average torque characteristics.

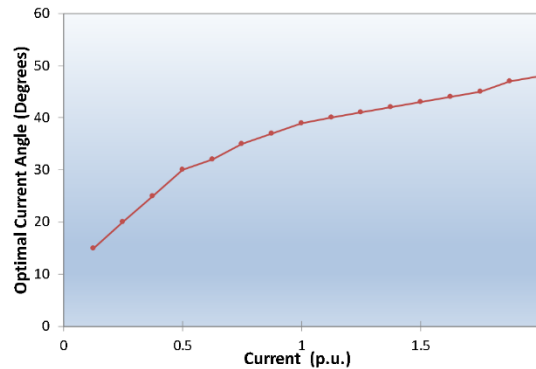


Figure 3-36 Optimal current angle.

saturation. The machines under study will become nonlinear and the two main torque components will be heavily coupled with each other. Firstly, the overall electromagnetic torque performances are investigated. The torque profiles are analysed under various current and advanced phase angle using FEA 2D nonlinear transient models. The armature current angle is the armature phase current leading the corresponding phase back EMF, and flux weakening can be achieved with positive current angles.

The average electromagnetic torque characteristics under different armature currents and advanced angle driving conditions are illustrated in Figure 3-35. The

results reveal that the average torque increases fast, which implies higher reluctance torque component. Thus, the optimal current angle to achieve the highest torque is long owing to the greater contribution from reluctance torque. The salient ratio and reluctance torque and larger optimal current angle are observed, as shown in Figure 3-36. According to Equation (3-44), the PM torque is proportional to current, and reluctance torque is proportional to the current square. So with the increase of armature current, the reluctance torque will grow more quickly than PM torque. Consequently, the contribution of the two torque components could be quite different under different loading conditions. To further investigate the torque productions, PM torque and reluctance torque is segregated using frozen permeability techniques.

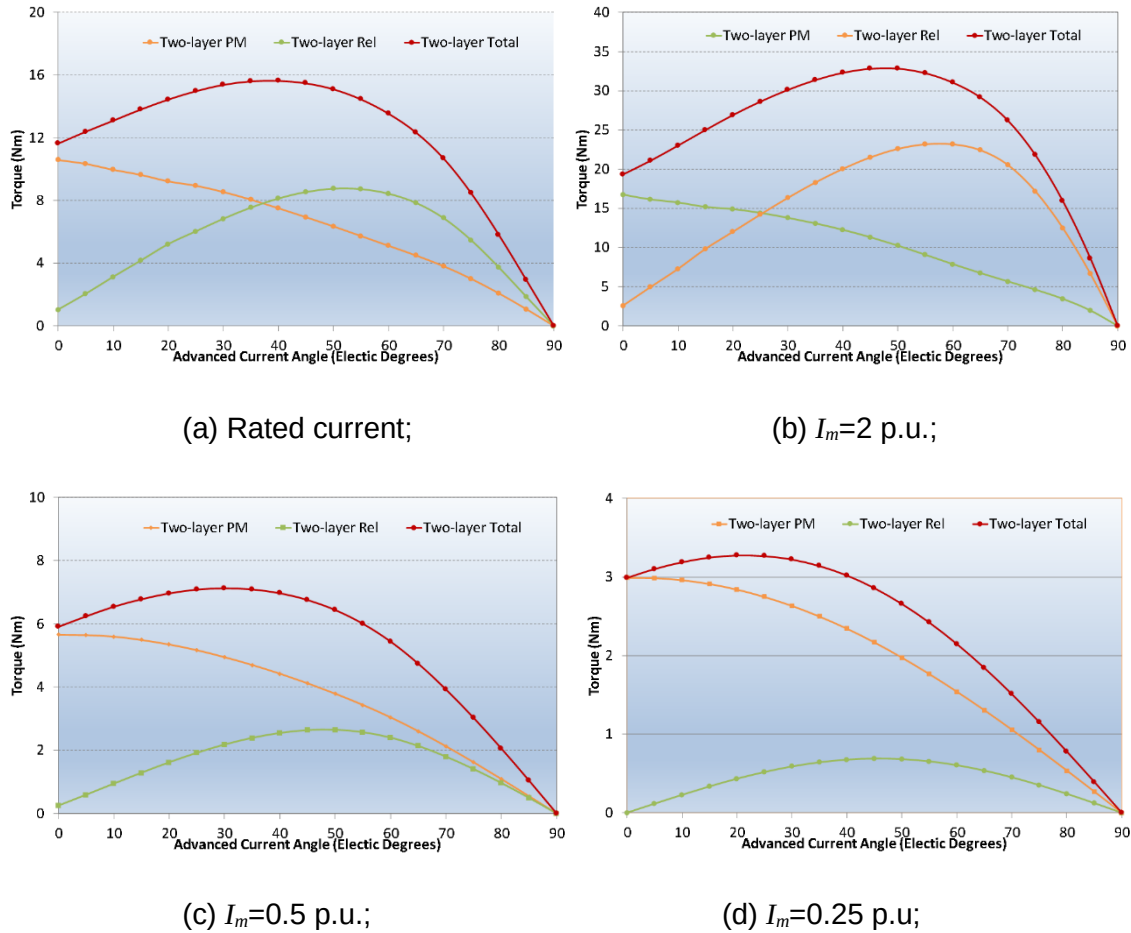


Figure 3-37 Torque components under different loading current.

The torque components of the machine under four typical loading conditions are derived and demonstrated in Figure 3-37. At the rated current excitation, as

shown in Figure 3-37(a). Reluctance torque contributes to a larger part of the overall torque output.

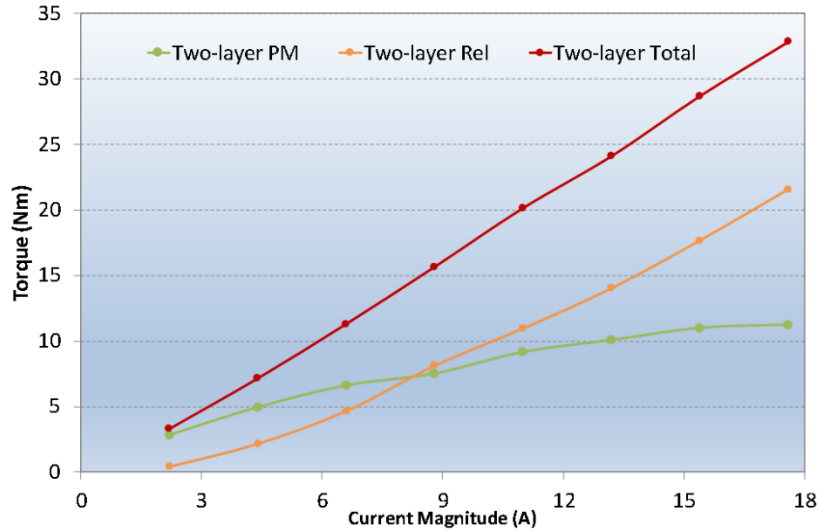
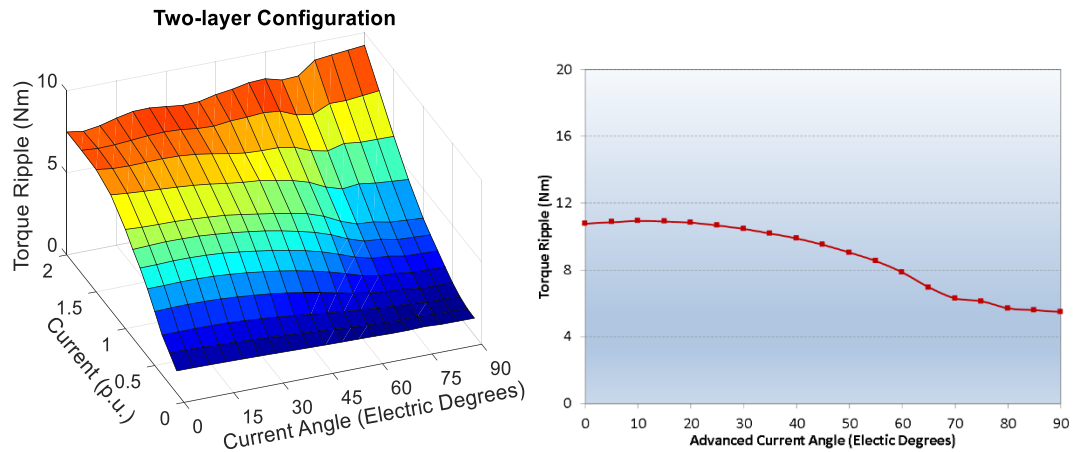


Figure 3-38 Torque segregations under maximum torque per ampere operations.

Since the reluctance torque increases more significantly than the PM component, it plays an even more important role in the overall torque output with armature current increased to overloading. On the other hand, the increment of PM torque will be further reduced by larger demagnetising d-axis current as a result of higher current angle. The machine shows good torque characteristics due to its high rotor saliency even at a highly saturated situation, as illustrated in Figure 3-37(b). Under the double rated current, it exhibits high torque density.

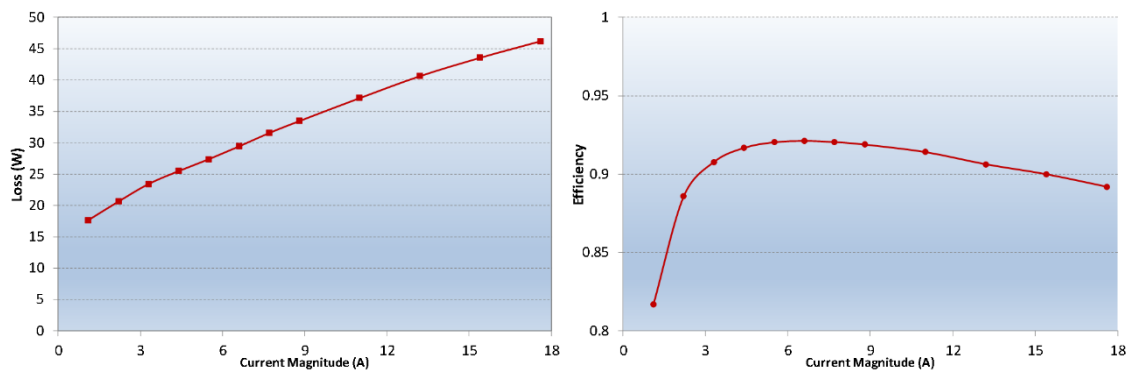
However, the situation reverses under medium and light load conditions. With the current reduced to half of the rated, reluctance torque of the machines declines dramatically. The overall torque characteristics under maximum-torque-per-ampere operations are summarized in Figure 3-38.

The torque ripple under different exciting current is depicted in Figure 3-39. The torque ripple is small due to the more sinusoidal flux distribution and well-designed secondary PM layer. But the ripple grows dramatically with the increase of current amplitude and advanced angle.



(a) Torque ripple under different currents and current angles; (b) Torque ripple at rated current

Figure 3-39 Peak-peak torque ripple.



(a) Core loss;

(b) Efficiency;

Figure 3-40 Core loss against current magnitude at rated speed.

As ferrite magnet materials are not conductive to electricity, the eddy current loss can be completely eliminated from the PMs. Figure 3-40(a) illustrates the core loss of the design under various current loading at rated speed. The core loss increases with a current magnitude due to the influence of the stronger winding reactive field. The loss in the two-layer cores increases much fast with the current as a result of much higher winding inductances and stronger armature reaction, Owing high open-circuit harmonic flux contents.

The two-layer machine shows the good average efficiency within rated operating range and achieves the efficiency of 92% at the rated operating point, striking the

right balance between torque capability and loss, as demonstrated in Figure 3-40(b). Figure 3-41 shows an example of ansys Maxwell FEA Parameters

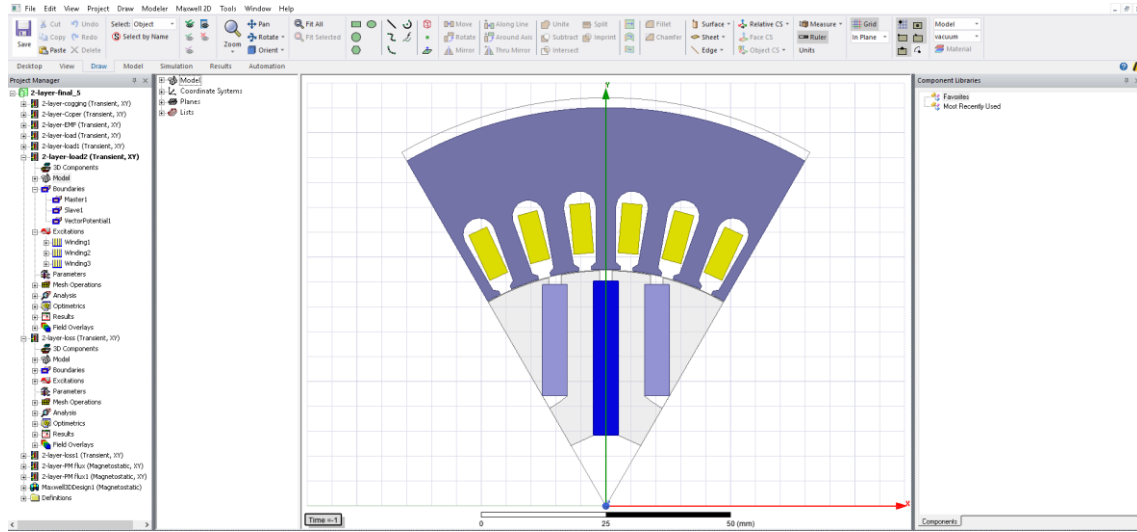


Figure 3-41 Example of the FEA parameters.

3.4.3 Performance at Higher Working Temperature

Due to the power losses in electric machines such as copper loss and iron loss, the considerable temperature rise is inevitable under continuous heavy load conditions. The magnetic properties of PMs are sensitive to the temperature and the residual flux density usually decreases with the temperature rise. For ferrite magnets, the temperature coefficient is -0.2% for residual flux density. As a result, PM machines could suffer a noticeable decrease in torque output owing to lower PM torque. According to [11], the residual flux density at certain temperature is given by

$$B_{rT} = B_{r20} [1 + k_{br} (T_c - 20)] \quad (3-45)$$

Where B_{rT} and B_{r20} are the residual flux density under T_c and 20°C, respectively, k_{br} is the temperature coefficient. According to Equation (3-45), the residual flux density is 11% lower under a working temperature of 75°C, which will definitely impose considerable effect on the PM torque and thus the torque profiles for PM machines.

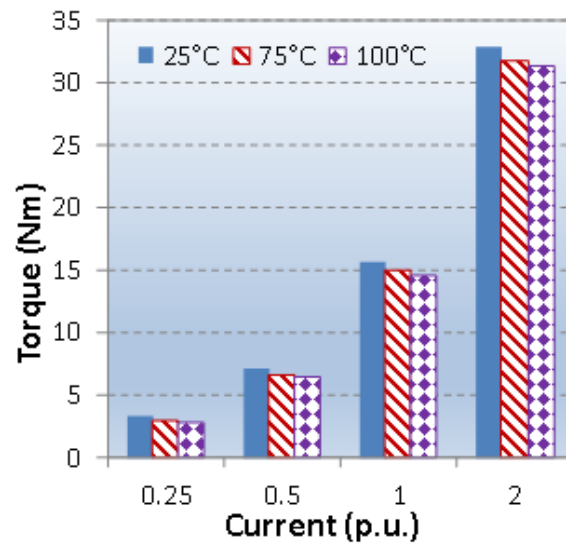


Figure 3-42 Influence of temperature on torque under different load conditions.

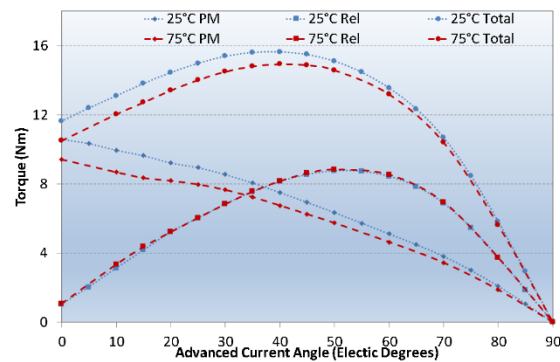


Figure 3-43 Influence of temperature on torque components at rated load.

The influence of temperature on torque production under different load is demonstrated in Figure 3-42. It is obvious that higher temperature deteriorates torque performance more severely because of lower PM flux field. Since the torque components changes with current amplitude, the influence of

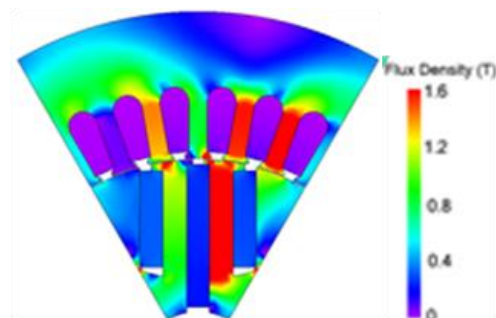


Figure 3-44 Flux distribution at rated current with optimal current angle.

The temperature rise on torque output varies under different load conditions. At light loading, torque output is more severely affected, and the torque decrease is about 10% at 75°C and 14% at 100°C under 1/4 rated current. As the increase of loading current, the influence of temperature decreases. Under double rated current loading at 75°C, the torque decrement 3.0%. Because PM torque contributes to the major part of the overall torque at light load, the overall torque will be more severely affected by the temperature rise. However, the increase in temperature has little influence on the reluctance component, and the overall torque is less influenced when the reluctance torque component dominates under high current loading.

Figure 3-43 depicts the torque profiles under 25°C and 75°C with rated current and indicates more detailed explanations for the different effects on the performances of the machine. The machine shows a good performance against temperature rise, and the torque decrease at 75°C is 4.5%, which makes the rated torque output higher. The temperature rise only affects the PM torque, while their reluctance torque profiles are almost the same under different temperature. As shown in Figure 3-44, heavy saturation appears in the rotor core between different PM layers due to the q-axis reactive field. Thus, the highly saturated area in the rotor cores is caused by reactive armature field.

3.5 Conclusion

The analytical methods for two-layer spoke type IPM machines are developed based on the equivalent magnetic circuit theory. Moreover, FEA models are built to verify the analytical method. Due to the large PM slots in the rotor, a sudden variation of airgap reluctance occurs in the PM slot areas and causes serious distortion for both PM and armature reactive field. For the sake of simplicity, infinite reluctance can be assumed for the analytical model. But slightly higher error percentage is observed comparing to FEA results. With the approximation of the flux distribution in the PM slot area, higher accuracy is obtained with a more complicated model by the improved analytical methods. To obtain simple and accurate relations between the machine performance and design parameters, the fundamental compensation solutions are also developed by using the analytical

method with simplified PM slot flux compensation for the fundamental components, and a compromise between simplicity and accuracy is achieved. Since no saturation is considered in the analytical model, there would be a noticeable error under heavy load conditions. However, variation trends of torque profiles against the key design parameters and current angle can be well predicted, which can still provide beneficial insight into the design and optimisation of multi-layer spoke type machines. The analytical model for spoke type machines with more PM layers can be developed by the same procedures proposed in the chapter.

Based on the analytical models derived, the influence of different design parameters is investigated to provide insight into the design of ferrite IPM machines. By introducing the secondary PM layer, the flux distribution can be improved with less harmonic contents, and the rotor saliency and reluctance torque can be significantly improved even with less PM.

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4 POWER LOSS AND THERMAL ANALYSIS OF FERRITE IPM MACHINE

Most studies in the field of electric machine design have only focused on electromagnetic aspects with constantly improving in the methods and modelling techniques. Along with this growth in these design methodologies, however, there is much lower concern regarding the thermal analysis and its

importance. Although this is true, the only thermal design aspect in the electromagnetic design methods is limiting the machine current to a value based on a standard value of the current density. The high-efficiency motor will reduce the overall operating cost of the electric drive system. In the chapter, the losses, efficiency and thermal analysis of the designed ferrite IPM motor will be comprehensively analysed.

4.1 Loss Composition and General Analysis

4.1.1 Energy Conversion and Loss Composition

There are various approaches to understand the force or torque developed in an energy conversion device. Figure 4-1 illustrates the energy conversion process based on the principle of conservation of energy, which states that energy can neither be created nor destroyed; can only be changed from one form to another. The electromechanical converter system has three essential parts: (1) an electric system, (2) a mechanical system, and (3) a coupling field[1].

The power in any motor can be transfer according to the following:

$$\begin{array}{l} \text{Electrical} \\ \text{input power} \\ \text{from source} \end{array} = \begin{array}{l} \text{Mechanical} \\ \text{output power} \end{array} + \begin{array}{l} \text{Increase in} \\ \text{stored power in} \\ \text{coupling field} \end{array} + \begin{array}{l} \text{power} \\ \text{losses} \end{array} \quad (4-1)$$

The electrical power cannot fully be converted into mechanical power, and the wasted power is the losses dissipated as heat. The electrical power loss is the copper loss due to current flowing in the winding of the energy converter, which

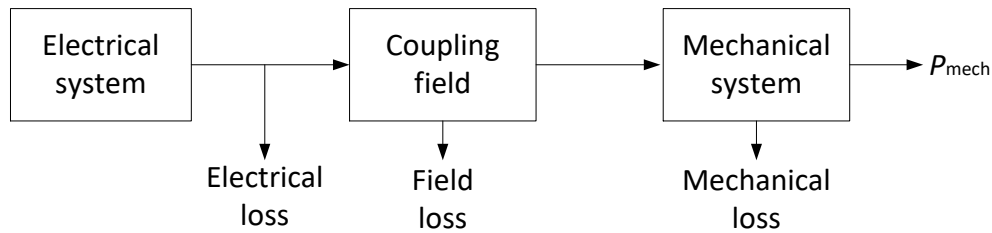


Figure 4-1 Electromechanical conversion system

is also known as heating loss or ohmic loss in the resistance of the winding. The field losses include the core loss in the stator and rotor lamination cores and eddy current loss in the PMs. The field losses are caused by the changing magnetic field in ferromagnetic materials. The mechanical loss is the friction and windage loss due to the motion of the moving components. All these losses are eventually converted into heat, which not only leads to a temperature rise of the motor but also influences its performance and lifetime to a certain extent.

Since the ferrite magnets used in our designed motor are nonconductive, there are no losses in the PMs. The relationship of power and losses in a ferrite IPM motor is illustrated in Figure 4-2.

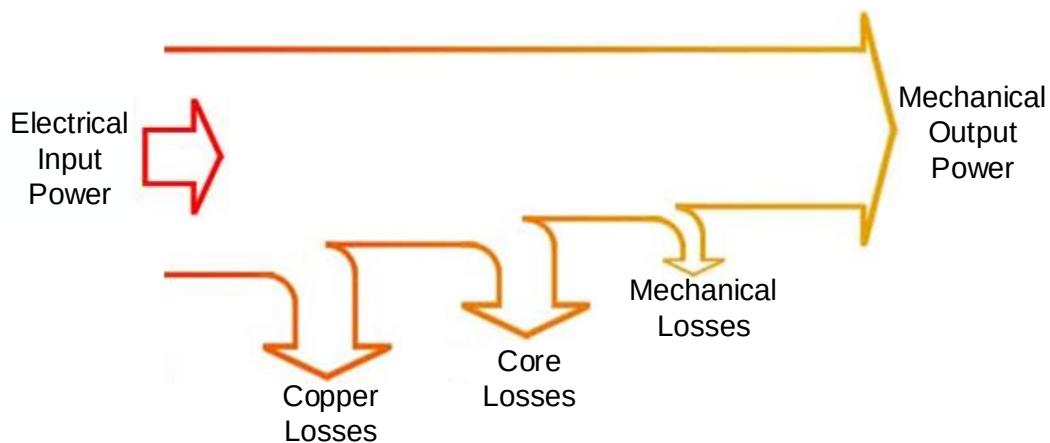


Figure 4-2 Power and loss flow diagram of the ferrite motor

The majority of losses come from copper losses and core losses, but mechanical losses cannot be neglected. The mechanical loss due to the bearing friction and wind drag is dependent upon rotational speed but independent on the load. It is difficult to estimate it directly by calculation because core loss is also dependent on rotational speed and coupled with mechanical loss. According to the rule of

thumb, the mechanical loss is normally approximated as 2% of the rated motor power

The two dominant losses, the copper loss in the windings and the iron loss in the stator and rotor core, are principally investigated in this study.

4.1.2 Theoretical Analysis of Copper Loss

Copper loss is always associated with the flowing of current in conductors due to their ohmic resistance. Thus, this resistivity of conductors is the main source of the ohmic power loss. This loss (copper) can be calculated from the following equation:

$$P_{cpl} = 3I^2R \quad (4-2)$$

Where I is the current in the coils and R is the copper coils resistance. The copper loss usually takes up a large part of the total loss. According to Ohm's law, the DC resistance of a conductor is dependent on its resistivity, length and cross-section area, and thus can be estimated. The electric resistance is calculated by

$$R = \rho \cdot \frac{L_{copper}}{A_{copper}} \quad (4-3)$$

Where ρ is the resistivity of copper, L_{copper} is the length of the copper wire in each phase, and A_{copper} is the cross-section area of the copper wire. The resistivity of copper at 20°C (293 K) is $1.68 \times 10^{-8} \Omega \cdot m$, namely original resistivity. However, most of the time, the motor is operating at 80°C at higher temperature because of the heat generated by the losses. After being started, the motor is warming up and stays at a specified temperature when the heat generation and dissipation reach their balance. According to the relationship between resistivity and temperature, the variation of the resistivity can be expressed by

$$\rho = \alpha \cdot \Delta T \cdot \rho_0 \quad (4-4)$$

Where $\Delta\rho$ is the change of the resistivity, α is the resistivity-temperature coefficient, ΔT is the change of temperature, and ρ_0 is the original resistivity. With

regard to copper, the resistivity-temperature coefficient $\alpha = 0.0039 \text{ K}^{-1}$, and the temperature difference $\Delta T = 60^\circ\text{C}$. Then the change of the resistivity can be calculated, which is $0.39312 \times 10^{-8} \Omega \cdot \text{m}$. So the resistivity of copper at 80°C is $2.07312 \times 10^{-8} \Omega \cdot \text{m}$.

All the coils of one phase are serially connected and the length of the copper wire can be estimated by the below equation [2]:

$$L_{copper} = \left(2L_{eff} + \left(L_{tooth} + \frac{B_{s1} + B_{s2}}{2} \right) \cdot N_p \right) \cdot N_c \cdot N_s / 3 \quad (4-5)$$

Where B_{s1} and B_{s2} are the top and bottom width of the slot. L_{eff} is the axial length of the motor, L_{tooth} is the width of the tooth, N_p is the winding pitch numbered in slots, N_c is the number of the conductors in a coil and N_s is the number of all slots, as mentioned before. And the cross-section area of the copper wire is

$$A_{copper} = e \cdot \frac{A_{slot}}{N_c} \quad (4-6)$$

Where e is winding space factor, which is 0.5 in this case; and A_{slot} is the cross section area of the slot, which can be obtained from the simulation model.

In sum, the copper loss can be computed by

$$P_{cpl} = I^2 \cdot \rho \cdot \frac{\left(2L_{eff} + \left(L_{tooth} + \frac{B_{s1} + B_{s2}}{2} \right) \cdot N_p \right) \cdot N_c^2 \cdot N_s}{3e \cdot A_{slot}} \quad (4-7)$$

When the motor is manufactured, the resistance of the winding can be easily measured by multi-meters, so that the copper loss can be calculated directly by equation (4-2).

In the alternating current (AC) system, some more issues that are necessary to be considered are skin effect and proximity effect, which make the resistance of a conductor increase with the AC frequency. The skin effect is the tendency of AC current becoming distributed so that the current density decreases greatly with the depth inside a conductor. In other words, the current tends to flow at the

skin of the conductor and the skin depth is usually used to describe the severity level of skin effect [3]:

$$\delta = \sqrt{\frac{\rho}{\pi f \mu_0 \mu_r}} \quad (4-8)$$

where δ is the skin depth, f is the operating frequency, ρ and μ_r are the resistivity and relative permeability of the conductor. In this case, copper wires with a diameter of 0.6mm are used, and the skin depth at rated speed is 8.23 mm that is much larger than the diameter of the copper conductor. So the skin effect can be neglected in our study case.

In multi-stranded coil systems, the current density distribution in one conductor becomes non-uniform because of the alternating magnetic field associated with the AC current in adjacent conductors. This phenomenon is termed the proximity effect. According to [3], the increment factor caused by the proximity effect can be represented by

$$k_{rpac} = \frac{R_a - R_{dc}}{R_{dc}} = \frac{\zeta}{2} \frac{\sinh \zeta + \sin \zeta}{\cosh \zeta - \cos \zeta} - 1 \quad (4-9)$$

$$\zeta = \frac{d_c \sqrt{\pi}}{2\delta} \quad (4-10)$$

where R_a and R_{dc} are AC and DC resistances of the winding respectively, ζ is an intermediate parameter, d_c is the diameter of the copper wire in the stator slots, and δ is the skin depth calculated by Equation (4-8). According to the parameters of the proposed motor design, the increment factor k_{rpac} is calculated by the above equations (4-9) and (4-10). k_{rpac} is less than 10^{-7} , so the proximity effect can also be ignored for the copper loss estimation.

From the discussion above, both of the skin effect and proximity effect can be ignored in our study, and thus the estimation of copper losses considers only considering the ohmic losses of DC resistance.

4.1.3 Theoretical Analysis of Core Loss

Another commonly found loss in motors is the iron loss that comes from the magnetic flux undergoing time variations and intersects with the core. It increases drastically at high rotation speeds and high magnetic flux densities. The increase of core loss can lead to a rise in temperature and a reduction in efficiency. Consequently, it is of great importance to predict iron loss at the motor design stage.

Iron loss incorporates hysteresis loss and eddy current loss. Hysteresis loss is produced by the magnetic hysteresis effect. As shown in Figure 4-3(a), there is net energy flowing from the current source to coil-core assembly during a cycle of the hysteresis loop. The loss of power is called hysteresis loss, which goes to heat the core. Eddy current loss is produced by eddy current that is generated when the ferromagnetic core undergoes time variations of magnetic flux. The cross section of a core is illustrated in Figure 4-3(b), through which the magnetic flux is rapidly changing. This change has induced by a voltage in the path and circulating (eddy) current flows around the path accordingly. This current generates an ohmic loss appears as heat in the core, which is called eddy current loss.

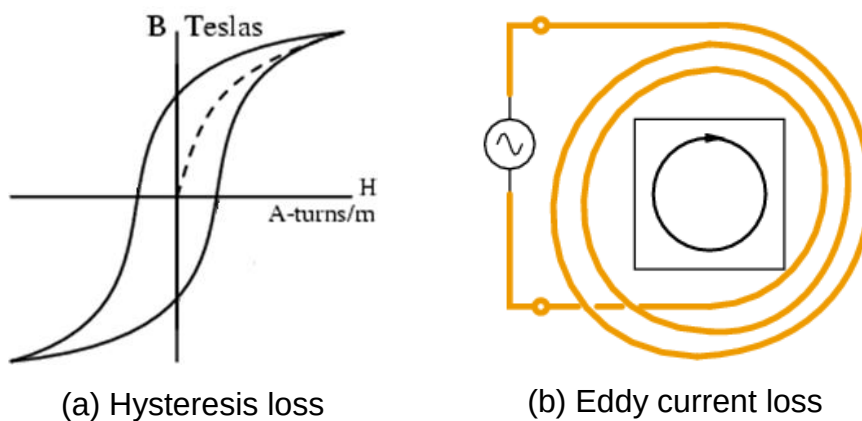


Figure 4-3 Mechanism of hysteresis loss and eddy current loss

Both of these losses increase when the magnetic flux density increases or the time variation frequency rises. The iron-core loss is expressed as:

$$p_v = p_h + p_c + p_e = K_1 B_m^2 + K_2 B_m^{1.5} \quad (4-11)$$

Where the eddy-current loss is

$$p_c = k_c (f B_m)^2 \quad (4-12)$$

The hysteresis loss is

$$p_h = k_h f B_m^2 \quad (4-13)$$

And the excessive loss is

$$p_e = k_e (f B_m)^{1.5} \quad (4-14)$$

Therefore

$$K_1 = k_h f + k_c f^2 \quad (4-15)$$

$$K_2 = k_e f^{1.5} \quad (4-16)$$

The classical eddy-current loss coefficient is calculated directly as

$$k_c = \pi^2 \sigma \frac{d^2}{\delta} \quad (4-17)$$

Where σ is the conductivity and d is the thickness of one lamination sheets. Minimise the quadratic form to obtain K_1 and K_2 .

$$err(K_1, K_2) = \sum_i [p_{vi} - (K_1 B_{mi}^2 + K_2 B_{mi}^{1.5})]^2 = \min \quad (4-18)$$

Where P_{vi} , B_{mi} – the i -th point of the data on the loss characteristics curve. The other two loss coefficients are obtained as

$$k_h = \frac{K_1 - k_c f_0^2}{f_0} \quad (4-19)$$

$$k_e = \frac{K_2}{f_0^{1.5}} \quad (4-20)$$

where f_0 is the testing frequency for B-H Cure.

Unfortunately, it is extremely difficult to obtain iron losses accurately in studies that use the magnetic circuit method or rules of thumb. In order to obtain them accurately, one needs to find the distribution and time variations of the magnetic flux density in each part of the motor after accounting for a fine geometry and the material's nonlinear magnetic properties. Using the finite element method (FEM) is essential in order to carry out this kind of detailed analysis. In order to simulate the physical phenomena in an analysis, firstly FEA models are created and simulated under various loading conditions. Then the time variations in the magnetic flux density produced in each part of the motor are obtained. With the results of the amplitude and frequency of the magnetic flux density variations for each element in the model, the iron loss distribution and iron losses can be calculated according to equations (4-11)–(4-20) by referring to the material's iron loss characteristics.

4.2 Iron Loss Analysis Considering PWM Effect

Vector controls using a PWM (Pulse Width Modulation) control are commonly included in the drive circuits of high efficiency IPM motors. A PWM control makes it possible to adjust the phase or amplitude of a current according to load and rotation speed so that they can achieve high efficiency in a wide operation range. The control frequency of a PWM is called a carrier frequency. To form the current waveform supplied by the PWM control, the carrier harmonic current is superimposed on the basic wave current. This carrier harmonic current applies a high-frequency magnetic field to each part of the motor. As a result, an extra iron loss is generated in the IPM motor's core. The total amount of these losses is not a dominant factor, but they can be a hindrance when trying to raise efficiency. The effects of a carrier harmonic against an IPM motor's iron loss can be evaluated by inputting the measured current waveforms into a magnetic field analysis.

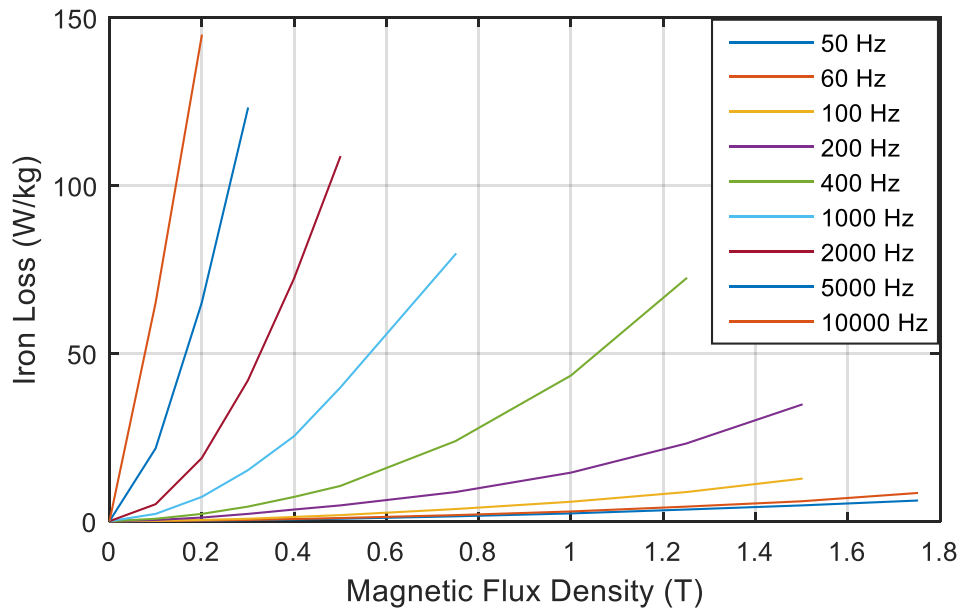


Figure 4-4 Iron loss characteristics of stator and rotor cores

Magnetic flux that varies with time is produced in each part of a motor when a current that contains a PWM carrier harmonic is input. The main components of iron loss are eddy current loss, which is caused by eddy currents produced from time variations in the magnetic flux, and hysteresis loss, which is produced from magnetic hysteresis. The iron loss characteristics of stator and rotor cores are illustrated in Figure 4-4. . Due to limitations of the software lines show combination of linearized lines, but it should be continues. The iron loss density is increased with magnetic flux density and variation frequency. When the frequency is very high, even if the magnetic flux density is relatively low, lots of losses can be generated in the motor cores. That is why the high-frequency harmonic components of the current generated by PWM inverter can increase iron losses.

The control frequency of the PWM inverter that used to drive the ferrite IPM motor is 8 kHz. When the motor is operating at rated speed and load, the three-phase current is measured (experimentally) and depicted in Figure 4-5. Since the motor has three pole-pairs, the fundamental frequency of the current waveform is 75 Hz ($f = \text{speed} \times \text{pole-pairs} / 60$) and the period is 0.013333 s at rated speed 1500 rpm. The amplitude of the fundamental current component is 8.8 A, which is the same as the amplitude of

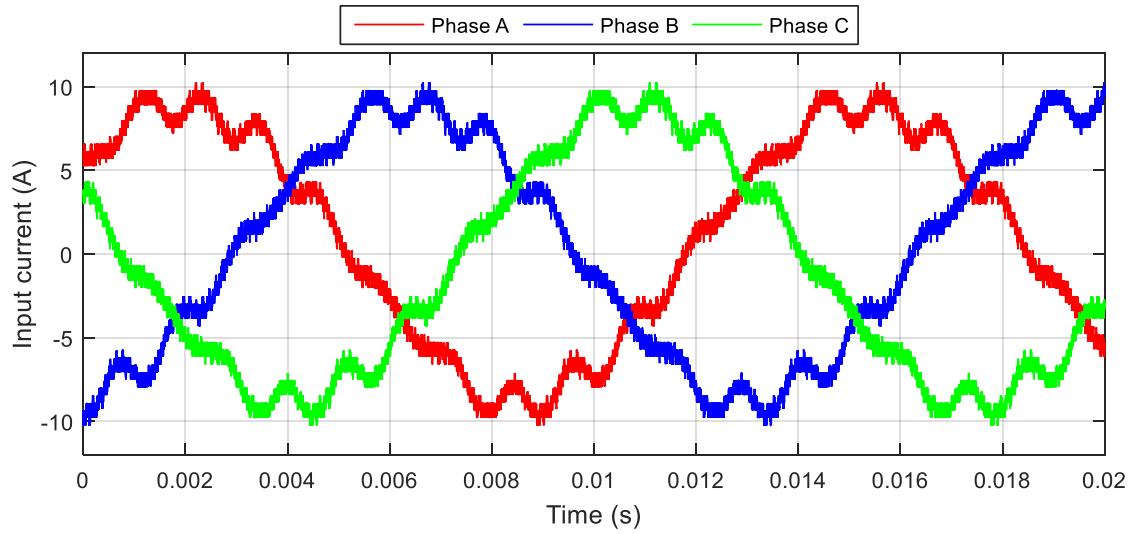


Figure 4-5 Actual current waveforms generated by PWM inverter

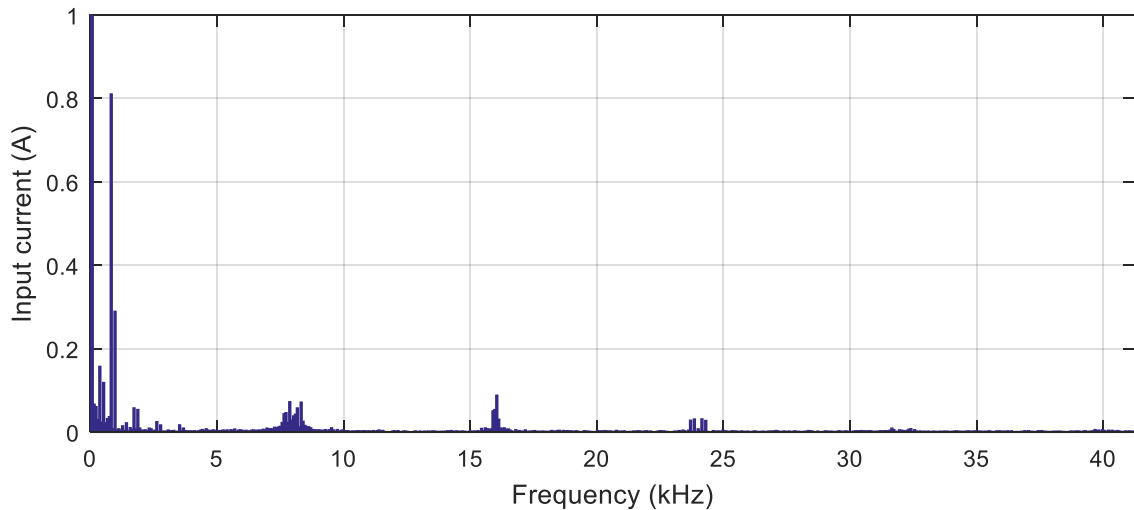


Figure 4-6 Spectrum of actual current waveforms generated by PWM inverter

conventional sinusoidal current. Apart from the fundamental component, the measured current has a lot of harmonic components that are displayed in Figure 4-6, which makes the actual current waveform bump not in an ideal sinusoidal shape. Aside from the major harmonics in relatively low frequency which is mainly due to machine configurations, it is worth noting that the PWM inverter generated current has many high-frequency harmonic components mainly concentrating around 8 kHz, 16 kHz, 24 kHz, 32 kHz and 40 kHz. These high-frequency components are caused by PWM carrier waves, the frequency of which is 8 kHz. Thus, the motor input current from the PWM inverter drive will induce more iron

loss than that from the conventional sinusoidal current drive. A comparison of the iron losses, hysteresis losses, and eddy current losses when the IPM motor is driven by a PWM inverter and a sinusoidal current are shown in Table 4-1.

Table 4-1 Iron losses caused by different current drives

	Rotor Core		Stator Core	
	Sinusoidal current drive	PWM inverter drive	Sinusoidal current drive	PWM inverter drive
Iron loss	9.49 W	13.55 W	38.28 W	46.54 W
Hysteresis loss	1.66 W	2.12 W	14.79 W	15.53 W
Eddy current loss	7.83 W	11.43 W	23.49 W	31.01 W

The iron losses in both the rotor core and the stator core increase, when the IPM motor is run by a PWM inverter. The eddy current losses increase the most severely, increasing 46% in the rotor core, and 32% in the stator core. The hysteresis losses are aggravated as well, rising 28% in the rotor core, and 5% in the stator core. The iron losses of the whole motor are increased from 48 W in the sinusoidal current drive to 60 W in PWM inverter drive by 26%, with eddy current losses increased by 36% and hysteresis losses increased by 7%. The different losses of the IPM motor driven by a PWM inverter and a sinusoidal current drive are clearly illustrated in Figure 4-7 and Figure 4-8 by frequency spectra, which help to make comparisons.

Since the studied ferrite IPM motor has 6 magnetic poles in the rotor and 36 slots in the stator, the fundamental frequencies are 75 Hz in the stator core and 450 Hz in the rotor core when it is run at rated speed 1500 rpm. Therefore, the significant components of eddy current loss and hysteresis loss in the rotor core are the multiples of the fundamental frequency 450 Hz, such as 900 Hz, 1800 Hz, and 2700 Hz, as shown in Figure 4-7. When the motor is driven by PWM inverter, more eddy current losses are generated in the rotor core, than it is driven by the sinusoidal current. That is because the eddy current loss in PWM inverter drive has a larger 2nd order component (900 Hz) and lots of high-frequency components around 8000 Hz that are caused by PWM carrier waves. The

difference of hysteresis loss in the rotor is primarily from the 2nd order component at 900 Hz. The PWM driven motor has 34% more hysteresis loss.

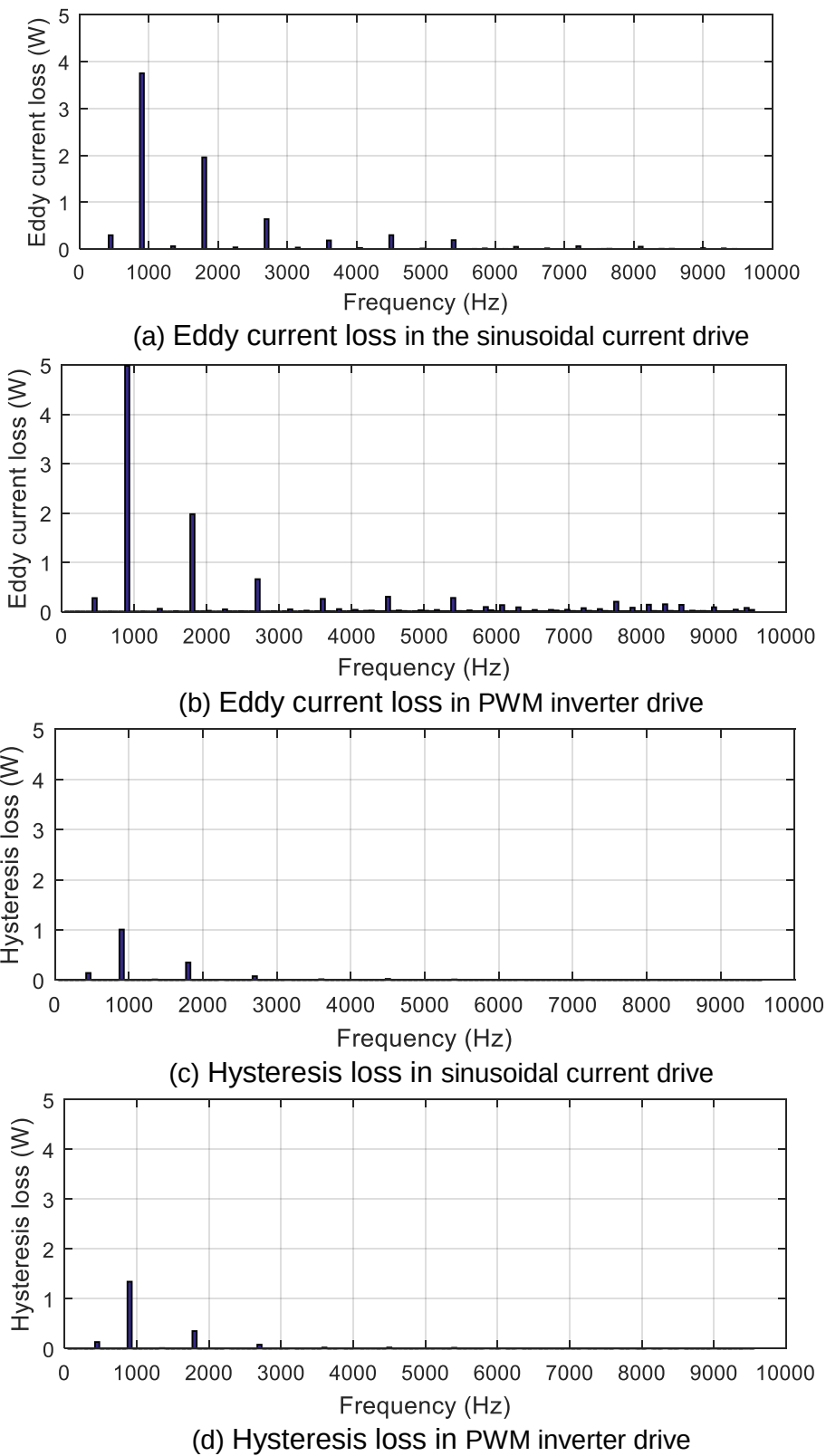
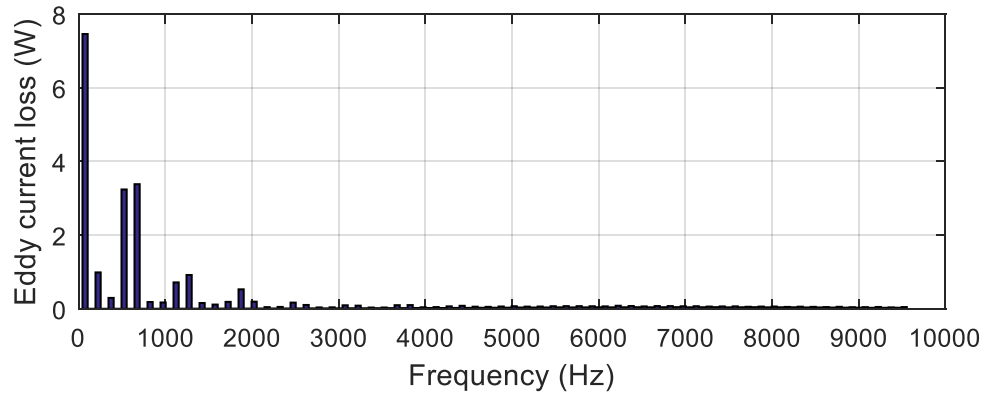
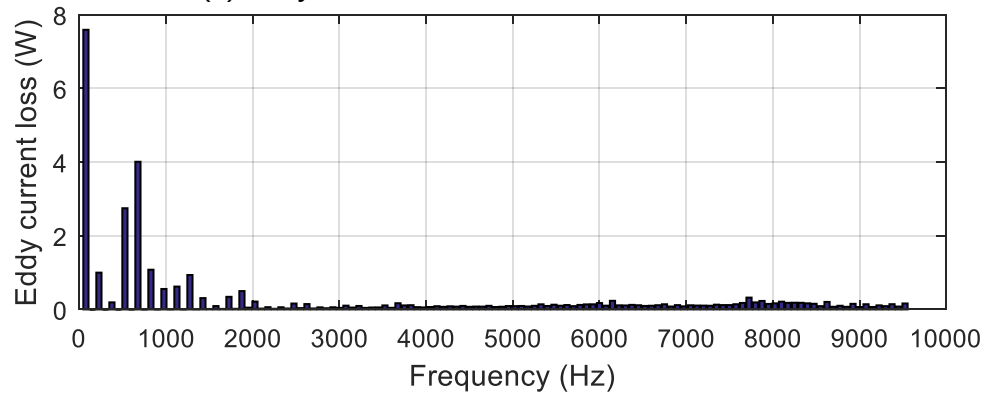


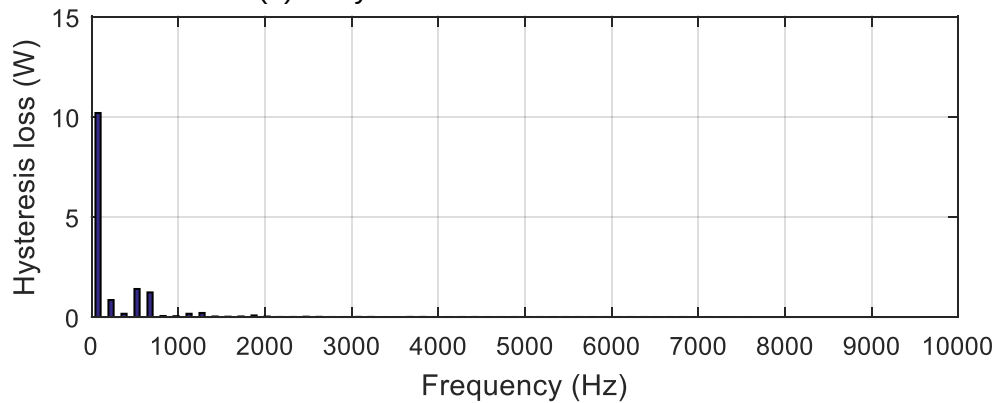
Figure 4-7 Iron loss frequency components in the rotor core



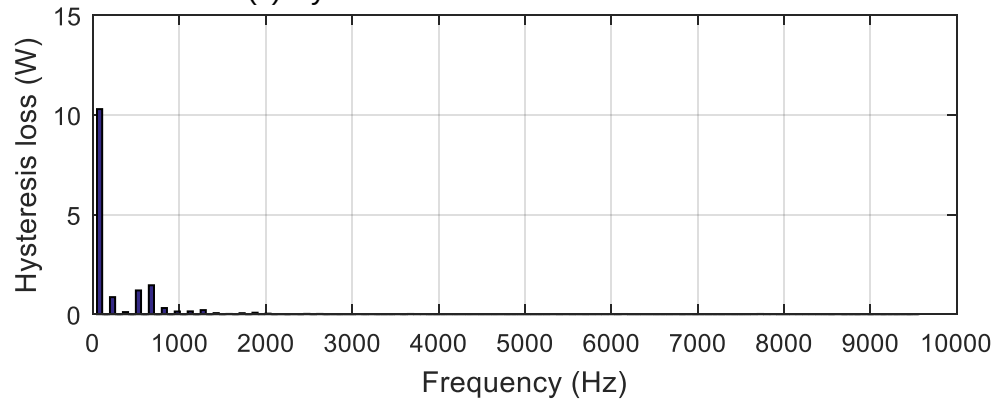
(a) Eddy current loss in the sinusoidal current drive



(b) Eddy current loss in PWM inverter drive



(c) Hysteresis loss in sinusoidal current drive



(d) Hysteresis loss in PWM inverter drive

Figure 4-8 Iron loss frequency components in the stator core

The significant components of eddy current loss and hysteresis loss in the stator core are the multiples of the fundamental frequency 75 Hz, such as 75 Hz (1st order), 525 Hz (5th order), and 675 Hz (7th order), as shown in Figure 4-8. The IPM motors driven by both PWM inverter and sinusoidal current have almost the same amount of eddy current loss in the stator in relatively low frequency, under 3000 Hz. But in the high-frequency region, the PWM inverter driven motor exhibits more eddy current losses, especially around 8000 Hz, which are generally induced by PWM carrier waves. There is no much difference in the hysteresis loss in the stator core, and only some of the low-frequency components in PWM inverter driven motor have slightly larger losses.

In addition, more than three quarters of the iron losses are generated in the stator core whatever the current drive is used, because the fundamental current component, the largest one, does not cause any iron loss in the rotor core in a synchronous motor. More than two-thirds of the iron losses are from eddy current losses whatever the current drive is used. That is because the eddy current loss is proportional to the square of variation frequency, but hysteresis loss is linear with variation frequency, which is indicated in equations (4-12) and (4-13). Hence more eddy current losses are produced when motor speed increases and the high-frequency harmonic components of input current will contribute more eddy current losses than hysteresis losses. This is obviously depicted in Figure 4-7 and Figure 4-8. In both rotor and stator cores, there are almost none or very fewer hysteresis losses in high frequency, but, in contrast, there are many perceivable eddy current losses in the high-frequency area.

The difference of iron losses generated by PWM inverter driven motor and sinusoidal current driven motor can also be observed in the iron loss distributions in Figure 4-9 and Figure 4-10. In the rotor core, the majority of the iron losses occur around the circumference of the rotor, where the flux density is high. Since the eddy current loss is larger than the hysteresis loss in the IPM motor, there are more high loss density regions in Figure 4-9 (a) and (b) than that in Figure 4-9 (c) and (d). Comparing the eddy current loss density in Figure 4-9 (a) and (b), it is found that the motor driven by PWM inverter has more eddy current loss at the

inner part of the rotor core. There is no much difference between the hysteresis loss density in Figure 4-9 (c) and (d), while the PWM inverter driven motor has a slightly higher loss in the inner side part of the rotor. Comparing with sinusoidal current drive, the PWM inverter driven motor has more iron loss distributed in the inner areas of the rotor core, and it has larger high loss density areas around rotor circumference, which are shown in Figure 4-9 (e) and (f). Figure 4-10 shows the loss distributions of eddy current loss, hysteresis loss and iron loss in the stator. In all these six distribution graphs, the loss density decreases from inside to outside. The loss density at the tooth tips is the highest, followed by the stator teeth, and at the stator yoke, the loss density is the lowest. Comparing the eddy current loss density in Figure 4-10 (a) and (b), it is seen that the PWM driven motor has higher loss density in the teeth and a little more eddy current loss distributed in the stator yoke. There is no observable difference of the stator hysteresis loss distribution in Figure 4-10 (c) and (d), as they have almost the same amount of hysteresis losses. Combining eddy current loss and hysteresis loss, the stator iron loss density in PWM inverter driven motor is higher than that in sinusoidal current driven motor, as shown in Figure 4-10 (e) and (f). As both of its rotor and stator have higher loss density, the motor powered by PWM inverter drive generates more iron losses. In this section, the effects of PWM carrier harmonics on the iron loss of the ferrite IPM motor are analysed. The current waveform that is supplied by PWM inverter has PWM carrier waves superimposed on the basic current wave control. And with this PWM carrier harmonic current, high-frequency of the magnetic field is applied to the motor core and more iron loss occurs.

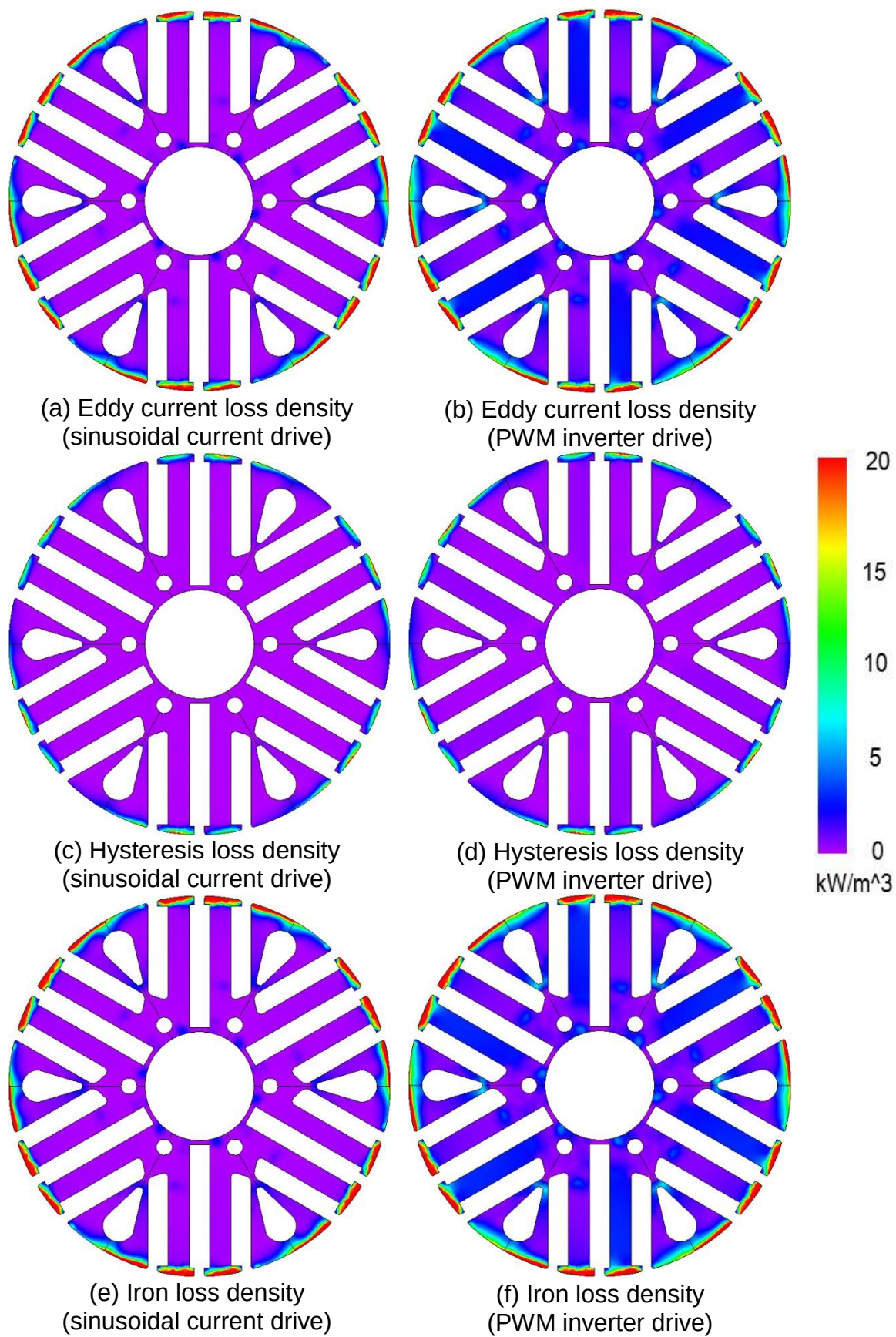


Figure 4-9 Iron loss density distribution in the rotor core

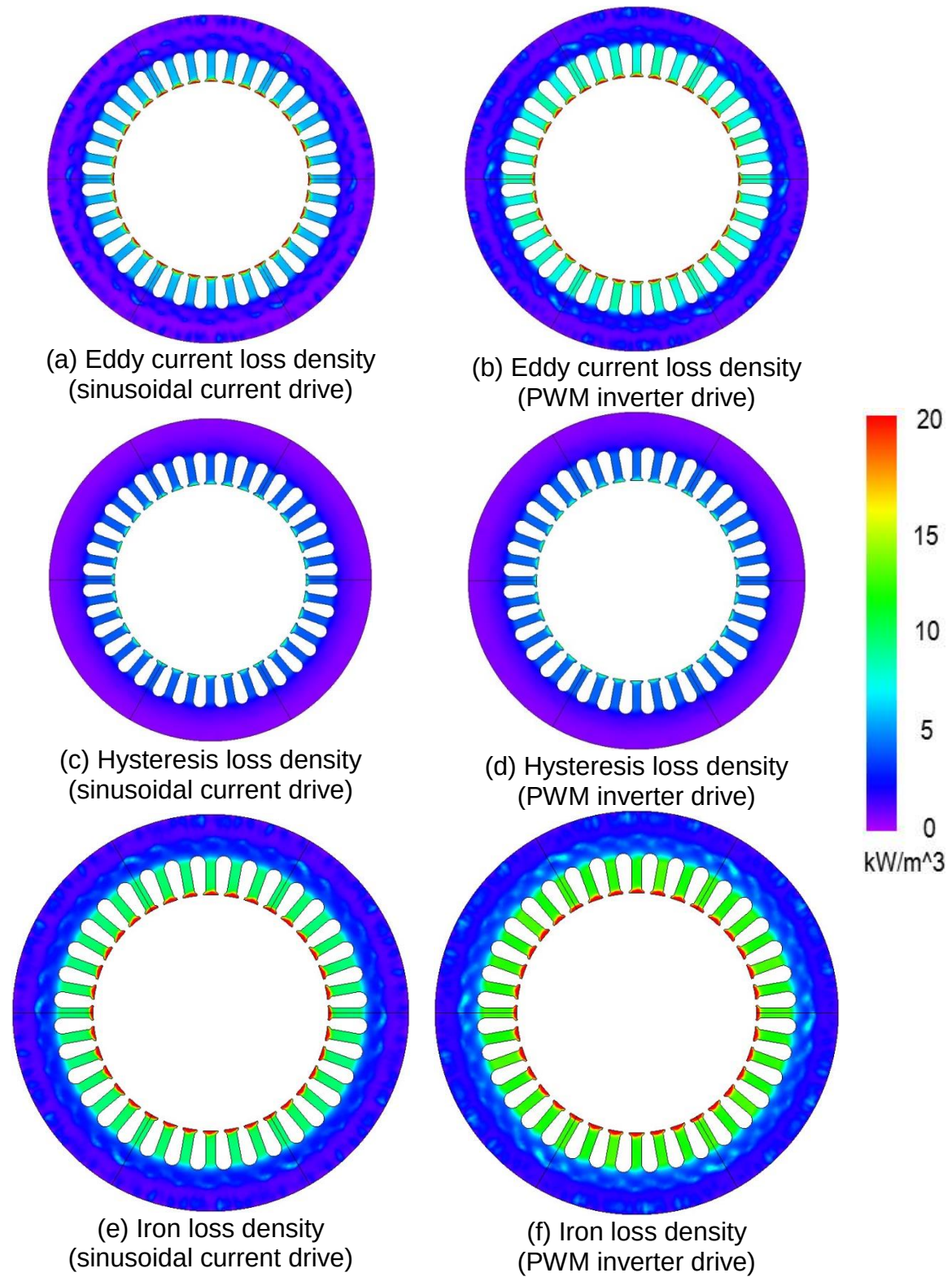


Figure 4-10 Iron loss density distribution in the stator core

The PWM affected iron loss is discussed in detail, broken down into eddy current loss and hysteresis loss, the losses in the rotor and in the stator, and it is

compared with the iron loss caused by conventional sinusoidal current drive. From the comparison of iron loss frequency components and iron loss distribution, it is concluded that the motor driven by a PWM inverter will generate more iron losses than that driven by the conventional sinusoidal current. Thus, the effect of PWM should be considered when an actual prediction of iron loss is required. In order to improve the efficiency of the PWM electric drive, the current harmonics should be further minimised. And increasing the frequency of the PWM carrier wave will be helpful to mitigate the effects of PWM on iron loss.

4.3 Thermal Analysis and Temperature Distribution

Most studies in the field of electric machine design have only focused on electromagnetic aspects with constantly improving the methods and modelling techniques. Along with this growth in these design methodologies, however, there is much lower concern regarding the thermal analysis and its importance. Although this is true, the only thermal design aspect in the electromagnetic design methods is limiting the machine current to a value based on a standard value of the current density. One of the feasible methods of achieving a higher power density machine is realized by choosing the maximum available limit of the standard current density for a given application. Simultaneously, high current density leads to a high current for a given conductor cross section area results in significant copper loss and, in turn, high hot-spot temperatures [4,5].

Furthermore, running the machine at high speeds leads to a higher current and frequency, which results in higher copper and iron losses. Improving the efficiency of the machine while keeping the torque or current density at a high level is one of the main goals of many recent researchers even when this improvement is a small fraction. The high computational power and the necessity of efficiency improvement are the motives behind the feasibility of including the thermal analysis as a tool for further optimisation. The loss in the magnetic field can be considered as the heat source of the thermal model, which can be obtained from the electromagnetic analysis. However, the thermal analysis of electrical motors still not investigated as the electromagnetic analysis, which is clearly shown in several published papers. The lack of thermal analysis due to

complexity is addressed by recently published papers [6-8]. Thermal analysis of an electrical machine comprises three basic types: analytical approach (lumped-circuit), finite element analysis (FEA), and numerical methods. The first approach is positively very fast to calculate; however, the main heat transfer paths need a huge effort to be modelled accurately [6-9]. Thermal analysis is formulated by assuming three types of heat transfer resistance for different parts of the machine, and they are conduction, convection, and radiation. The advantage of this formulation is the simplicity of resistances representation. The conduction resistance is equal to the path length divided by the product of the path area and the thermal conductivity of the material. The convection and radiation resistances are equal to one divided by the product of the surface area and the heat transfer coefficient. Nevertheless, an accurate FEA solution requires not only the knowledge of geometry and material properties of the machine construction but also the thermal parameters. The main benefit of using FEA solution is the accurate calculation of conduction heat transfer in complex geometric shapes, such as heat transfer through strands of a copper conductor in a slot. A comprehensive understanding and modelling of the loss production mechanism are required to accurately predict the main heating source in the machine [10]. Thus, it is also important to study a thermal design that improves heat dissipation and reduce the temperature rise to an acceptable range. In electrical machines temperature influences lifetime/degradation and performance. Current density, which is usually proportional to torque density, is thermally limited. Over-temperature will lead to premature failure, for example of the stator winding insulation. The insulation lifetime of a machine varies inversely with working temperature. Thus, it is necessary to evaluate the thermal properties of the motor at the design stage.

The losses discussed in the above sections are the heat sources in the motor's thermal system. A magnetic field-thermal analysis model is created based on the motor's structure and operating environment, which is used to investigate the motor's thermal properties in a steady state and transient state. The consecutive operating time in the overload condition is evaluated as well, which is a critical characteristic of a motor's power capability.

4.3.1 Thermal Modelling of the Machine

To carry out an accurate thermal analysis, it is necessary first correctly to understand the amount and location of heat generation. It would be advantageous to calculate the losses in a magnetic field analysis simulation using the finite element method, and from there to carry out a thermal analysis using the resulting loss distribution.

Table 4-2 Loss in each part at rated load

Heat generation parts	Iron loss in the rotor core	Iron loss in the stator core	Copper loss in the coils
Losses (W)	9.49	38.28	117.8

Table 4-2 shows the loss in each part of the motor at rated load and speed. Copper loss in the coils and iron loss in the core are the dominant heat sources, so this analysis mainly evaluates the effects of this heat. The loss in the coil is relatively large, about 2/3–3/4 of the total loss, so it is expected to be the main source of heat. On the other hand, most of the iron loss comes from the stator core, because in synchronous machines the fundamental component of the magnetic field does not generate an iron loss in the rotor, only harmonics causing losses. Even though the loss in the rotor is only about 5% of a total loss, the heat will be accumulated due to poor dissipating condition and cause rotor temperature to increase. The properties of ferrite magnet, such as residual flux density and coercivity, are sensitive to temperature. What is more, the heat resistance of the ferrite magnet is low, so lots of attention is required to prevent it from being irreversibly damaged. The models of the machine are demonstrated in Figure 4-11. According to the structure of the designed ferrite IPM machine, a thermal circuit model is built and shown in Figure 4-12. The square blocks in the circuit represent the motor's core parts that generate and transmit heat, and they meshed in FEA simulation. The circle blocks represent the surroundings and ambiances, and they absorb the heat from the inner motor.

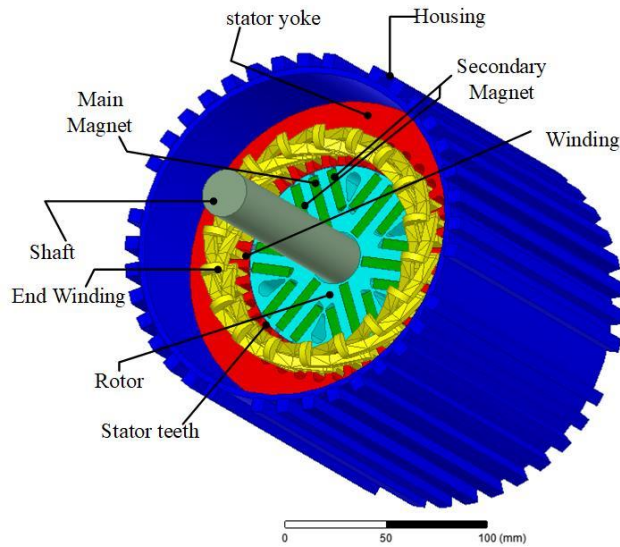


Figure 4-11 3D models of the two-layer spoke-type machine.

The resistance signs are illustrated as *thermal resistance* among the square and circles blocks, such as thermal resistance between coil and stator core, stator core and cover, cover and air, and so on. The capacitor sign attached to cover is a *heat capacitor*, and the cover's heat capacity is related to its volume, density and specific heat. All of the heat will eventually transfer into the ambient air. As is illustrated in Figure 4-12, the rotor has three routes to transfer its generated heat. Some of the heat is brought out through the rotor shaft to the housing cover; some go through the internal air to the cover, and the others are transferred to the stator core by the air gap. For the heat occurs in copper coils, most of it is transmitted to stator core and the rest is dissipated to housing cover through internal air. Since stator core is seamlessly connected to housing cover and has large heat radiating surface, the majority of the heat generated in the motor is finally dispelled away through stator core to the cover, and only a small fraction is transferred through internal air to the cover. Based on the thermal circuit model the temperature of the ferrite IPM machine can be calculated.

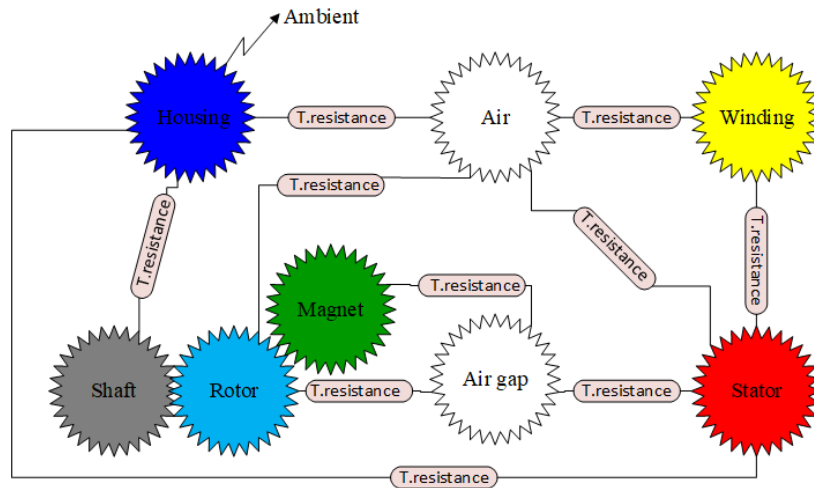


Figure 4-12 Thermal circuit model of the designed motor

4.3.2 Temperature Distribution at Steady State

The simulation of the thermal circuit model is carried out with a rated motor load. The motor operates continuously until a thermal balanced state is reached. The final balanced state is also called a steady state, and the temperature distribution of it is shown in Figure 4-13. According to the temperature distribution, the motor is generally divided into three thermal regions: the copper coils, the rotor and the stator. The highest temperature region is the copper coils, as the copper loss is the dominant source of heat. The highest temperature is about 104°C. Most of the heat from the coils is transferred to the cover through the stator core, and the rest goes through the air inside of motor to the cover and finally dissipates to the ambience. The moderate temperature region is the rotor. Since ferrite PM is non-conductive, no loss is generated in magnets. The heat accumulated in the rotor is mainly from its iron loss. As the majority of the rotor's surface area is encompassed by air, which has high heat transfer resistance, it is hard for the rotor to dispel the generated heat. The accrued heat increases the temperature of the rotor and the magnets embedded. The rotor has a higher temperature than the stator core. Most of the heat generated in the rotor flows through the air around it and then to the stator and cover, and the rest is transferred to cover through the bearings and shaft.

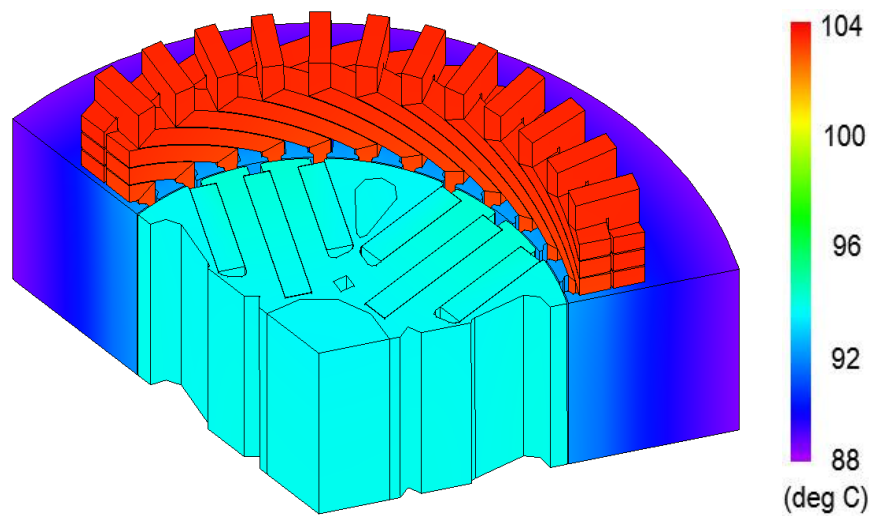


Figure 4-13 Temperature distribution at a steady state

The relatively low temperature region is the stator core, with the lowest temperature about 88°C at the contact surface to cover. The stator core is not only a heat source that generated from iron loss but also a very important heat transfer channel/medium, through which most of the heat generated inside of the motor is transferred to the motor housing/cover and finally dissipated into the ambient air. It is depicted in Figure 4-13 that from the outer to the inner side of the stator core the temperature is increasing gradually. The outer side is cooler because it is connected to a cooling source, the cover. The inner side is hotter as heat is concentrated, where the stator iron loss density is the highest and it also receives a lot of heat from copper windings and rotor core.

With the help of temperature distribution analysis, several thermal design approaches can be concluded for the ferrite IPM machine:

- Fill the gap between coils and stator core with high thermal conductivity material, so that heat from copper loss can be more easily transferred to stator core and dissipated to the ambience.
- Minimise the air gap between rotor core and stator core, to improve the heat transfer efficiency from the rotor to the stator core.
- Enhance the air circulation inside of the motor to speed up the heat transferring through the air, such as installing fans.

- Cool down the motor housing/cover, the larger the temperature difference between stator and cover, the faster the heat is transferred.

4.3.3 Temperature Variation Process

To further understand the process of temperature changing, the temperature of some typical points in the motor is measured and compared. The measuring points are indicated in Figure 4-14, which include points on rotor core, coil, magnet, and stator core. The initial temperature is 20°C. The temperature measurement lasts for four hours and it is recorded in Figure 4-15. The temperature of the four points increase fast in the first two hours and gradually become stable in the last two hours. In contrast with the temperature distribution at a steady state, the increased temperature is quite close to the balanced thermal temperature after four hours of running.

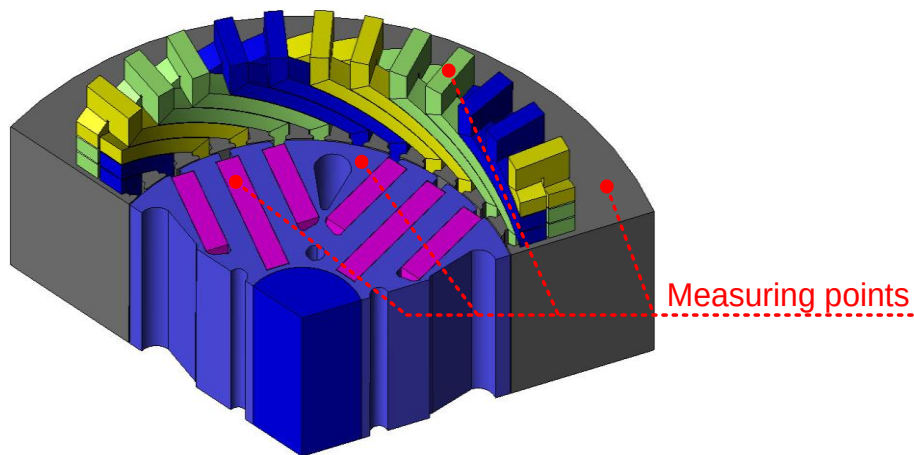


Figure 4-14 Points of measuring temperature for each part

The highest temperature appears in the coil. As the main source of heat comes from copper loss, the winding temperature is growing faster at the beginning and more than 10°C higher than other measuring points at the end. In the first 100 minutes, the temperature of the stator core is higher than the rotor core and magnet, but it is exceeded later. The losses in the stator are much higher than the rotor, and the stator temperature rises much faster than the rotor at first. However, the rotor has worse heat dissipation conditions as it is largely surrounded by air and the heat generated in the rotor is difficult to transfer to the

outside; while the stator core is directly (interference) fitted to motor housing that transfer the heat from stator efficiently. The accumulated heat in rotor makes it hotter than stator in relatively long running time. The results are consistent with the temperature distribution of the steady state in Figure 4-13.

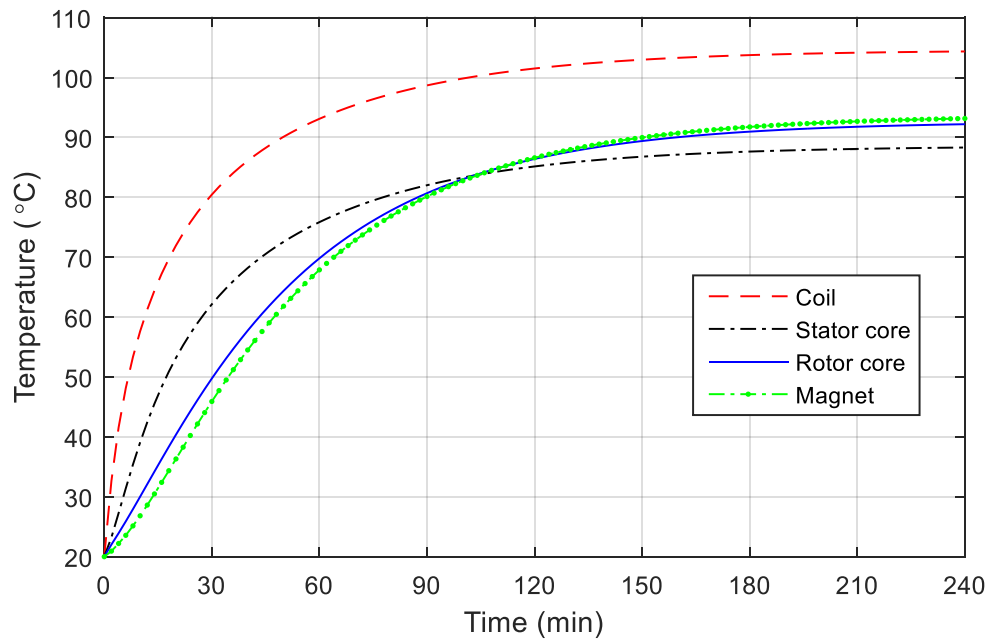


Figure 4-15 Temperature variation of the measured points

The rotor core itself is also a heat source from its iron loss. The ferrite magnet does not generate heat but is heated by its adjacent rotor core. So, there is a temperature difference at the beginning between the two parts. But the difference is reduced with time and almost to zero in the thermal balanced state. In addition, the property of the ferrite PM is affected by the increased temperature of the rotor. Thus, special attention should be paid to the rising temperature during operation, and make sure the PMs are operating in an acceptable temperature range.

4.3.4 Continuous Operating Time at Over-load Conditions

In some applications, motors are required to work in overloaded conditions for a short while. The over load capacity is an essential characteristic to evaluate a motor and useful for motor selection. Larger output power requires a larger input current. Although increasing the current produces higher torque, it also boosts

the heat generation in terms of copper losses and core losses. Owing to the large input current, the copper losses generated in the windings increase exponentially. Along with coil temperature surging up, the resistivity of the copper rises, which further deteriorates the copper losses. The iron losses in stator and rotor cores are also increased due to the large armature reaction of the designed IPM machine. Accordingly, the temperature of the stator and rotor will increase dramatically with current, and the properties of the ferrite PMs will be worsened by high temperature.

Table 4-3 Current and average torque

Input Current	100%	200%	300%
Average Torque (pu)	1	2.1	3.1

Therefore, it is vital to find the longest consecutive operating time under overloaded condition before causing any irreversible damage. In addition to the evaluation under rated load in subsection 4.3.3, simulations with two and three times the rated input current are carried out, and the temperatures of coil and magnet are analysed and discussed. The rated current is 6.2 A (RMS value), designated as 1 C, and the corresponding average torque is 13.32 Nm, designated as 1 pu. When the input current is increased from 1 C to 3 C, the output torque grows from 1pu to 3.1pu, which are summarised in Table 4-3.

The temperature measuring points are the same as the ones in Figure 4-14, and temperature variations of the coil and magnet with various currents are recorded in Figure 4-16 and Figure 4-17 respectively. The enamelled copper wire used in the designed motor is in thermal Class H (180°C), and the temperature limit is indicated in the green dash-dot line in Figure 4-16. The shape of the temperature variation lines with various currents is similar, but the slop of them is different. The larger the input current, the steeper the temperature changing line is. In rated load condition (input current is 1 C) the temperature of the coil is always under the temperature limit; in double load condition (input current is 2 C) it takes about 13 minutes to reach the temperature limit; and in triple load condition (input current is 3 C) the coil's temperature exceeds the temperature limit in about 3 minutes, which is much shorter than double load.

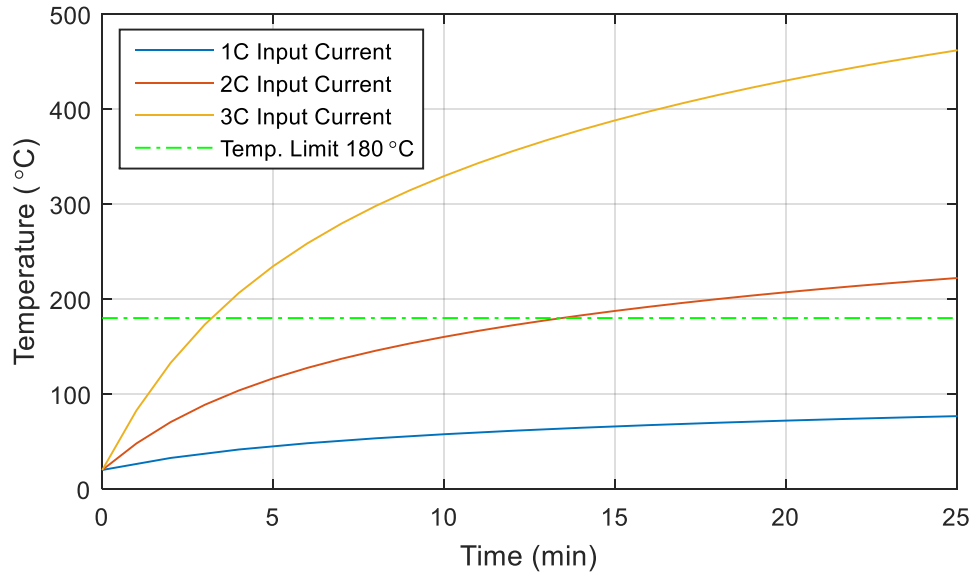


Figure 4-16 Temperature variation of the coil with various currents

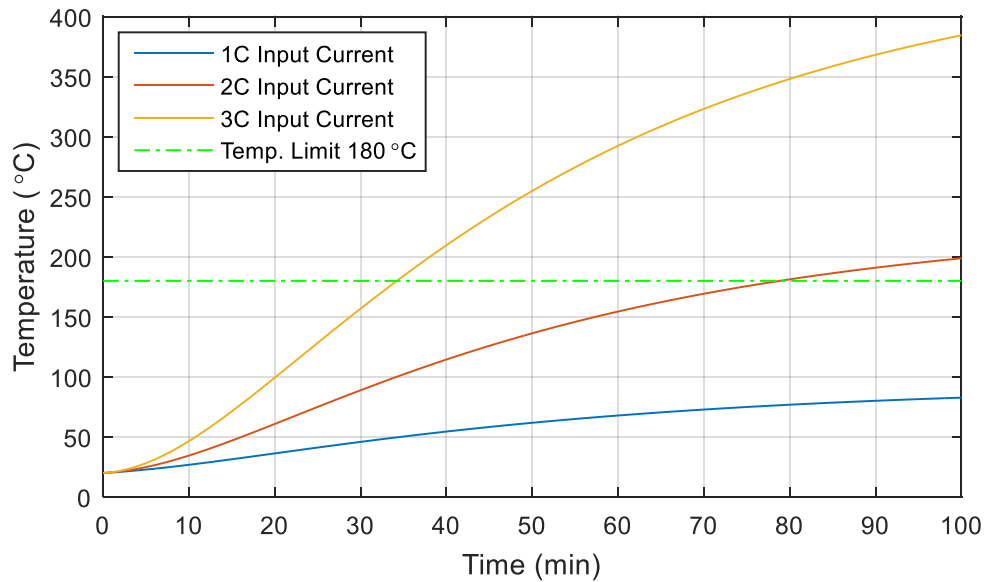


Figure 4-17 Temperature variation of the magnet with various currents

The magnet material used in the designed motor is ferrite Y30, the maximum working temperature of which is 180°C that is indicated in green dash-dot line in Figure 4-16 and Figure 4-17. Similar to the temperature variation of the coil, the temperature trend of the magnet with various input currents have similar shapes but different slopes. In the rated load condition (input current is 1 C) the magnet's temperature stays under 70°C; in double load condition (input current is 2 C) it does not reach the temperature in the 78 minutes running time, and in triple load

condition (input current is 3 C) the magnet's temperature exceeds the temperature limit in no more than 34 minutes. The consecutive operating time with various currents is determined by the weaker part, the one with a shorter time to reach the temperature limit. Therefore, considering both of the coil and the magnet, it is concluded that the consecutive operating time is 13 minutes under double load and 3 minutes under triple load. The consecutive operating time should be paid serious attention when a motor is working in an overloaded condition. Beyond the allowable time, it is possible for the motor to be permanently damaged.

4.4 Summary

The demand for high performance and low cost motors has grown from the need to accommodate devices that incorporate energy efficiency in their designs. In order to meet this demand, motors have to improve their output density and reduce their losses. The losses of ferrite IPM mainly consist of copper loss and iron loss, both of which are thoroughly analysed in the chapter. And the iron losses analysis also considers the effects of the PWM harmonics. With the consideration of PWM harmonics effect, the iron loss estimation of the studied motor becomes more accurate. The losses and efficiency maps of the studied motor are plotted from the FEA results and compared with that of IM. The proposed ferrite PM machine exhibits great superiority to the IM in terms of efficiency below the rated speed, with about 3-4% higher efficiency. All the losses are converted into heat and increase the temperature of a motor. The rising temperature will worsen the properties of PMs and is possible to wreak irreversible damage to a motor. A detailed thermal analysis is implemented to the ferrite IPM, and the consecutive operating time with over load is estimated as well.

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5 Benchmarking a Novel Ferrite Interior Permanent Magnet (IPM) Motor

5.1 Introduction

The research concerns the novel design of a ferrite two-layer interior permanent magnet (IPM) configuration for high performance applications. The proposed design improves PM flux by flux-focusing techniques and maximises the reluctance torque with a multi-layer structure. High torque density and efficiency are achieved with comparable performance to a rare-earth equivalent. For benchmarking, a standard industrial IM is used. Induction machines, especially the cage ones, are the most wide machine type in the industry due to its low cost and rigid rotor structure. Since one aim of the project is to replace the low efficiency IMs in the vast industrial and domestic market, it is perfect to benchmark a commercial IM with the same housing and stator dimensions. As the high-price rare-earth materials are excluded from the design, the cost of the ferrite PM motors is comparable with IMs. On the other hand, by using PM excitation and salient rotor structure, higher efficiency, torque density and power factor can be achieved for the proposed ferrite IPM motor. Furthermore, due to the increasing mandatory efficiency standards to be implemented, the proposed ferrite machine will become more and more competitive in the near future.

5.2 Performance Comparison with Induction Machine

For the performance evaluation of the proposed machine design, a benchmark is necessary to reveal its advantages. IMs, especially the cage ones, are the most widely used machine type in the industry due to the merits of the mature manufacturing technology, robustness, high reliability, low maintenance requirements and cost, IMs are utilized in various applications for instance pumps, fans, generator, traction motors and so on. With the developments of technology and material, the IMs show disadvantages in efficiency, power factor and power density compared with their PMSM counterparts. However, the PMSMs relies heavily on high-energy rare-earth metals [1,2]. The future supply of rare-earth materials is likely to be restricted because of the strategical nature,

and the price is prohibitive and highly unstable in the global market. As a result, IMs are still particularly attractive owing to the advantage of much lower material costs and take up at least 90% of the overall industrial installation.

However, the excitation field in IMs is set up by the magnetizing component of the stator winding current. Thus, the power factor is inevitably lower than PM machines. What is more, due to the existence of considerable copper loss in the rotor cage, the efficiency of IMs is usually inferior to their PM counterparts, which puts them in a disadvantageous position to meet the higher power saving requirements. Currently, mandatory efficiency requirement standards have been enacted in many countries including the US, EU, China and Brazil. The efficiency regulations cast the burden of designing more efficient electric machines on the manufacturers. It is of significant importance to improve efficiency, not only for business purpose but also for the mandatory requirements [3-5]. Researches have shown that the total loss can be reduced by 15-20% and the efficiency can be improved by 1-2% by using die-cast copper rotors for IMs [1,5]. But the melting point of copper is much higher aluminium, and the copper die-casting process is not yet mature and quite costly. Thus the copper die-casted IMs usually have shortened lifetime and much higher manufacturing costs, especially for the small and medium scale IMs [6].

The cross sections of the proposed ferrite two-layer IPM and commercial IM models are demonstrated in Figure 5-1, and the detailed dimensional specifications are listed in Table 5-1. The size and geometry of the stator lamination are exactly identical, and the inner and outer diameters of the rotor are also of the same size. Both machines are facilitated by short-pitched single-layer distributed windings. The IPM has 36 slots in the stator and 6 poles in the rotor with two slots per pole per phase, while the IM has 36 slots in the stator and 44 slots in the rotor with 4-pole armature windings. Since the number of poles is not the same for the two machines, the windings distributions are different from each other. The ferrite IPM is equipped with single-layer chain windings, and the IM is facilitated by single-layer cross windings. Not to lose generality, the maximum electrical loads of the stators are kept the same.

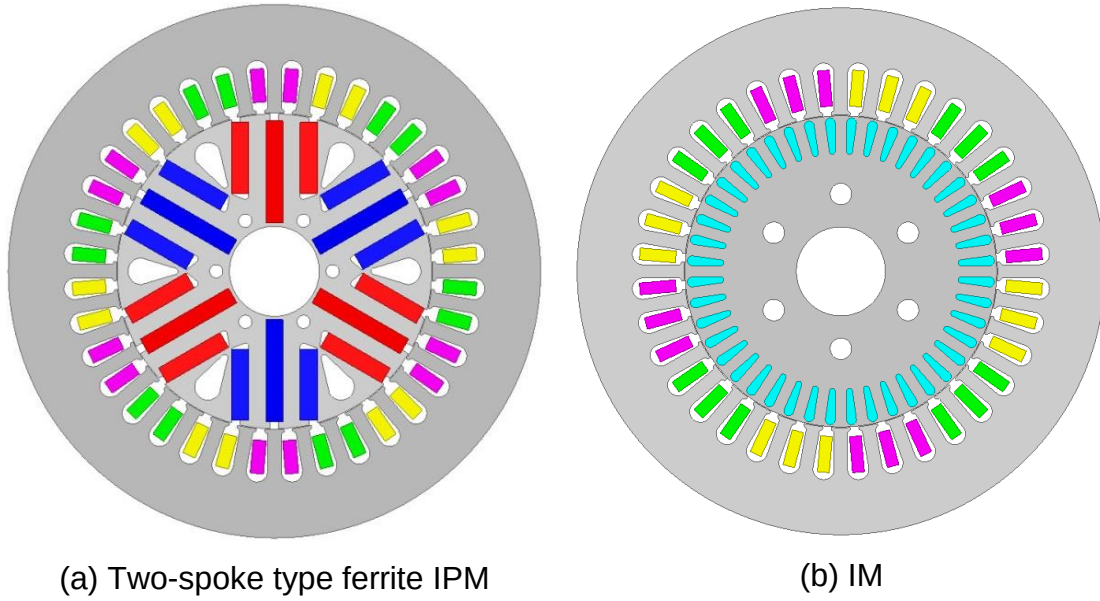


Figure 5-1 2D Models of the ferrite IPM and IM.

Table 5-1 Dimensional specifications of the ferrite IPM and IM

Parameters	PM	IM	Unit
Stator Outer Diameter	160	160	mm
Stator Inner Diameter	95	95	mm
Rotor Outer Diameter	94.5	94.5	mm
Rotor Inner Diameter	27	27	mm
Airgap Length	0.25	0.25	mm
Stacking Length	90	90	mm
Number of Stator Slots	36	36	
Number of Poles	6	4	
Number of Rotor Slots	-	44	
Number of Conductors per Slot	108	100	
Material of Rotor PM/Cage	Ferrite	Aluminum	
Material of Lamination	50W470	50W470	

5.2.1 Torque Capability

The torque profiles against armature current are illustrated in Figure 5-2 for both the IPM and IM.

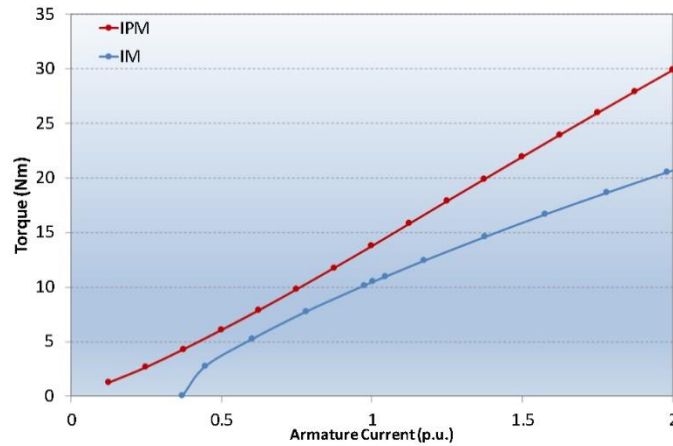


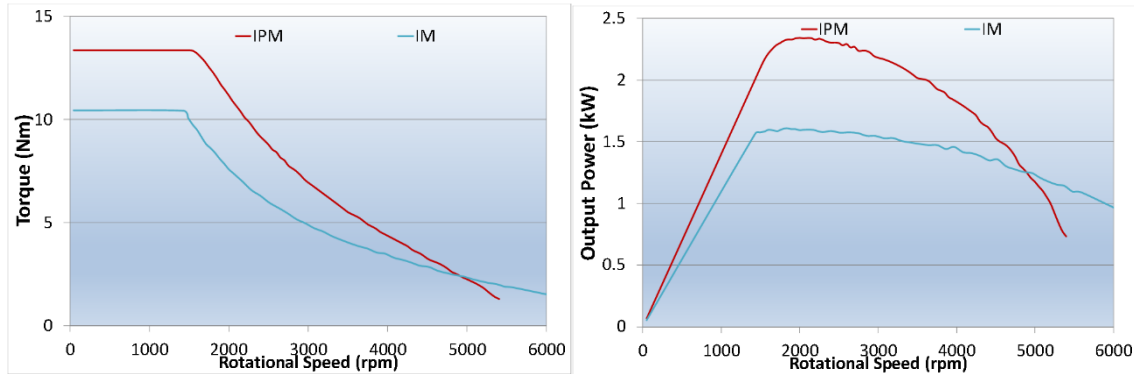
Figure 5-2 Comparison of average electromagnetic torque with armature current.

The ferrite IPM exhibits much better torque capability than its IM counterpart. Since the excitation field of the IM is generated by armature current, the current is 0.36 p.u. under no-load condition. For the IPM, the excitation field is mostly produced by the PMs under light and medium loadings, which improves the utilization of the armature current for torque production purposes. Thus, the no-load current of the IPM is very low, ideally only to overcome the no-load losses. With the increase of armature current, the increment in torque against the current is close for the two machines. As the current grows higher, though the PM excited field abates owing to the saturation, the reluctance torque increases and dominates under heavy and overloaded conditions, which would help to increase the torque ability of the IPM further. On the other hand, the current in the squirrel cage increases dramatically with the load for the IM due to larger rotor slip, which consumes part of the input power and places a negative influence on the torque increment under overload conditions. Thus, the ferrite IPM has a much better torque profile. At rated current, the torque output of the IPM is 1.27 times that of the IM, which means 27% higher torque and power density is achieved with the same electric loading. And the IPM exhibits even more remarkable advantages over the IM under overload conditions.

5.2.2 Torque-Speed and Torque-Power Curve

Figure 5-3(a) depicts the torque-speed characteristics of the IPM and IM at rated current under maximum torque per ampere control strategy. Since the

torque capability of the ferrite PM has been improved greatly, much higher torque can be delivered within the rated base speed range comparing to the IM.



(a) Torque-speed curves;

(b) Power-speed curves;

Figure 5-3 Torque-speed and Power-speed characteristics of the IPM and IM

But under flux weakening operation, the torque output of the IPM reduces faster with speed. Hence, the torque difference between the two machines becomes smaller and the torque delivered by the IM eventually exceeds that of ferrite PM when the speed is over 4500 rpm. What's more, the IM has a wider operational speed range of over 6000rpm, while the induced voltage of the ferrite PM exceeds that of power supply at the speed of 5400rpm.

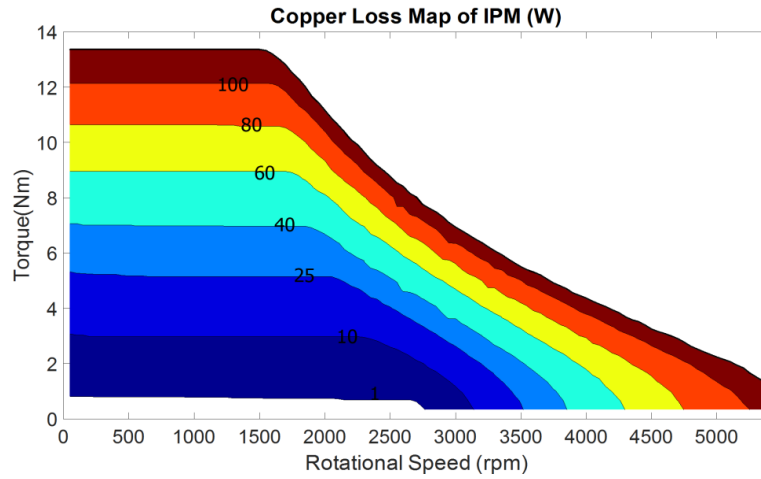
The power-speed curves of the ferrite PM and IM are illustrated in Figure 5-3(b). With rated current, the ferrite PM machine can deliver 2.1kW at rated speed and 2.3kW at 2000rpm. Then the output power reduces with speed and drops below 2.1kW when the speed exceeds 3100rpm. The IM achieves 1.5kW at rated speed and 1.6kW at 1600rpm, and its constant power speed range (CPSR) of the IM reaches 3500rpm. Due to the better flux weakening performance, the power of the IM drops much slower than the IPM. The IM delivers 1.1kW power at 5400 rpm and nearly 1kW at the speed of 6000rpm.

Because the excitation field of IMs is generated by armature current, it can be directly adjusted by controlling the reactive current. Thus better flux weakening performances and wider operational speed range can be achieved for the IM. For PM machines, the exciting field is mainly provided by PMs, which is not adjustable since the PMs are pre-magnetized. To operate over the rated base speed, the

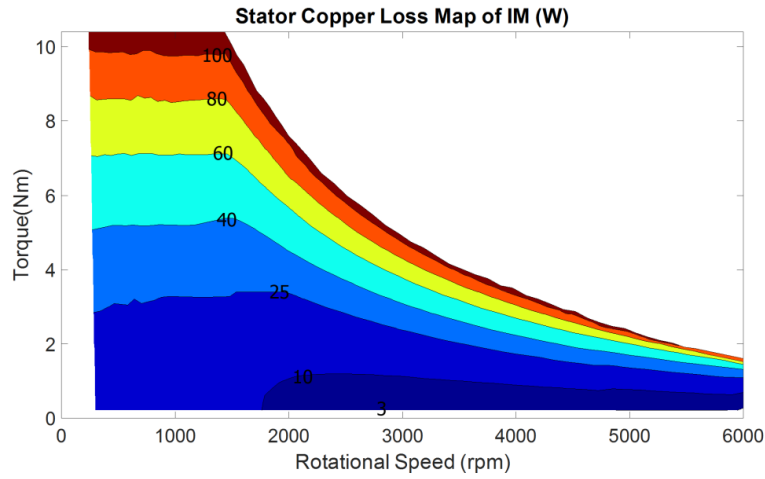
flux weakening control strategy is applied to generate armature reactive field opposing the PM magnetization for weakened excitation. The flux weakening performance depends on the ratio of the product of current amplitude I_m and d-axis inductance L_d over PM flux linkage Ψ_f . According to d-axis magnetizing inductance derivation in subsection 3.2.1.2, L_d is greatly influenced by the width of PMs and reduces with the width. Besides, wider PMs will enhance PM flux field and thus Ψ_f . Hence, the flux weakening performance deteriorates dramatically with the increase of PM width. However, the design of spoke type machines requires a certain PM width for sufficient PM excitation and rotor saliency to achieved high torque density, which inevitably leads to relatively small L_d and inferior flux weakening ability.

5.2.3 Losses

In electric machines, the estimation of the losses is of great importance concerning the thermal limitation and energy conversion efficiency. Because the losses would eventually be transferred into heat and result in a temperature rise of the machines, the maximum continuous loading abilities and lifespan of the machines are significantly influenced. The losses in electric machines can be usually divided into four categories: copper loss in stator and rotor windings, core loss in stator and rotor lamination cores, eddy current loss in back iron and PMs, and mechanical losses. Since eddy current loss is completely eliminated from the ferrite PMs and does not exist in the IM, it is not included in this section. Additionally, the mechanical loss due to the bearing friction and windage is dependent upon rotational speed but independent on the load. It is difficult to estimate by a direct calculation but can be approximated empirically as 2% that of rated power under full load condition, and proportional to the rotating speed square [7]. In this study, the total loss is dominated by electrical and electromagnetic losses. As a result, only copper loss and core loss are investigated and compared in detail for the ferrite PM and IM.



(a) Ferrite IPM;



(b) IM;

Figure 5-4 Stator copper loss maps of IPM and IM.

5.2.3.1 Copper Loss

The copper loss usually takes up a large part of the total loss. The copper loss can be derived from the current flowing in the armature winding and the estimated winding resistance, and hence is also called I^2R loss. As the basic condition for the performance comparisons between the ferrite PM and IM, the stator copper losses are set to be equal at rated operational condition, so that the electrical loading and thermal condition of the stators are the same.

Since torque can be approximately regarded as proportional to current below rated speed, copper loss of stator windings increases with torque and is

independent on the speed for both the IPM and IM, as shown in Figure 5-4. As a precondition set for the comparisons, the maximum copper loss in the stator is the same for both models. But because of the different operating principles due to different machine types, their copper loss features change differently under various operational conditions. When below the rated speed and torque, the ferrite PM machine achieves lower copper loss level because of the advantage of PM excitation, while high copper loss is observed for the IM as excitation current is required to establish the flux field. The difference is more obvious at light or no-load conditions when the copper loss is close to zero for the ferrite PM and over 15W for the IM.

However, the result is completely different when it comes to high speed range. The decrease in copper loss is observed with speed under no-load condition since the excitation current is reduced during flux weakening control. But under loaded condition, torque generating current increases to maintain the same torque, and thus copper loss still increases with speed. On the other hand, much higher stator copper loss is observed for the IPM as the larger d-axis current is necessary to achieve flux weakening operation. Accordingly, the copper loss increases dramatically with speed.

For the IM, there is a considerable amount of copper loss in the rotor side due to the existence of a squirrel cage. Since the current in the rotor cage is induced from the stator reactive field, the copper loss increases with stator current and rotating speed, as depicted in Figure 5-5. The rotor copper loss of the IM can reach up to half of stator copper loss at rated torque output, which means 50% higher overall copper loss for the IM compared with the ferrite IPM. Considering only the stator copper loss, those of the IM and IPM are comparable below rated speed. But with the rotor copper loss taken into consideration, the overall copper loss of the IM is much higher than that of ferrite PM when the speed is lower than 3000 rpm. Nevertheless, under very high speed operations, the copper loss of the ferrite PM machine is still higher because the extremely large d-axis current is required for the deep flux weakening control, as demonstrated in Figure 5-6.

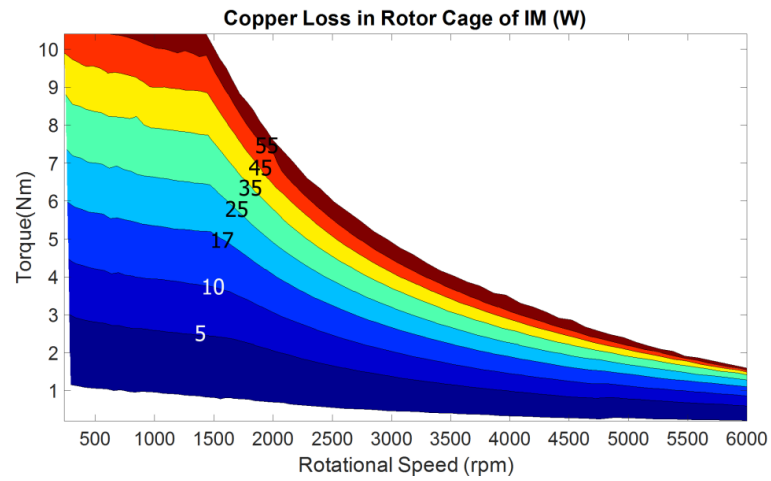


Figure 5-5 Copper loss in the rotor cage of the IM.

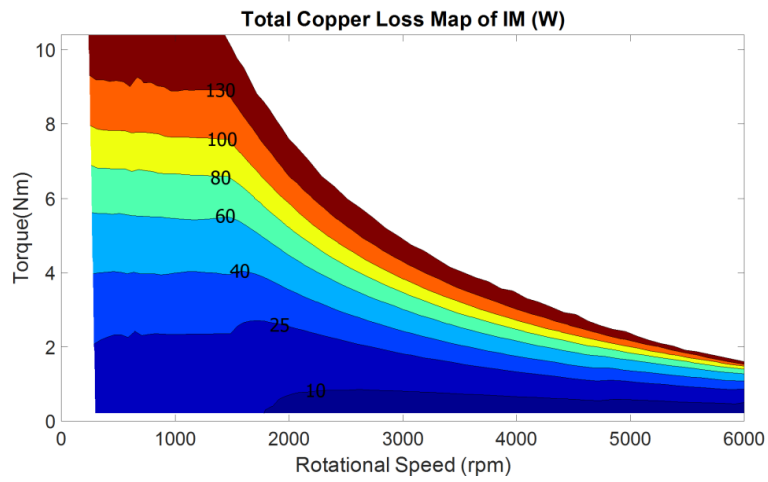


Figure 5-6 Total copper loss of the IM.

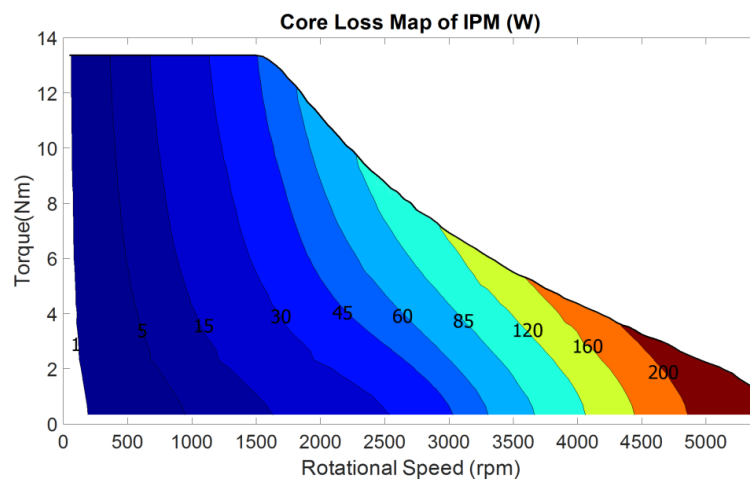


Figure 5-7 Core loss map of the IPM.

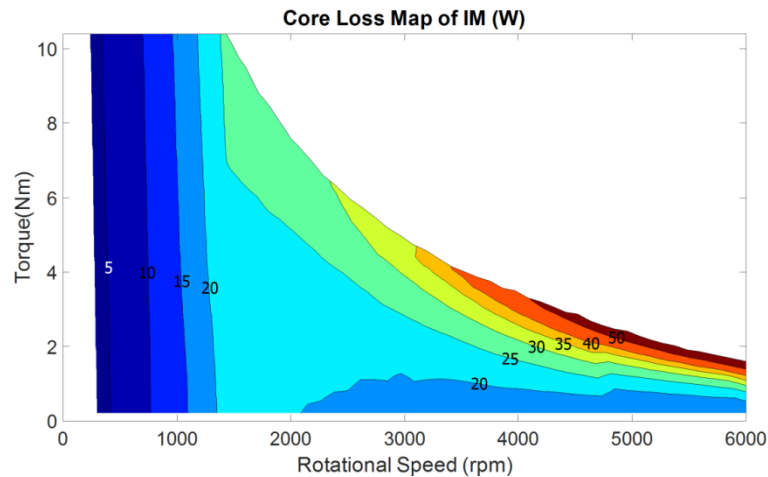


Figure 5-8 Core loss map of the IM.

5.2.3.2 Core Loss

Core loss, which consists of hysteresis loss, eddy current loss and excessive loss in the lamination cores, is another major source of heat, especially at high speed operations. Core loss is significantly influenced by the operating frequency, and hence, the losses increase with rotational speed exponentially for both the ferrite PM and IM, as demonstrated in Figure 5-7 and Figure 5-8. Moreover, the flux density also has a considerable impact so that the armature current could aggrandise the core loss. Since the PMs largely decide the excitation field for the IPM, the rotational speed has a major influence on the core loss over the whole operational range. The armature current shows the minor effect, and core loss rises as the increase of loading due to the stronger armature reaction. As for the IM, the flux field is almost constant when below the rated speed, because excitation is controlled by armature current. As a result, core loss can be approximately regarded as only dependent on rotational speed. But under flux weakening operations, the excitation current decreases and lower core loss can be observed under no-load condition. But as the increase of output, core loss increases significantly due to the influence of the reactive field at high speed.

The core losses of both machines are quite small under rated speed. Compared with the IM, the core loss of the IPM is slightly higher because of higher flux density owing to the PM excitation and higher electrical frequency due to the higher number of poles. The core loss of ferrite PM becomes much higher than

IM when the speed is over nominal value. As a result of better flux weakening ability for the IM, much lower flux density can be achieved as the speed grows, which results in a much slower rise in core loss. On the other hand, the flux density of the ferrite PM stays at a relatively high level owing to the fixed PM excitation and inferior flux weakening ability, which leads to an exponential increase in core loss with speed. This is certainly one of the factors that deteriorate the performance of the IPM at the high speed range.

5.2.4 Efficiency

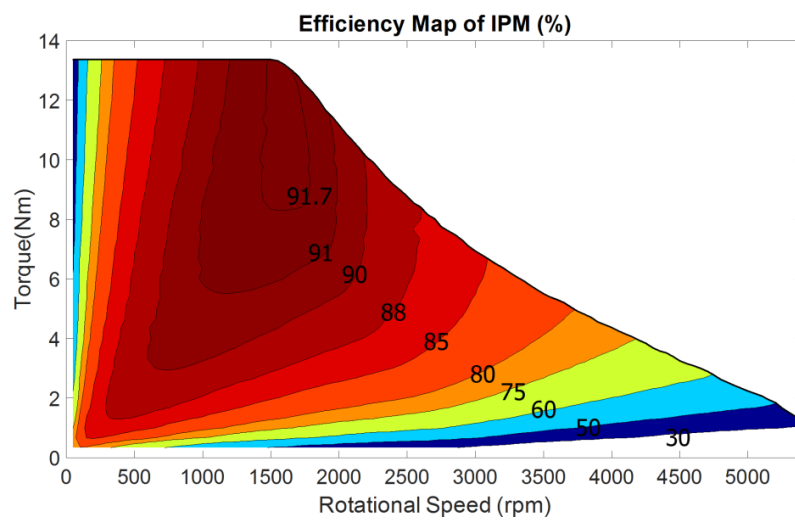


Figure 5-9 Efficiency map of the IPM.

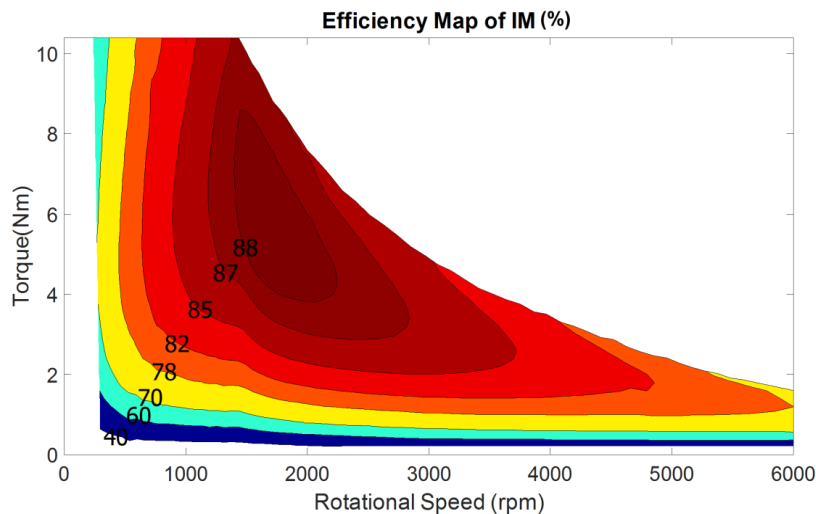


Figure 5-10 Efficiency map of the IM.

By compiling the losses calculated in the previous sections, efficiency maps of the proposed ferrite PM and the commercial IM are derived, as shown in Figure 5-9 and Figure 5-10. Due to the different loss features under different torque and speed conditions, the whole speed-torque operation region is divided into four sub-regions for detailed comparisons and discussions.

- Sub-Region *I*: Low Speed and Low Torque

In this sub-region, the load torque is less than 40% of rated and the rotating speed is lower than 1000 rpm. As the speed is relatively low, the core loss for the two machines is similar. The difference in behaviour mainly lies in a copper loss. PMs establish the main flux field of the IPM and thus the armature current is much lower, while a large part of the armature current is for the build-up of field excitation. As a result, the efficiency of the IPM is much better than the IM for lower copper loss in the region.

- Sub-Region *II*: Low Speed and High Torque

Since the current is high under these circumstances, copper loss dominates the overall loss. Although the maximum stator electric loadings are kept the same for both the machines, extra copper loss in the rotor bar can add up to 25% of the total loss for the IM. What's more, higher torque per ampere can be achieved for the IPM, which means higher power output. Combining the two factors, much higher efficiency is expected for the IPM in this sub-region.

- Sub-Region *III*: Medium Speed and Medium to High Torque

Normally the most frequent operational ranges for a motor lie in this region, and both of the motors are designed to exhibit the highest efficiency under the torque-speed range of 1000 to 2000 rpm speed and a half to the full load torque. Apparently, the ferrite IPM shows better performance in efficiency against its IM counterpart. At rated working points, the proposed IPM machine operates with an efficiency of over 91.7%, against that of 87.8% for the IM. Within this sub-region, the IPM achieves over 90% of efficiency with the highest of 91.9%, while the IM obtains over 85% with the highest efficiency of 88.7%. And 2-4% higher efficiency can be attained compared with the IM, which makes a big difference for energy saving because the motors work in this region for most of the time in operation.

- Sub-Region IV: High Speed

The motors enter deep flux weakening operations in this sub-region. For the IPM, a negative d-axis current is necessary to weaken the PM exciting field to reduce the induced voltage during high speed operations. The increased demagnetising current will not only increase the copper loss but also intensify the spatial harmonic of the armature MMF resulting in higher core loss. Thus, the efficiency of the IPM decreases significantly with speed. On the other hand, the excitation current of the IM can be reduced for weaker flux field, and both copper loss and core loss are much lower than the IPM. Therefore, the efficiency of the IM exceeds that of IPM when the speed is over 3000rpm and becomes far more superior over 4000rpm.

To conclude, the IPM achieves higher efficiency in most of the operational region, and most importantly, in the most frequently operating sub-region *III*. Although the IM obtains higher efficiency during the high speed range, it is not the main working region for most of the small-scale industrial applications. On this basis, the IPM exhibits much higher overall efficiency.

5.2.5 Power Factor

The Power factor is an important index for electric machines, especially in inverter-fed variable speed systems. High power factor can reduce the power rating of the inverter and thus reduces the cost for power electronics since lower apparent power and current can be achieved.

The IPM achieves a relatively high power factor from low to medium speed range owing to PM excitation, as illustrated in Figure 5-11. Its power factor decreases slightly with the increase of torque load when below rated speed because higher current generates higher armature reactive field, which increases the leading phase angle of the armature voltage vector. When the speed is over the rated, a larger current advanced angle is required for flux weakening, and the power factor increases as the phase angle difference between winding voltage and current reduce. When entering the high speed sub-region, an even larger current angle

is necessary. This makes power factor change from lagging to leading, and decreases dramatically with speed.

The power factor map of the IM is depicted in Figure 5-12. Due to the inherent disadvantage of the IM, it has a much lower power factor when below rated speed. The power factor is extremely low under no-load or light load because of its indispensable magnetizing current, even though the active power is very low. With the increase of loading, the active power rises and so does the power factor. The IM only shows its advantage over the IPM during the high speed range, when the excitation current is reduced.

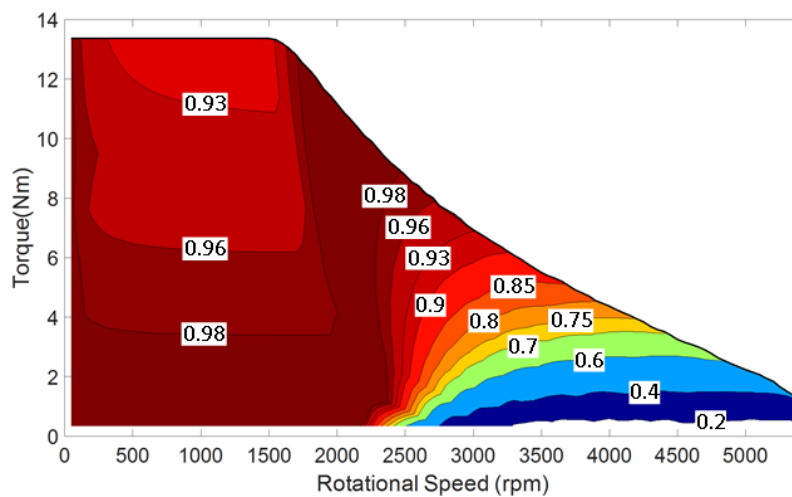


Figure 5-11 Power factor of the IPM.

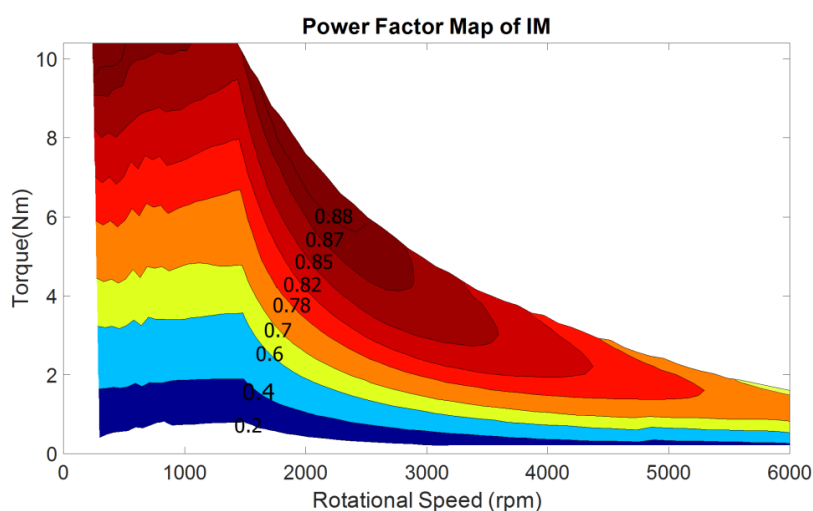


Figure 5-12 Power factor of the IM.

5.2.6 Material Costs

The economic performance and weight of the motor are also very important from the market perspective. The main materials of electric machines include silicon steel lamination, copper wires, ferrite PMs, and aluminium. The weight and material cost of the ferrite PM and IM are listed in Table 5-2. The total weight of the two motors is approximately the same since the same stator lamination, shaft, rotor dimension and housing assembly are used. The IM has slightly heavier rotor lamination as the IPM motor uses a large amount of ferrite material in the rotor. And since the mass density of ferrite is lower than silicon steel, the overall weight of the IPM is marginally smaller. The material cost estimation is based on the UK market price in [8]. The costs of motor cores and stator windings are almost the same due to similar material usage. The difference is reflected in the rotor. The IM rotor contains only die-casted aluminium bars besides lamination core, making IM rotor one of the cheapest structures in electric machines. Although highly expensive rare-earth PM materials are excluded for the IPM, the considerable amount ferrite used still costs more than the IM rotor. Therefore, the overall cost of the PM motor is 15% higher. However, the ferrite IPM machine can obviously deliver a much higher torque density. If the power output is taken into consideration, the cost per kilowatt of the IPM is about 18% lower than the IM, which makes the two-layer ferrite IPM much more attractive for various applications.

Table 5-2 Material costs of the ferrite PM and IM

	IPM	IM
Lamination Mass (Kg)	10.0	10.7
Copper Mass (Kg)	2.1	2.1
PM/Aluminum Mass (Kg)	1.0	0.5
Total Mass (Kg)	13.1	13.2
Lamination Cost (£)	20.0	21.4
Copper Cost (£)	14.7	14.7
PM/Aluminum Cost (£)	8.0	2.0
Total Cost (£)	42.7	37.1
Cost per Kilowatt (£/kW)	20.3	24.7

5.3 Conclusion

Comprehensive comparisons of the performances between the proposed ferrite PM machine and a commercial IM are carried out at first. The same stator lamination geometry and housing assembly are used for both of the two machines. Despite the winding arrangements are different, the amount of copper wires and stator copper loss at rated conditions are kept the same to set the basis for performance comparisons. The proposed ferrite PM machine exhibits great superiority to the IM in terms of torque capability, power density, efficiency and power factor below the rated speed. It is noteworthy that ferrite PM achieves 38% higher torque density, 3-4% higher efficiency as well as much higher power factor. Even when the operational speed reaches twice the rated at 3000 rpm, the ferrite PM still holds some advantages over the IM in the overall performances. The only disadvantage of the proposed ferrite PM to the IM is the poorer flux weakening ability owing to a large amount of PM materials used. The cost of materials is estimated based on the market price, and the cost of the ferrite machine can be regarded as similar to that of IMs since no expensive material is used.

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6 EXPERIMENTAL VALIDATION

A prototype ferrite IPM machine is manufactured according to the motor design in Chapter 3. In order to obtain the characteristics of the prototype machine, a series of tests is performed under various working conditions. The proposed machine, as well as the rotor of IM, are validated by the comprehensive comparisons of the performances between FEA and experimental results. The detailed discussion and analysis are shown in the following sections.

6.1 Validation of the Motor Design

6.1.1 Manufacture of Prototype Motor

Some major steps of the prototype motor manufacture are demonstrated in Figure 6-1. Due to the brittle physical nature of ferrite materials, there is a risk of chipping and breaking the magnets when stress is exerted on them. For the convenience of magnet manufacture and rotor assembly, each of the PM poles is divided into two parts in an axial direction and then inserted inside the rotor from a different end of the rotor lamination. The required cuboid ferrite magnets are shaped and magnetised first, the primary and secondary magnet pieces shown in Figure 6-1(a). After being fixed on the rotor shaft and inserted with the magnet pieces, the rotor lamination is clipped and machined to the final outer circumference, shown in Figure 6-1(b). Finally, two pieces of aluminium end-plates are used at each end of the rotor lamination to increase the robustness of the rotor, and two bearings are installed at each end of the rotor shaft. The completed rotor is shown in Figure 6-1(c). The housing and stator lamination are directly adopted from a commercial IM. In order to adapt the design of the ferrite IPM motor, the stator coils are re-winded with four strands of 0.6mm diameter copper wires. Because thinner copper wires and chain coil configuration with shorter end-winding are employed for the IPM, there are eight more conductors per stator slot than the original IM. The new stator windings are shown in Figure 6-1(d). and the completed stator, including the cover, is shown in Figure 6 1(e).

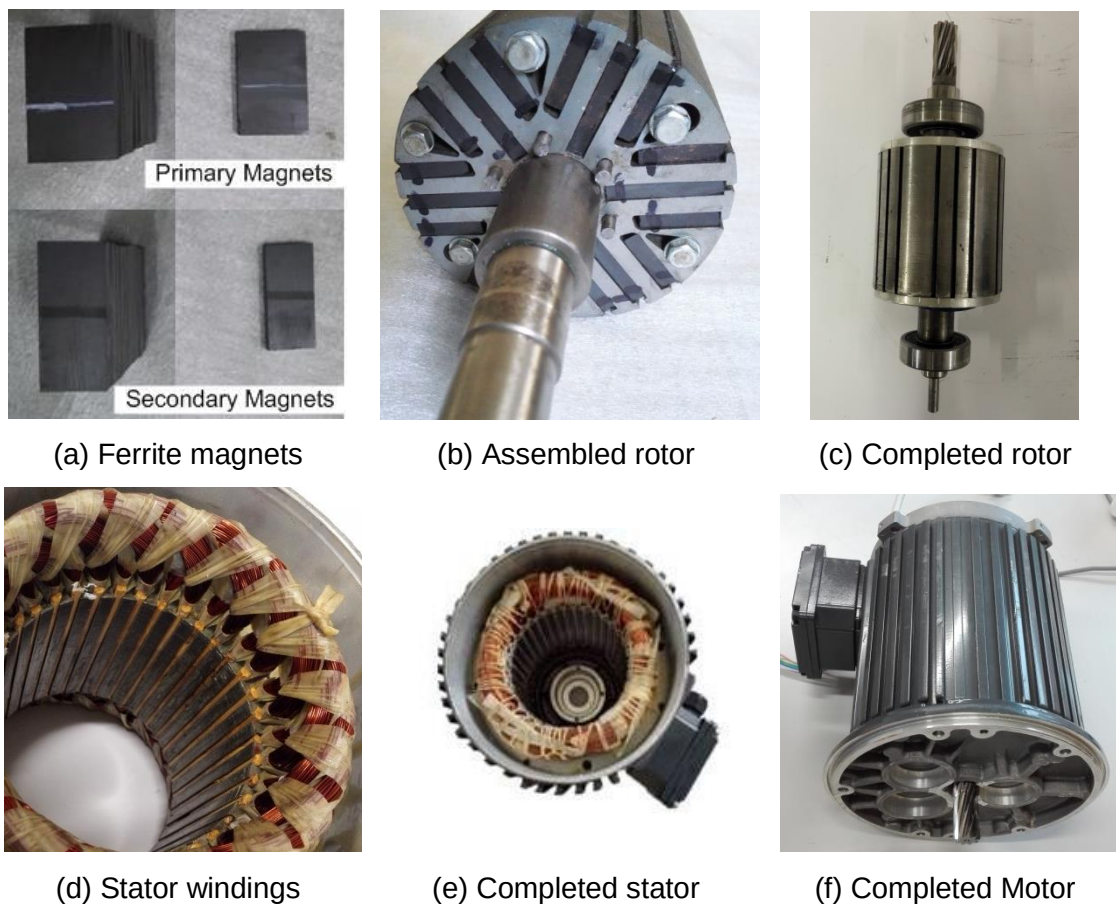


Figure 6-1 Manufacture of prototype motor

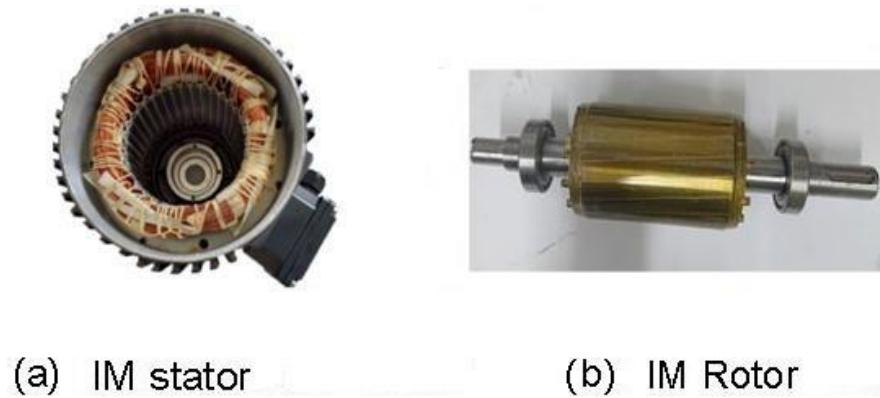


Figure 6-2 IM motor

At last, all the prototype components are carefully assembled together to make the final motor, shown in Figure 6-1(f). The IM stator and rotor construction can be seen in Figure 6-2 and as mentioned previously both machines have the same

stator. In the following several sections, a series of experimental tests will be carried out on the motors to validate the design of the machine and FEA results.

6.1.2 Open-Circuit Test

Open Circuit Test is performed on the prototype machine to obtain the back EMF of the proposed ferrite IPM machine. The machine is driven at a constant speed and a voltage acquisition device measures the open terminal voltage. The rig for the back EMF test is illustrated in Figure 6-3. An IM that is controlled by a variable frequency converter works as a primary motor and drives the prototype motor through a coupling. Since the IM is primarily used for low speed applications, only 300 rpm can be achieved for the back EMF testing. Motor

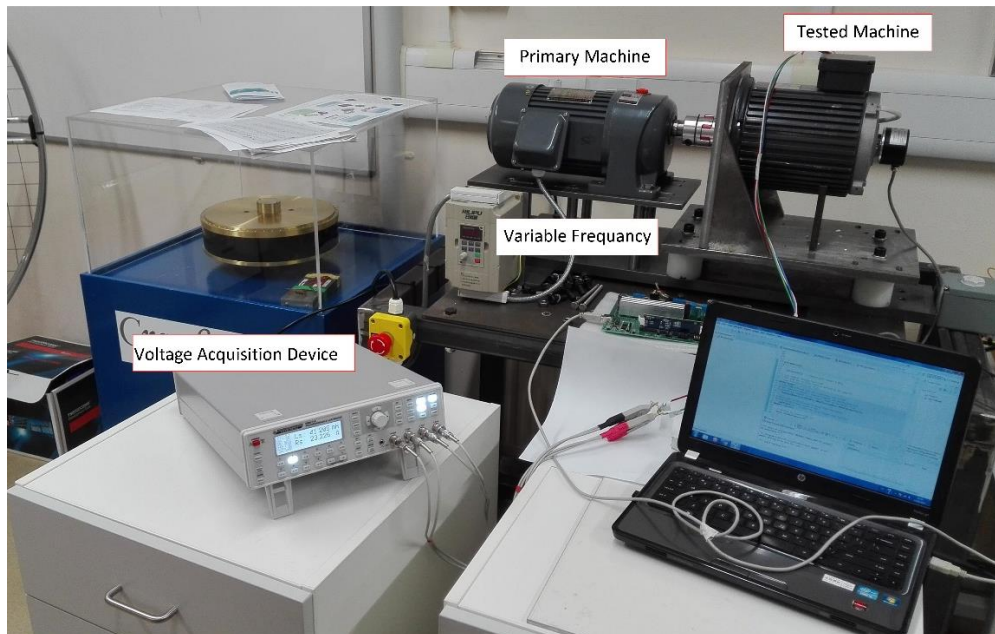


Figure 6-3 Back EMF test rig

speed is measured by an incremental encoder and fed back to monitoring computer by a speed feedback controller. An oscilloscope or voltage acquisition device obtain the back EMF of the prototype motor, and their waveforms and spectra are depicted in Figure 6-4. The fundamental component of the back EMF is about 20 V, and it also consists of some noticeable harmonics, such as the 3rd and 11th order. That is mainly caused by the special structure of the IPM motor that has a strong cogging torque, and it is also affected by demagnetisation.

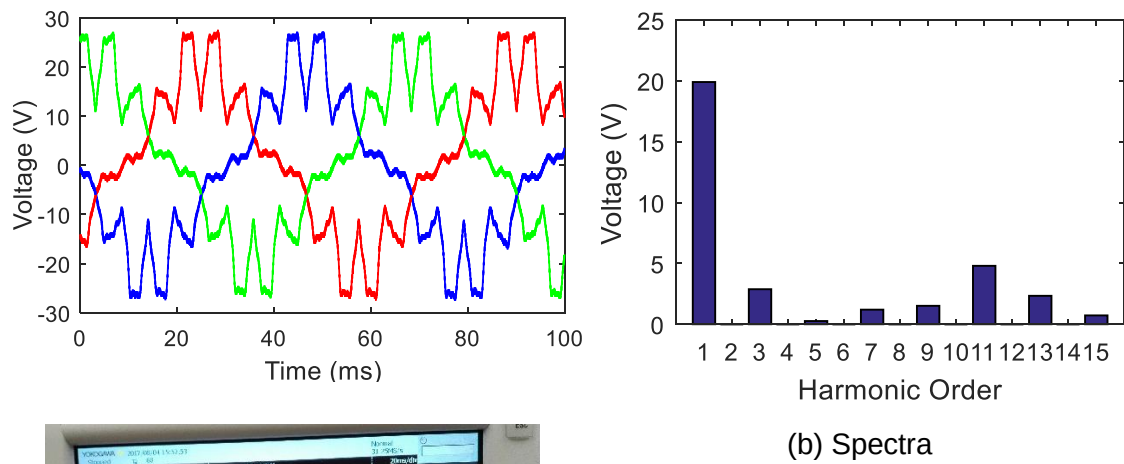


Figure 6-4 Waveforms and spectra of the prototype machine's back EMF

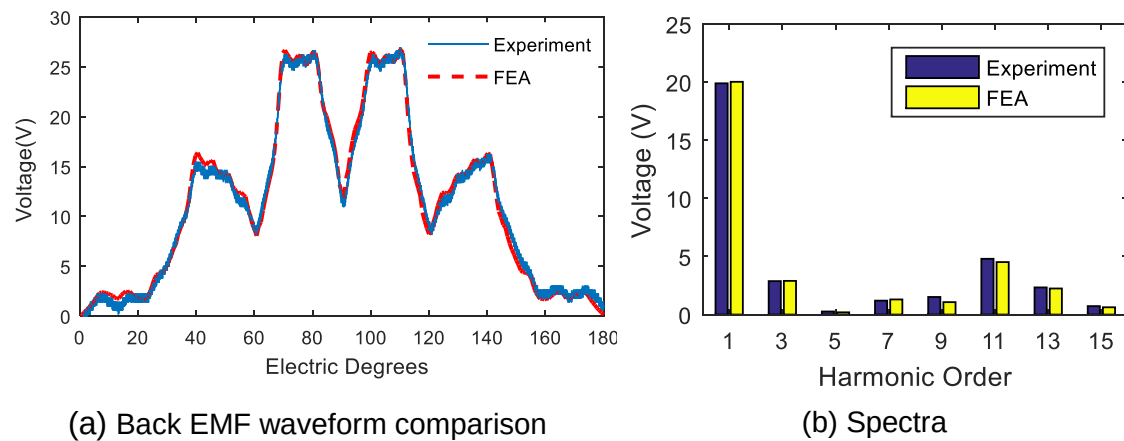


Figure 6-5 Back EMF comparisons between FEA and experiment

In order to validate the FEA simulation results in Chapter 3, the back EMF of which is compared with that of experiments. Figure 6-5 shows the comparison of back EMF waveforms and their spectra between FEA and experiment results. It is seen that FEA results have good agreement with tested results, and the slight difference may result from manufacture errors and material inconsistency.

In conclusion, the open-circuit test validates the FEA simulation results and they have a good agreement. And the noticeable harmonics should be aware when carrying out control algorithm design or motor test.

6.1.3 Load Test

After the open-circuit test, the prototype and IM motors are transported to motor performance test rig, which is shown in Figure 6-6.

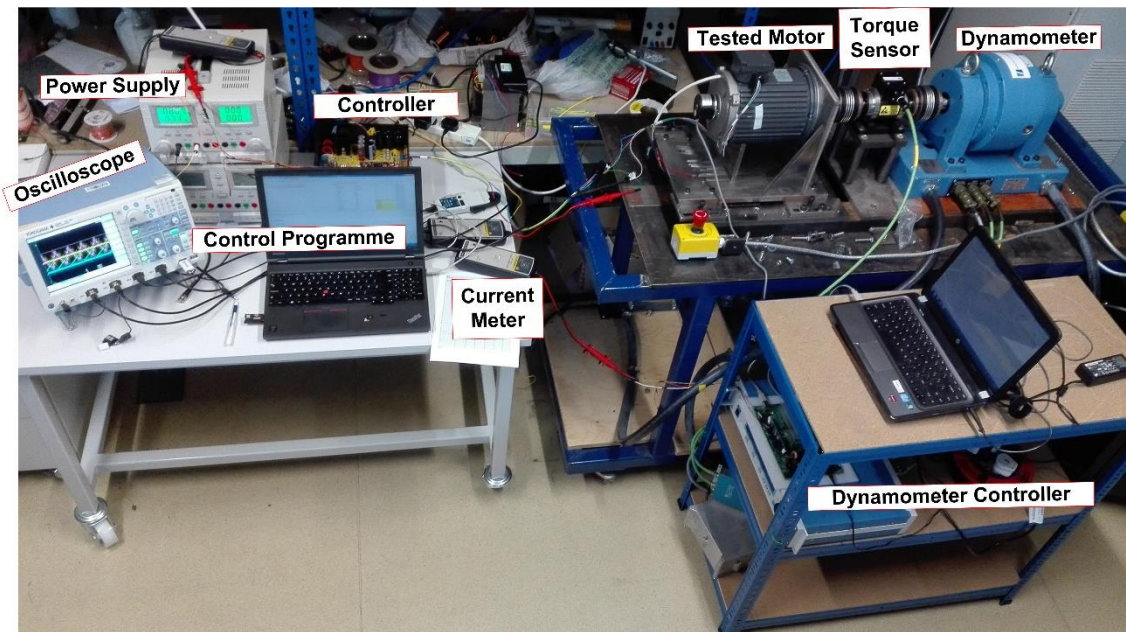


Figure 6-6 Setup of the motors test rig

In this test rig, the prototype motor is powered by TI HVDMC Kit. The output shaft of the motors is coupled with a torque sensor that is connected to a dynamometer, so the actual speed and output torque are measured by them and fed back to the dynamometer controller. The output power of the test motor is absorbed by a dynamometer and transferred into heat, and the dynamometer is cooled down by flowing water that circulates from a water tank. An instruction of the programmable dynamometer system is attached in Appendix D and Figure 6-7 shows the schematic diagram the testing configuration. During the motor test, the online current and voltage are measured by the current meter and voltage meter

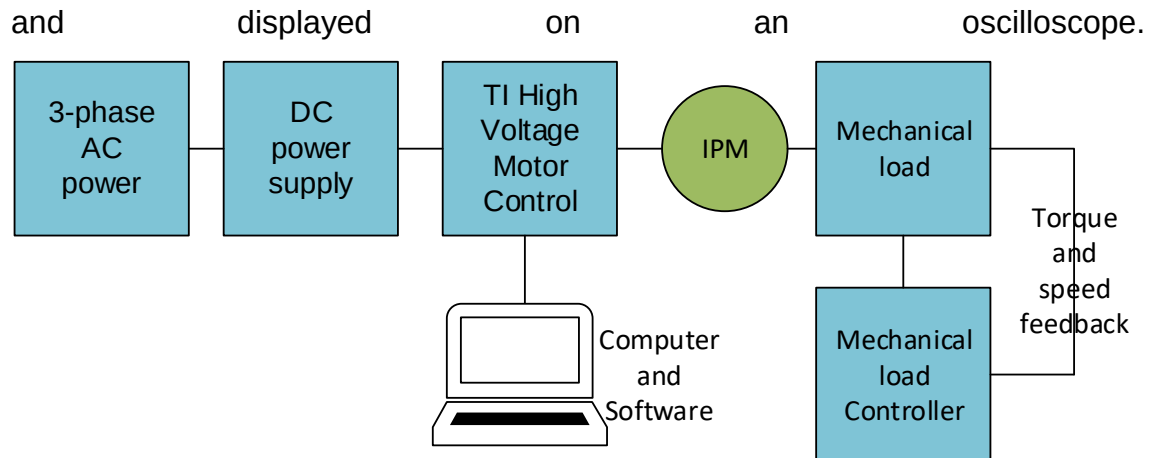


Figure 6-7 a schematic diagram of the testing system

The rotor shaft is exerted different loads by the programmable dynamometer to explore torque capacity and efficiency of the motor prototype. More detailed data and discussion are described in the following sections.

6.1.3.1 Torque Characteristics

Figure 6-8 demonstrates the comparisons of tested torque-current profiles of the IPM and IM. Due to the restrains of the power module of TI HVDMC Kit, the rated torque is not tested as the maximum current available is 0.87 p.u. for the IPM. However, the trend of the torque-current curves of both IPM and IM agree fairly well with FEA result, and the tested torque is about 0.3-0.4Nm lower than the FEA simulations. Some possible factors might cause a difference. There might be manufacturing imperfections, the actual property of materials might be marginally different from the database, and the mechanical loss could be slightly higher than estimated due to the test rig connections. The differences are only minor, and it can be confirmed that the experimental tests validate the FEA results.

By comparing the torque performances between the IPM and IM, it should be noted that the IPM is able to deliver much higher torque under the same p.u. current. Especially under light load conditions, the advantage is tremendous. Even under heavy loading conditions, the torque density of IPM is over one third higher than the IM.

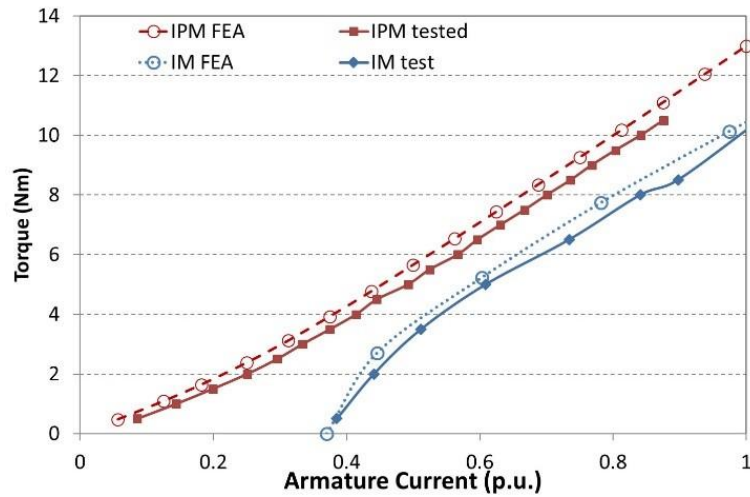


Figure 6-8 Torque output versus current amplitude in (p.u)

Figure 6-9 depicts the phase voltage and current under 3 and 6Nm at 200 rpm with the MTPA method. As the fundamental current component increases with the load, the current waveforms are much more sinusoidal than no-load, which indicates low harmonic current content. Because the motor is fed by the PWM controller inverter, 8k Hz filter is applied to the phase voltage for a clear view.

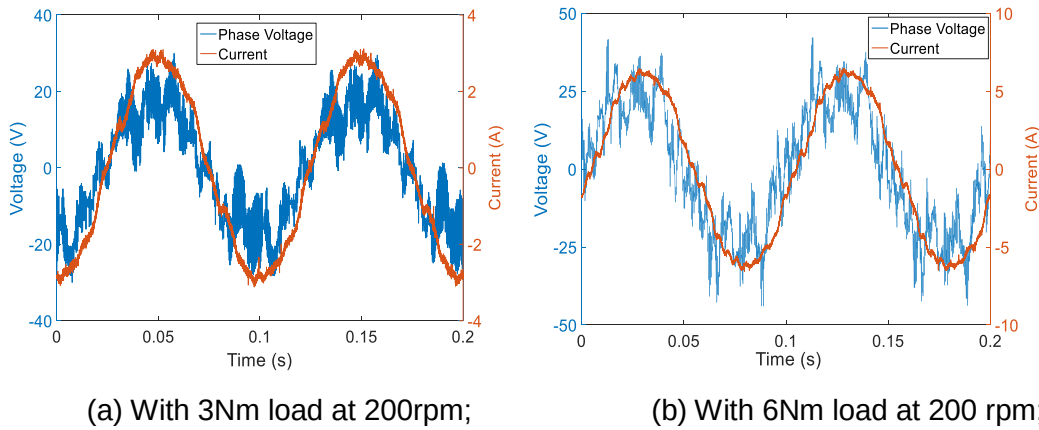


Figure 6-9 Phase voltage and current under loaded conditions.

As shown in Figure 6-9(a), the phase angles of voltage and current are almost aligned, indicating high power factor under low load conditions. With the increase of output, a slightly larger phase shift between voltage and current can be observed in Figure 6-9(b), owing to higher armature reaction and larger reluctance torque component.

6.1.3.2 Efficiency

Due to the power limitation of the controller available, the prototype machine is tested from 1–8 Nm with a maximum speed of 1500 rpm for power and efficiency validation.

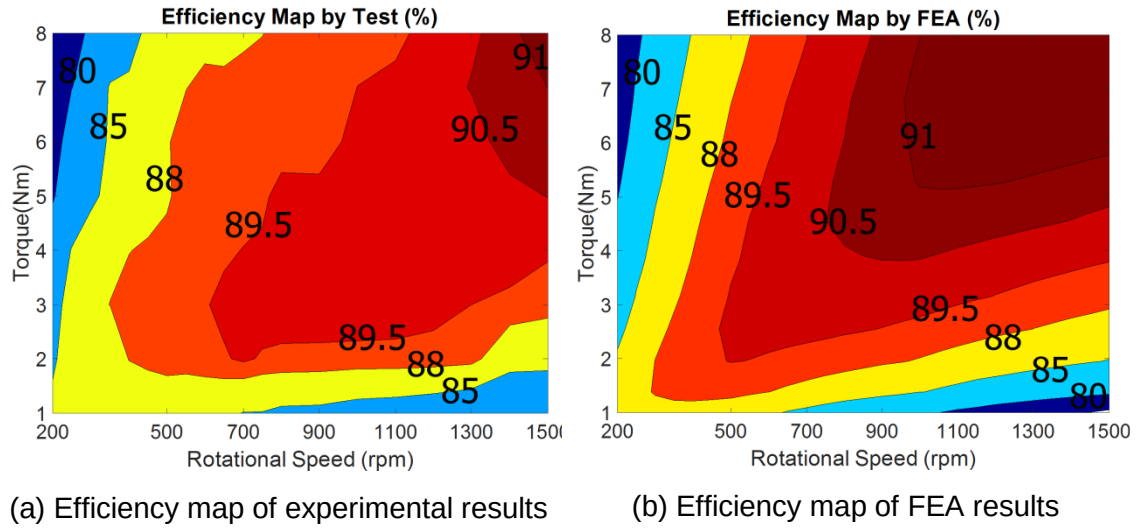


Figure 6-10 IPM Efficiency maps of experimental and FEA results

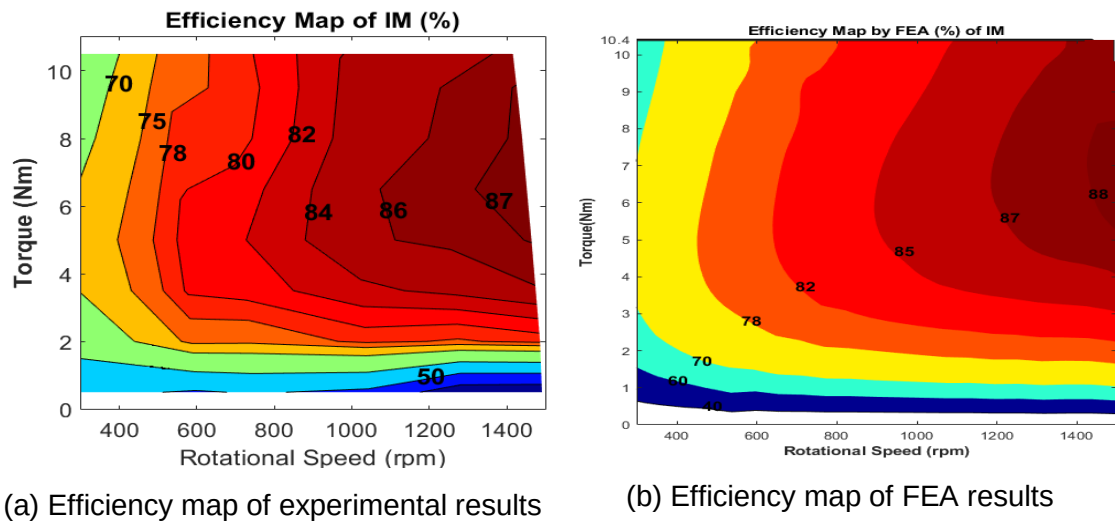


Figure 6-11 IM Efficiency maps of experimental and FEA results

The efficiency and Power factor maps of the two machines are obtained and compared in Figure 6-10 and Figure 6-11. The results agree with FEA analysis in the previous section. Higher efficiency can be achieved around rated speed under the high torque region for the IPM, while the IM exhibits higher efficiency under medium torque at rated speed. The IPM shows much higher efficiency throughout

the tested operational range and exceeds the IM by 4-5% at the highest efficiency region. If all the induction motors for industrial use are replaced this low-cost ferrite machine, the power saving is enormous. As demonstrated in Figure 6-10, the efficiency of the prototype machine is 0.5–1.5% lower than FEA at relatively high speed and torque operating conditions, and slightly high efficiency is obtained for the experiment when the loading torque is small. Two possible reasons would contribute to the difference. To simplify the FEA simulation, the armature resistance is calculated at 80°C and set as a constant. But under light load conditions, the temperature rise is very low. Thus, the resistance for simulation is actually higher than the actual value, resulting in higher copper loss and lower efficiency for the FEA. What is more, since it is too time-consuming to consider PWM signal in FEA simulation, the pure sinusoidal current is injected to the windings instead. As a result, the core loss in practical should be higher and the efficiency is lower since the harmonics caused by PWM control is ignored in simulation. Under high speed high torque operation when core loss is serious, larger differences is seen in efficiency.

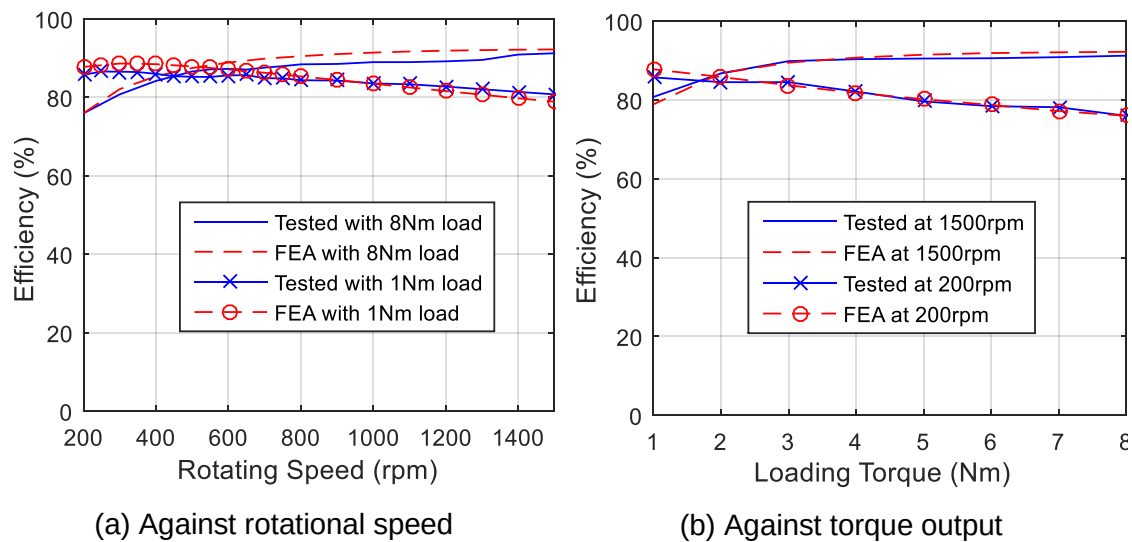


Figure 6-12 Efficiency comparisons between the tested and FEA results of IPM

The fact that the efficiency by test and FEA agrees with each other better under low speed high torque conditions can also be explained based on the theory. Because the core loss is very low and the armature temperature in simulation is

closer to that in actual situation within this operational range, the working conditions are quite close to the actual testing environment, and good accuracy is achieved. Figure 6-12(a) depicts the efficiency of experimental and FEA results against rotational speed. Efficiency increases with speed at higher torque loading while decreases during light load. Because copper loss contributes more to the overall losses at high load and can be regarded as constant under such constant torque conditions, the efficiency increases at a higher speed when power is increased. On the other hand, core loss dominates at light load and is aggrandized more significantly with speed, resulting in a decrease in efficiency.

Figure 6-12(b) shows that under constant speed conditions, core loss is far less sensitive to armature current compared to copper loss, and the mechanical loss is constant. On the other hand, the copper loss is proportional to the current square and increases exponentially with the load. Thus, at low speed, when the copper loss takes up contributes to most the overall losses, the motor efficiency decreases with load due to the higher loss increment. However, the efficiency increases with load at rated speed when core loss and mechanical loss take up a considerable portion of the total loss whose increment is less significant.

6.1.3.3 Power Factor

The power factor maps of tested and FEA result of the prototype machine and IM are illustrated in Figure 6-13 and Figure 6-14. The power factors of the two motor vary in the opposite direction against loading as shown. The power factor of IPM decreases gradually with torque, as the armature reaction increases the phase shift between voltage and current vectors. What is more, a larger portion of reluctance torque is harnessed at higher torque demands, and hence bigger advanced current angle is necessary for maximum torque production. On the contrary, the reactive current is almost constant below rated speed for the IM, and the active current increase with torque and the power factor increase dramatically when higher torque is required. But over the whole tested operating range, the IPM shows much higher power factor, 7-8% higher under heavy

the motor, so the resultant temperature can also indirectly validate the amount of losses. The temperature in a steady and transient state has been simulated by FEA simulation in chapter 4, and in this section it will be validated by experiment results.

6.2.1 Thermal Experiment Setup

A wired thermal meter (shown in **Figure 6-15**) and infrared thermal meter (shown in **Figure 6-16**) are used to measure the temperature of different parts of the prototype motor. The infrared thermal meter measures the temperature of the motor housing surface and is movable and convenient to measure at any point of the surface. Inside of the housing frame, the temperature of the copper windings and stator core are measured by the wired thermal meter.



(Courtesy: RS company)

Figure 6-15 Wired thermal meter



(Courtesy: CENTER Technology Corp.)

Figure 6-16 Infrared thermal meter

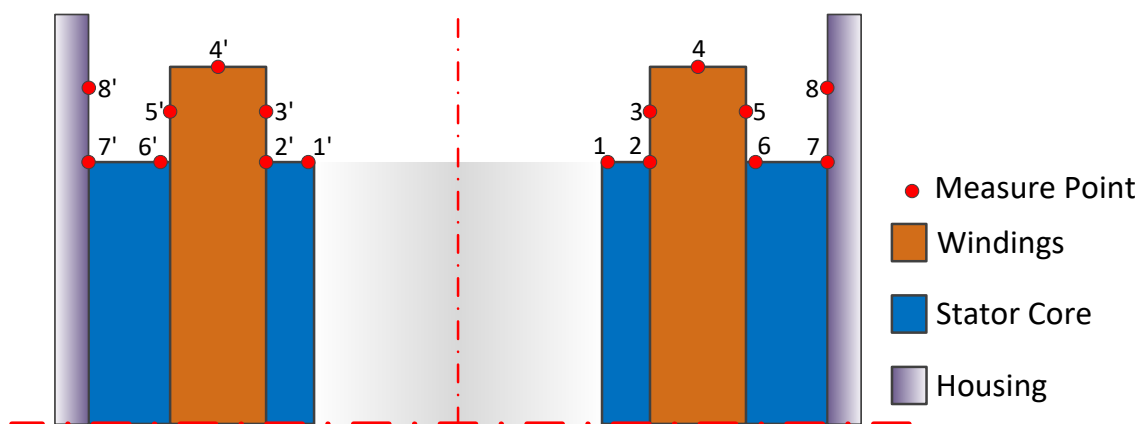


Figure 6-17 Cross-section of stator end and thermal measuring points

As shown in Figure 6-17, 16 thermal probes are divided into two groups, and each group has 8 probes and are numbered from 1 to 8. Measure points 2, 3, 4, and 5 are fixed on the surface of copper windings, where point 2 is around the corner between the winding and stator teeth. Measure point 1, 6, and 7 are fixed on the top surface of the stator core, where point 6 is at the corner between winding and stator core. And measure point 8 is mounted on the inner surface of the housing frame.

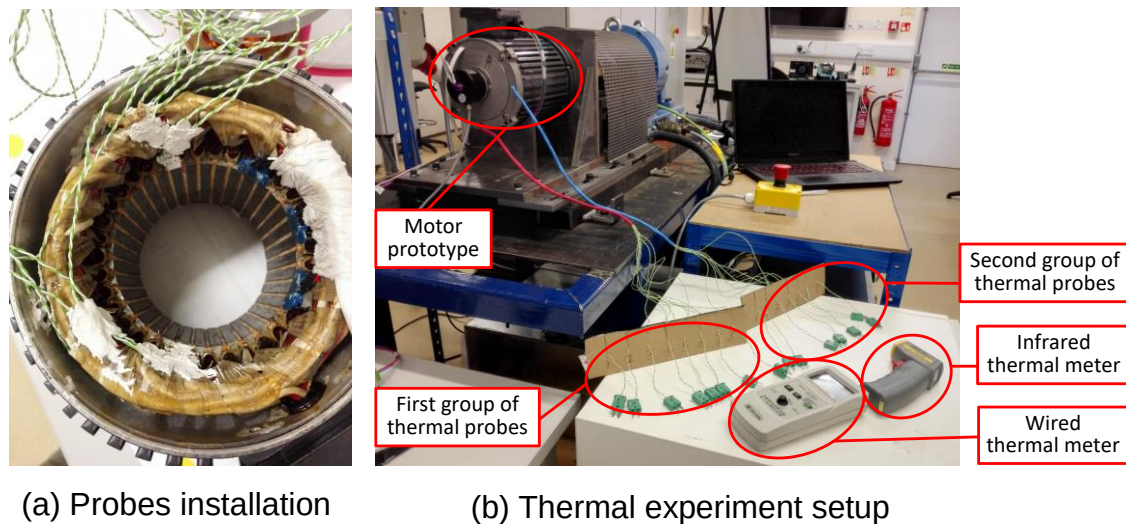


Figure 6-18 Thermal probes installation and experiment measurement

When all the thermal probes have been properly fixed inside of the motor housing, shown in **Figure 6-18(a)**, the prototype motor is reassembled and installed on the testing rig. As shown in **Figure 6-18(b)**, the setup of the thermal testing consists of two groups of thermal probes, a wired thermal meter and an infrared thermal meter. The wires of the first and second group of thermal probes are bundled in red and blue heat-shrink tubes respectively. The temperature of each measuring point inside of the motor can be measured by the external end of each thermal probe. During the thermal test, the internal temperature of the prototype motor is read by the wired thermal meter, and the temperature of the motor housing exterior surface is measured by the infrared contactless thermal meter.

6.2.2 Temperature measurement and discussion

The thermal test of the designed ferrite IPM motor takes 240 minutes and the measuring results are displayed in **Figure 6-19**. Since the relative measuring

positions of the two groups of thermal probes are the same, the temperature results shown in **Figure 6-19** are the average of the two groups. Before starting the motor, the temperatures of all the parts of the machine are the same, which is the ambient temperature 27.2°C. In order to compare with the simulation results conveniently, the beginning temperature is adjusted to 20°C while the temperature increases are kept unchanged.

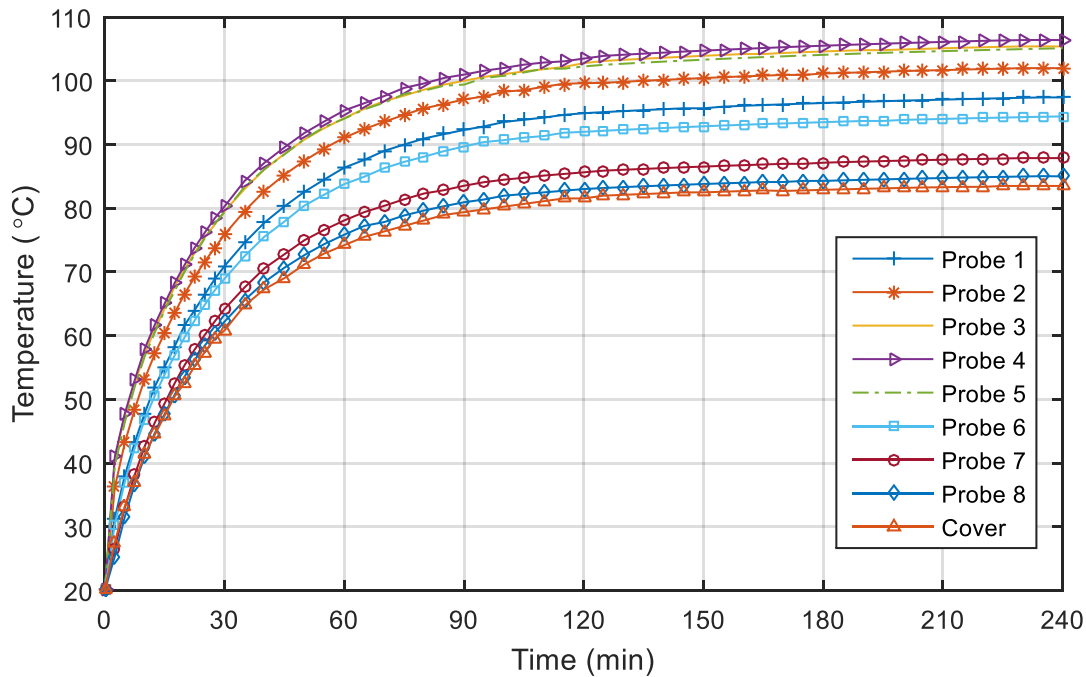


Figure 6-19 Measured temperature of the ferrite IPM motor

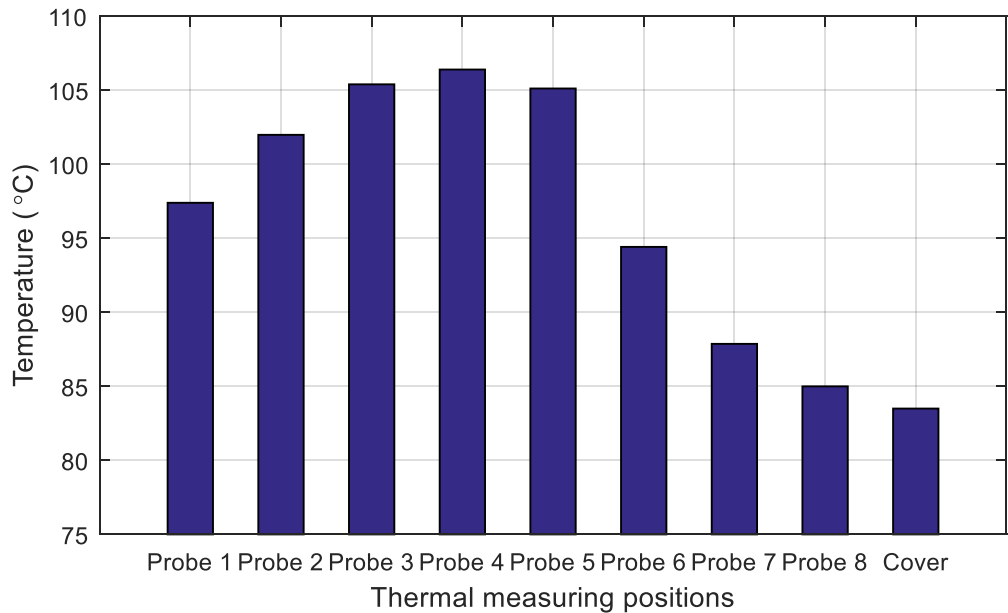


Figure 6-20 Temperature distribution of the motor at the steady sate

As shown in **Figure 6-19**, the temperatures of the motor increase in the first hour and gradually slowdown in the second hour. After two hours of running, the motor is getting into a thermal balanced state, and there is a little variation of temperature in the last two hours. Although the temperature growing rates are different for these various measuring positions, there are no cross lines in the temperature changing curves. That means all the measuring points maintain the same thermal gradients all the time, and the temperature distribution of the motor at a steady state is depicted in **Figure 6-20**. The temperature sequence of the measurement is: *Probe 4 > Probe 3 > Probe 5 > Probe 2 > Probe 1 > Probe 6 > Probe 7 > Probe 8 > Cover*. Since the copper loss is the primary heat source, the temperature of the winding increases rapidly and is always higher than any other part of the motor. As illustrated in **Figure 6-17**, the probe 3, 4 and 5 measure the copper winding temperature. The measuring point 4 is at the top of the coil, and it is much harder for it to dissipate heat to the stator core, so its temperature is a little higher than the other two points'. Since all the generated heat is transferred from the inside to the outside and the area near the housing is cooler, the temperature of point 5 is slightly lower than that of point 3. The point 2, at the corner of the copper coil and stator tooth is heated by the winding but is easier than point 3, 4 and 5 to dissipate its heat as it adheres to the stator core. Measuring points 1, 6 and 7 are all glued on the stator core, from the tooth tip to stator yoke. From the analysis in Figure 6.6, it is known that the iron loss density at the tooth tip area is the highest, followed by the stator tooth and then stator yoke. In addition, the rotor is also a heat source caused by the iron loss of rotor core, and most of its heat is transferred through the air gap to the stator core. Hence the inner part of the stator core is hotter than the outer part, and the temperature of measuring point 1 is higher than point 6 that is higher than point 7. Point 8 is fixed on the interior side of the motor housing but not on the stator core, so its temperature is between point 7 and exterior cover. The temperature of the exterior cover is the lowest, which is the thermal interface between the motor and the ambient air. Due to the temperature difference, the generated heat can be continuously brought away from the inside of the motor to the ambience. In addition, in the current design, the motor does not consist of a cooling fan to

help dissipate heat. Since it is solely dependent on natural cooling, it takes a relatively long time for the motor to get into thermal steady state and the steady temperature is relatively high. A cooling fan can be involved in the next generation of motor design, so that the overall temperature will be lower and it will take a shorter time to become thermal steady.

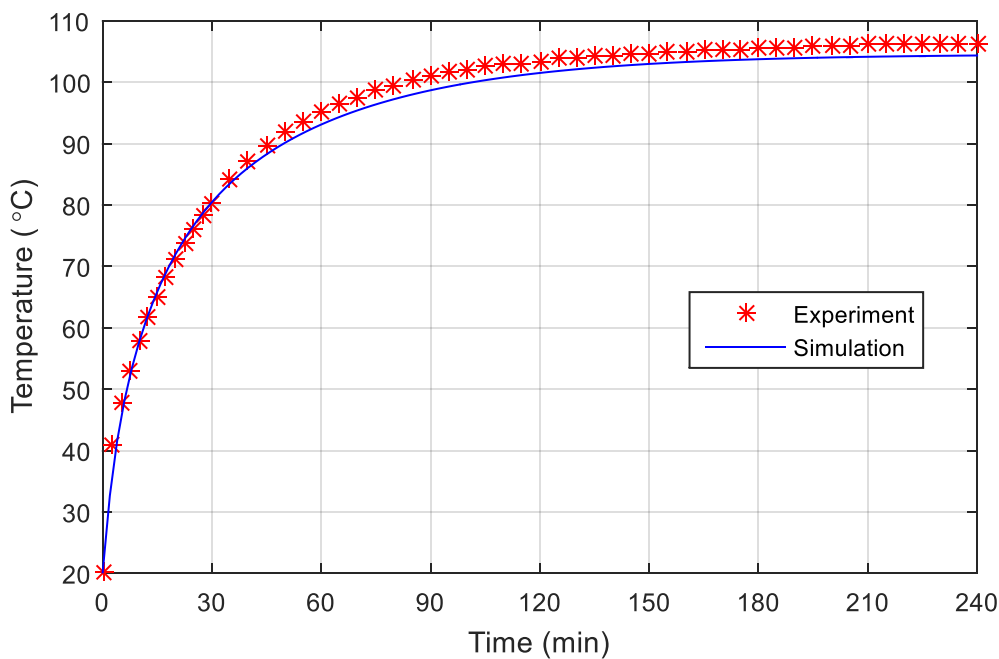


Figure 6-21 Temperature comparison of copper coil

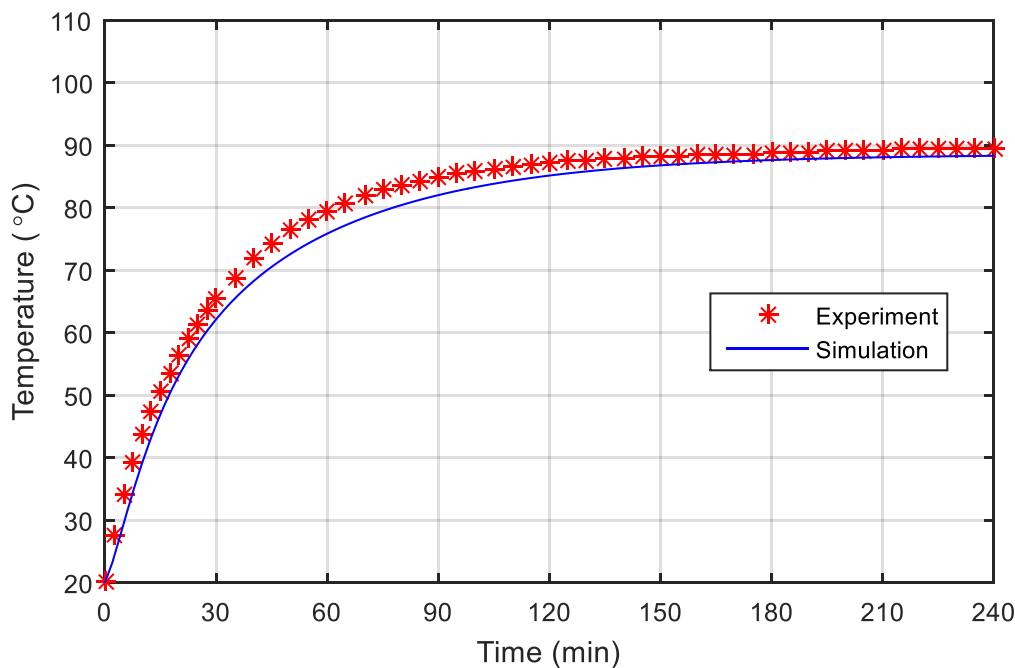


Figure 6-22 Temperature comparison of the stator core

The thermal testing results of the designed ferrite IPM motor are compared with its simulation results that are obtained in chapter 4. The comparison of copper coil temperature and stator core temperature are depicted in **Figure 6-21** and **Figure 6-22**, respectively. As shown in **Figure 6-21**, the temperature results of the experiment have a close agreement with that of simulation in the first half hour, but after that, the experimental temperature is about 2°C higher than the simulated temperature. This might be because the actual thermal testing is conducted in a closed room and the ambient temperature is increased during the experiment, which is caused by the heat that is generated from the running motor. But the ambient temperature in the simulation is set to be constant, with virtually no increase during the operating process.

The stator core temperature at the corresponding position is estimated from the temperatures of its nearby measuring point 6 and 7 and compared with simulated results in **Figure 6-22**. The actual testing stator core temperature grows a little faster than the simulated temperature, and all get into the equilibrium state after around 150 minutes. The testing temperature is higher, and the largest difference is about 4°C happening around the 60th minute. And in the steady state, the temperature is about 1.5°C. The inconsistency of lamination material may cause the difference in the stator core temperature. The actual loss coefficient of steel sheets might be higher than the one used in the simulation. The increasing room temperature also affects the experimental results. Despite the discrepancies between the experiment and simulation results, the general tendencies are the same. Therefore, the thermal analysis of the ferrite IPM motor is validated by the experiment testing.

6.3 Conclusion

A prototype of the proposed ferrite design is manufactured and tested under various load conditions. Meanwhile, a commercial IM with the same housing and stator lamination is also tested as a benchmark. The testing results validate the FEA simulations and have confirmed the advantageous overall performance of the proposed ferrite motor over the commercial IM. The ferrite IPM is able to deliver over 33% higher torque density and achieved 4-5% higher efficiency at

the highest region in the efficiency map with much higher power factor. Considering the huge number of IMs installed in industry, there is a vast potential market in developing ferrite PM machines to replace the IM for various industrial applications.

The four-hour continuous thermal test shows the temperature distribution and change of the stator. The simulation results of the temperature changes of copper coil and stator core exhibit good agreement with the experimental results. Therefore, the proposed methods and ideas are well validated by the experiments.

7 CONCLUSION AND FUTURE WORK

7.1 Conclusions

A novel two-layer configuration for the ferrite IPM machine is presented and investigated in this thesis to provide an alternative solution to replace IMs for various industrial applications. The investigations were focused on the systematic design of the proposed ferrite machine.

An extensive literature review is conducted to understand the background and current development of electric machines. To achieve the aims of the project, a novel two-layer ferrite interior PM machine is proposed based on the literature as the most potential solution. Then the optimal design is carried out with the development of analytical model so as to reveal the influence of key design parameters of such machines. FEA models are also built to verify the analytical methods. With a commercial 1.5kW IM as a benchmarking motor, the proposed design adopts the housing and stator of the commercial 1.5kW IM, with the rotor replaced by the two-layer ferrite IPM rotor. Comprehensive performance comparisons in terms of torque profiles, losses, efficiency, power factor and material costs, are carried out to exhibit the great superiority of ferrite IPM design over the commercial IM.

Finally, a prototype of the proposed ferrite design is manufactured and tested under various load conditions. Meanwhile, the commercial IM with the same housing and stator lamination is also tested as a benchmark. The testing results validate the FEA simulations and confirmed the advantageous overall performances of the proposed ferrite motor over the commercial IM.

7.1.1 Two-Layer Configurations

Spoke type configurations and PMASynRMs appeared to be the most potential candidates for high performance ferrite machines in literature. Spoke type IPMs used flux focusing structure and were able to achieve a similar PM flux field with rare-earth machines and high PM torque. PMASynRMs obtained high saliency ratio and harness high reluctance torque. By introducing multiple flux barrier concept of PMASynRM into the flux spoke type machine, a better compromise

between PM torque and reluctance torque can be met, and a most promising multi-layer spoke type structure was possible to achieve better torque density due to high PM flux density and excellent rotor saliency.

7.1.2 Design of Two-Layer Ferrite Machine

Analytical models of the two-layer IPM is developed based on the equivalent magnetic circuit theory to provide a fast and accurate way of optimal design for such structure. The influences of key design parameters are also revealed to offer guidelines of performing optimisation. In this project, the power and torque density are the major focuses, and the optimal design is achieved by striking the optimal compromise between PM torque and reluctance torque.

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7.1.3 Benchmarking Two-Layer Ferrite Machine

The performances of the novel two-layer spoke-type ferrite motor is investigated, with a commercial IM having the same housing and stator laminations as benchmark. Their performances, including torque capability, power density, losses, efficiency and power factor, are analysed and compared comprehensively via FEA simulations to reveal the advantages IPM machine. Cost analysis is also carried out. The FEA simulation results exhibits superiority to the IM in terms of per kilowatt cost, torque capability, power density, efficiency and power factor.

7.1.4 Core loss and thermal analysis

High-frequency PWM carrier waves are superimposed on the basic current forms when it is supplied by PWM inverter. The induced high-frequency magnetic field will increase core losses of the electric machine. The iron loss analysis is carried out by taking the PWM harmonics into consideration, and a more accurate loss estimation is achieved to provide useful advices for motor housing installation and motor PWM control. From the FEA simulation, it is shown that the high temperature region appears in the windings, as copper loss is the dominant source of heat. The moderate temperature region is the rotor, where the heat is accumulated from rotor iron loss but difficult to dissipate. The temperature of

stator core is relatively low, because the heat generated from stator core loss is easy to run away into the ambience. The thermal analysis of the studied motor provides a deeper understanding of motor losses and helps to determine consecutive operating time in overload conditions.

7.1.5 Experimental validation of the machines

A prototype ferrite IPM machine is manufactured and tested to validate the proposed motor design methods and ideas. A series of tests is performed under various working conditions to obtain the characteristics of the prototype machine, which include open-circuit test, no-load test, loaded test and thermal test. From the comprehensive comparison between the FEA and experimental results, it can be concluded that the FEA analyses have close agreement with the actual prototype tests in terms of back EMF, torque capacity power factor and motor efficiency. In the four-hour continuous thermal test, the experimental results of the temperature changes in the copper coils and stator core exhibit good agreement with the simulation results. Therefore, the proposed methods and ideas are well validated by the experiments.

In all, despite slightly poorer flux weakening ability, the proposed ferrite design exhibits great superiority to the IM in performance:

- Although the total machine cost of ferrite motor is slightly higher, the per kilowatt cost of ferrite motor is 20% lower;
- The ferrite achieves 38% higher torque density;
- 8-10% higher efficiency under the operating range
- More than 20% higher power factor, especially under light load conditions.

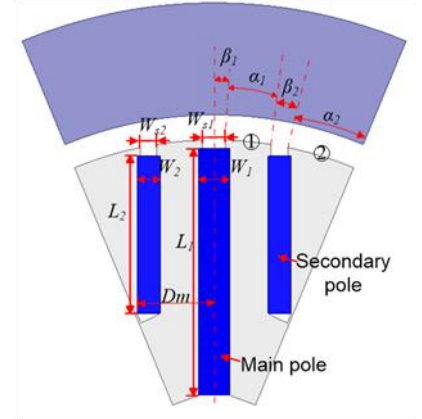
The proposed novel two-layer ferrite IPM design shows great potential for various applications in domestic and industry. With thorough theoretical and experimental validation of the prototype machine, the advantageous performance is confirmed.

7.2 Future Work

1. A processing approach can be designed to help in calculating L_d , L_q and U_f more precisely using Universal Fuzzy Logic approximator combined with a meta-heuristic optimisation algorithm. The preliminary results listed in (Appendix E) shows the possibility to use this design method.
2. Control strategies of the electric drive system based on MTPA control have been implemented in the electric drive. Some other control strategies could also be used to further improve the capability of the drive system, such as unity power factor control, flux weakening control, direct torque control, maximum efficiency per ampere control, and so on. The Universal Fuzzy Logic approximator could be used in the control algorithm calculations.
3. The torque ripple of the two-layer configurations is relatively high. In order to extend the usage of ferrite PM machines to the applications with higher torque requirements, it is of great importance to improve the design to reduce the ripple in the torque.
4. In the thesis, only the machining accuracy is considered, and other factors are assumed ideal and constant. In future works, the factors such as material consistency and assembling accuracy should be included in the electric machine.
5. Since there is a large amount of ferrite materials inserted inside the machine rotor with very long PM poles, the only very thin connection in the rotor is left to hold the whole rotor together. And sometimes even modular rotor designs are used with the rotor part completely disconnected. Thus, it is of great importance to investigate the mechanical design of the rotor, especially under high speed applications.
6. To reduce the cost of the overall drive system for ferrite IPM machines. the low-cost controller should also be developed accordingly, for example using single resistor current sampling and Sensorless control to reduce the cost of the whole system further. Since the harmonic contents under light load conditions are relatively high, improvements should be made to extract useful information from the current feedback.

APPENDICES

Appendix A Analytical Models of Multi-layer Spoke Type Machines



$$H_{m1} \cdot W_1 + 2 \cdot H_{g1} \cdot g_c = 0 \quad (\text{A1})$$

$$H_{m1} \cdot W_1 + H_{s1} \cdot W_{s1} = 0 \quad (\text{A2})$$

Fig. A1 Flux loops excited by PMs of two-layer configurations.

For the flux passing through area ② and secondary pole slots, the following can be obtained:

$$H_{m1} \cdot W_1 + 2H_{m2} \cdot W_2 + 2H_{g2} \cdot g_c = 0 \quad (\text{A3})$$

$$H_{m2} \cdot W_2 + H_{s2} \cdot W_{s2} = 0 \quad (\text{A4})$$

And based on the flux conversion law, flux equations are given by

$$\phi_{m1} = \phi_{g1} + \phi_{g2} + \phi_{ml1} \quad (\text{A5})$$

$$\phi_{m2} = \phi_{g2} + \phi_{ml2} + \phi_{bri} \quad (\text{A6})$$

where

$$\begin{cases} \phi_{m1} = B_{m1} \cdot L_1 \cdot l_{ef} \\ \phi_{m2} = B_{m2} \cdot L_2 \cdot l_{ef} \end{cases} \quad (\text{A7})$$

$$\begin{cases} \phi_{ml1} = B_{ml1} \cdot L_{s1} \cdot l_{ef} \\ \phi_{ml2} = B_{ml2} \cdot L_{s2} \cdot l_{ef} \end{cases} \quad (\text{A8})$$

$$\phi_{g1} = l_{ef} \cdot \int_0^{\frac{\pi}{p}} B_g \cdot R_{ro} d\theta = l_{ef} \cdot \left(\int_{\frac{\pi}{p} - \alpha_1 - \beta_1}^{\frac{\pi}{p} - \beta_1} B_{g1} \cdot R_{ro} d\theta + \int_{\frac{\pi}{p} - \beta_1}^{\frac{\pi}{p}} B_g \cdot R_{ro} d\theta \right) \quad (\text{A9})$$

$$\phi_{g2} = l_{ef} \cdot \int_0^{\alpha_2 + \beta_2} B_{g2} \cdot R_{ro} d\theta = l_{ef} \cdot \left(\int_0^{\alpha_2} B_{g2} \cdot R_{ro} d\theta + \int_{\alpha_2}^{\alpha_2 + \beta_2} B_g \cdot R_{ro} d\theta \right) \quad (\text{A10})$$

$$\begin{cases} \beta_1 = \arcsin \frac{W_{s1}}{2R_{ro}} \\ \beta_2 = \arcsin \frac{D_m}{R_{ro}} - \arcsin \frac{D_m - W_{s2}}{R_{ro}} \end{cases} \quad (\text{A11})$$

$$\begin{cases} \alpha_1 = \arcsin \frac{D_m - W_{s2}}{R_{ro}} - \beta_1 \\ \alpha_2 = \frac{\pi}{p} - \arcsin \frac{D_m}{R_{ro}} \end{cases} \quad (\text{A12})$$

The saturation bridges of the secondary poles are design to operate at completely saturated conditions, and thus the flux through the bridge can be approximately estimated as:

$$\phi_{bri} = l_{ef} B_{sat} \frac{h_{bri}}{2} \quad (\text{A13})$$

where B_{sat} is the saturation flux density of the laminations.

For the PMs,

$$B_m = B_r + \mu_0 \mu_m H_m \quad (\text{A14})$$

For the other air regions

$$B = \mu_0 H \quad (\text{A15})$$

Firstly assume that at the PM slot area, the flux density is zero, then (A9) and (A10) can be simplified as:

$$\phi_{g1} = l_{ef} \cdot \int_{\frac{\pi}{p} - \alpha_1 - \beta_1}^{\frac{\pi}{p} - \beta_1} B_{g1} \cdot R_{ro} d\theta = l_{ef} B_{g1} R_{ro} \alpha_1 \quad (\text{A16})$$

$$\phi_{g2} = l_{ef} \int_0^{\alpha_2} B_{g2} \cdot R_{ro} d\theta = l_{ef} B_{g2} R_{ro} \alpha_2 \quad (\text{A17})$$

By solving equation (A5)(A3)-(A6), the airgap flux density can be derived:

$$B_{g1} = \frac{B_r L_1 \left(\mu_m \frac{g_c L_2}{W_2} + \frac{2g_c L_{s2}}{W_{s2}} + R_{ro} \alpha_2 \right) - B_r L_2 R_{ro} \alpha_2 + \frac{B_{sat} h_{bri} R_{ro} \alpha_2}{2}}{\mu_m \frac{g_c L_2}{W_1} R_{ro} \alpha_2 + \left(\mu_m \frac{g_c L_2}{W_2} + \frac{2g_c L_{s2}}{W_{s2}} + R_{ro} \alpha_2 \right) \left(2\mu_m \frac{g_c L_1}{W_1} + \frac{2g_c L_{s1}}{W_{s1}} + R_{ro} \alpha_1 \right)} \quad (\text{A18})$$

$$B_{g2} = \frac{B_r L_1 - \left(2\mu_m \frac{g_c L_1}{W_1} + \frac{2g_c L_{s1}}{W_{s1}} + R_{ro} \alpha_2 \right) B_{g1}}{R_{ro} \alpha_2} \quad (\text{A19})$$

Considering $W_1 = W_{s1}$, and $W_2 = W_{s2}$, (A18) can be simplified as

$$B_{g1} = \frac{2B_r L_1 W_1 (\mu_m g_c L_2 + 2g_c L_{s2} + R_{ro} \alpha_2) - 2B_r L_2 W_1 W_2 R_{ro} \alpha_2 + B_{sat} h_{bri} W_1 W_2 R_{ro} \alpha_2}{2\mu_m g_c L_2 W_2 R_{ro} \alpha_2 + 2(\mu_m g_c L_2 + 2g_c L_{s2})(2\mu_m g_c L_1 + 2g_c L_{s1} + R_{ro} \alpha_2 W_1)} \quad (\text{A20})$$

To take the PM slot flux into consideration, then (A9) and (A10) can be written as

$$\phi_{g1} = l_{ef} B_{g1} \left[R_{ro} \alpha_1 + \frac{2g}{\pi} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_1}{2g} \right) \right] = l_{ef} B_{g1} R_{ro} \alpha_1 \cdot k_{rs1} \quad (\text{A21})$$

$$\phi_{g2} = l_{ef} B_{g2} \left[R_{ro} \alpha_2 + \frac{4g}{\pi} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_2}{4g} \right) \right] = l_{ef} B_{g2} R_{ro} \alpha_2 \cdot k_{rs2} \quad (\text{A22})$$

where k_{rs1} and k_{rs2} are the rotor slot coefficient for PM slots, and

$$k_{rs1} = \left[1 + \frac{2g}{\pi R_{ro} \alpha_1} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_1}{2g} \right) \right] \quad (\text{A23})$$

$$k_{rs2} = \left[1 + \frac{4g}{\pi R_{ro} \alpha_2} \ln \left(1 + \frac{\pi \cdot R_{ro} \beta_2}{4g} \right) \right] \quad (\text{A24})$$

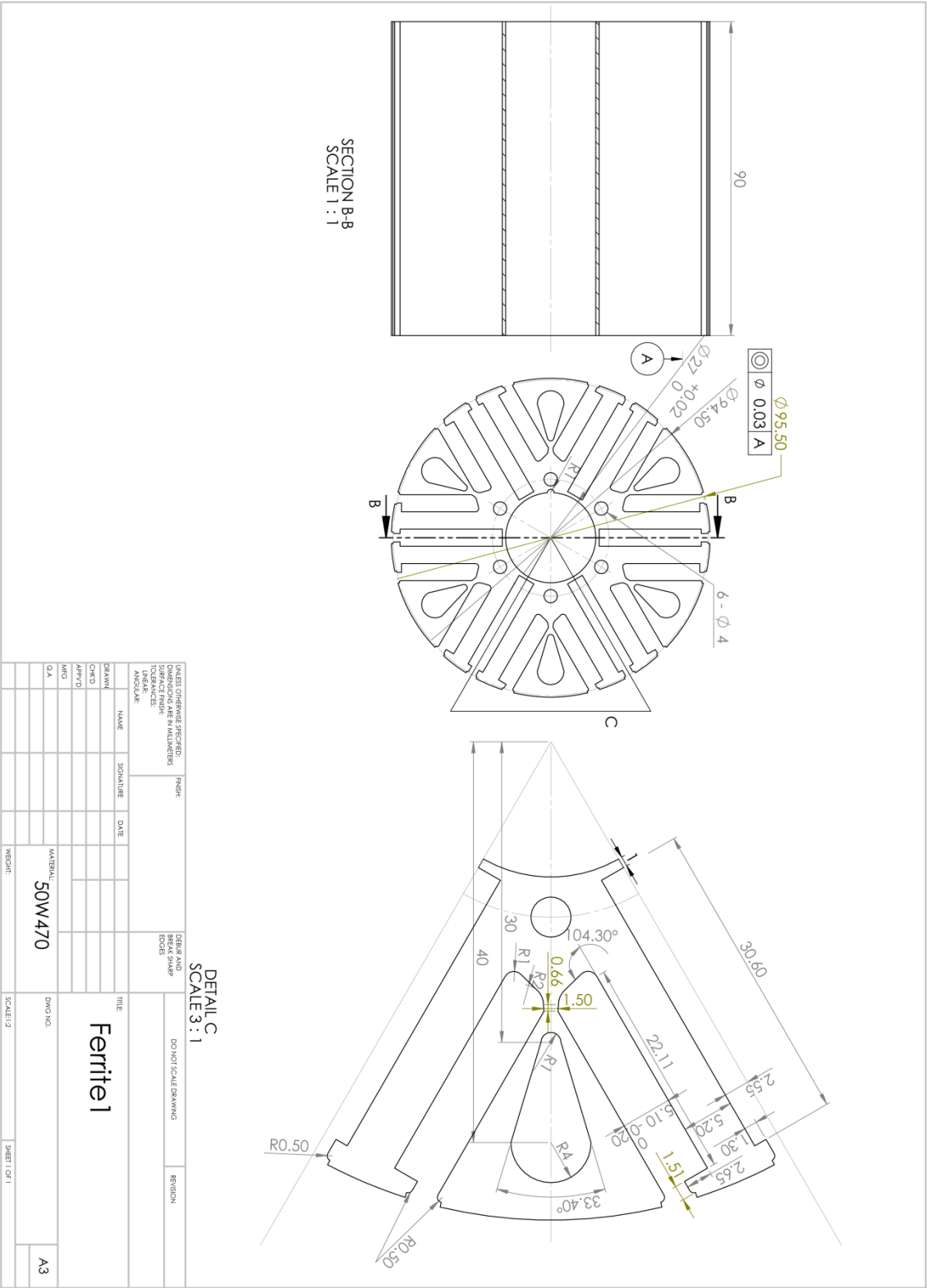
Then

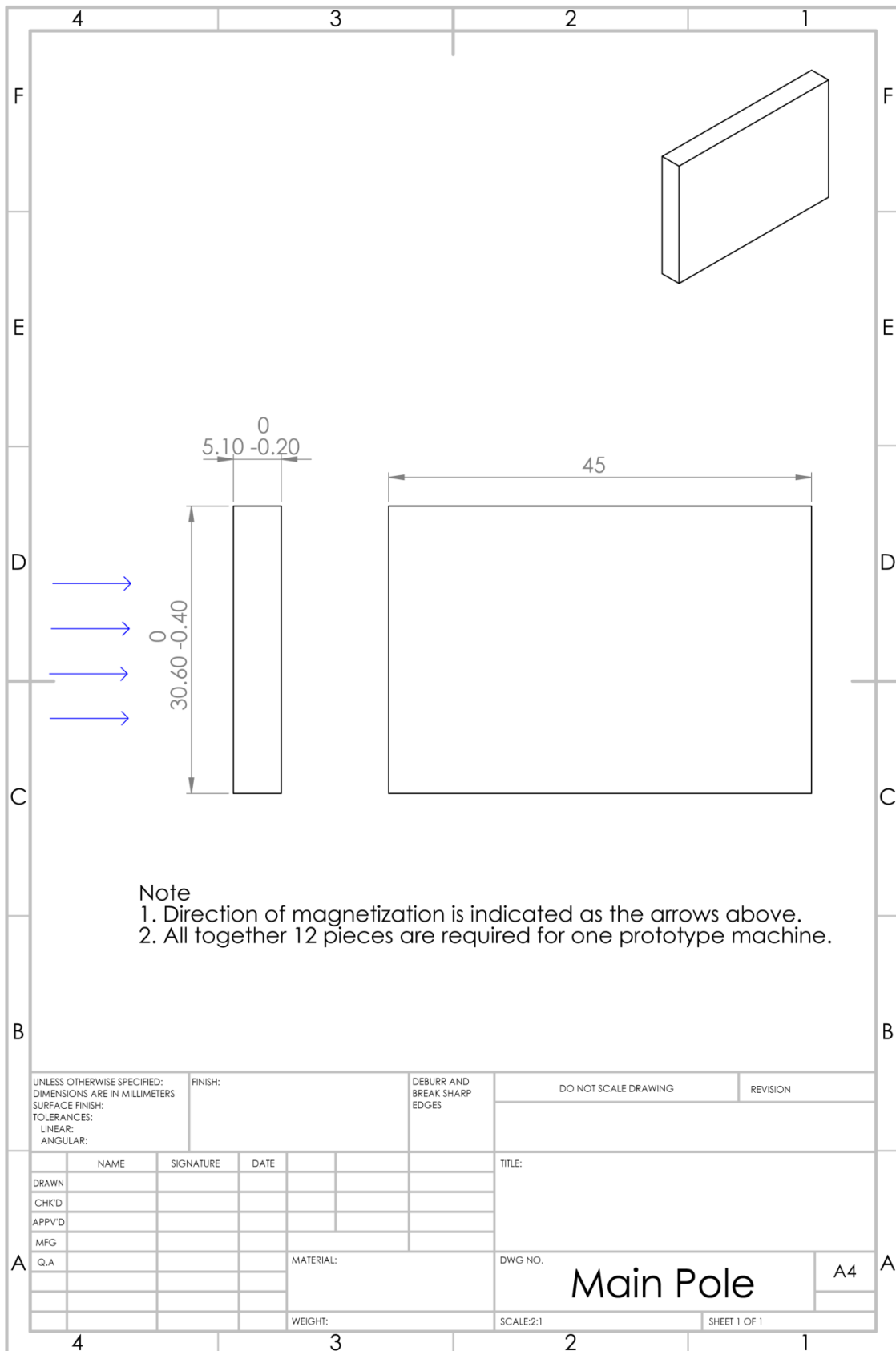
$$B_{g1} = \frac{B_r L_1 \left(\mu_m \frac{g_c L_2}{W_2} + \frac{2g_c L_{s2}}{W_{s2}} + k_{rs2} R_{ro} \alpha_2 \right) - \left(B_r L_2 - \frac{B_{sat} h_{bri}}{2} \right) k_{rs2} R_{ro} \alpha_2}{\mu_m \frac{g_c L_2}{W_1} k_{rs2} R_{ro} \alpha_2 + \left(\mu_m \frac{g_c L_2}{W_2} + \frac{2g_c L_{s2}}{W_{s2}} + k_{rs2} R_{ro} \alpha_2 \right) \left(2\mu_m \frac{g_c L_1}{W_1} + \frac{2g_c L_{s1}}{W_{s1}} + k_{rs1} R_{ro} \alpha_1 \right)} \quad (\text{A25})$$

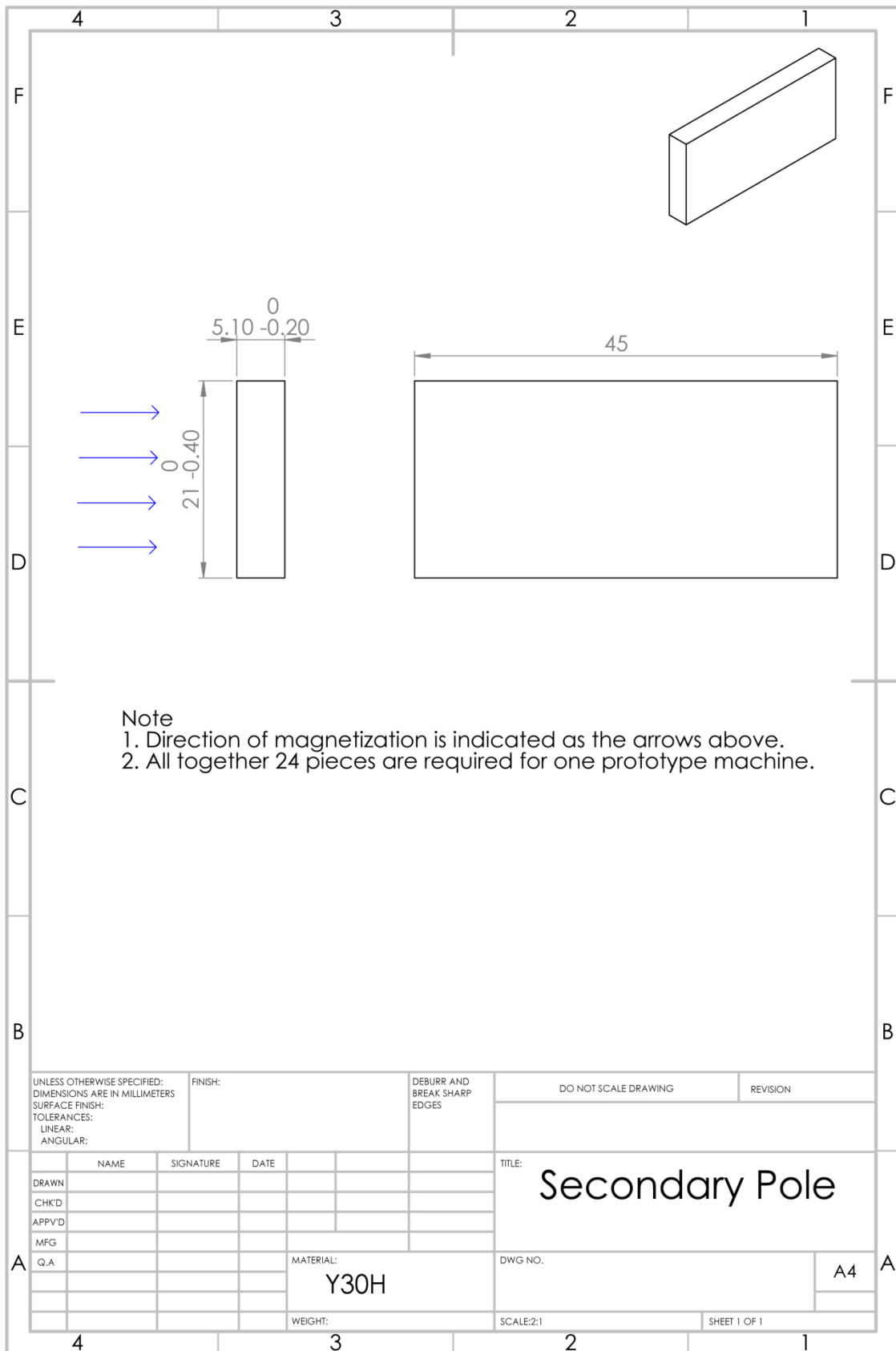
$$B_{g2} = \frac{B_r L_1 - \left(2\mu_m \frac{g_c L_1}{W_1} + \frac{2g_c L_{s1}}{W_{s1}} + k_{rs2} R_{ro} \alpha_2 \right) B_{g1}}{k_{rs2} R_{ro} \alpha_2} \quad (\text{A26})$$

By decomposing the flux distribution into Fourier's series, the v th order of flux distribution can be obtained:

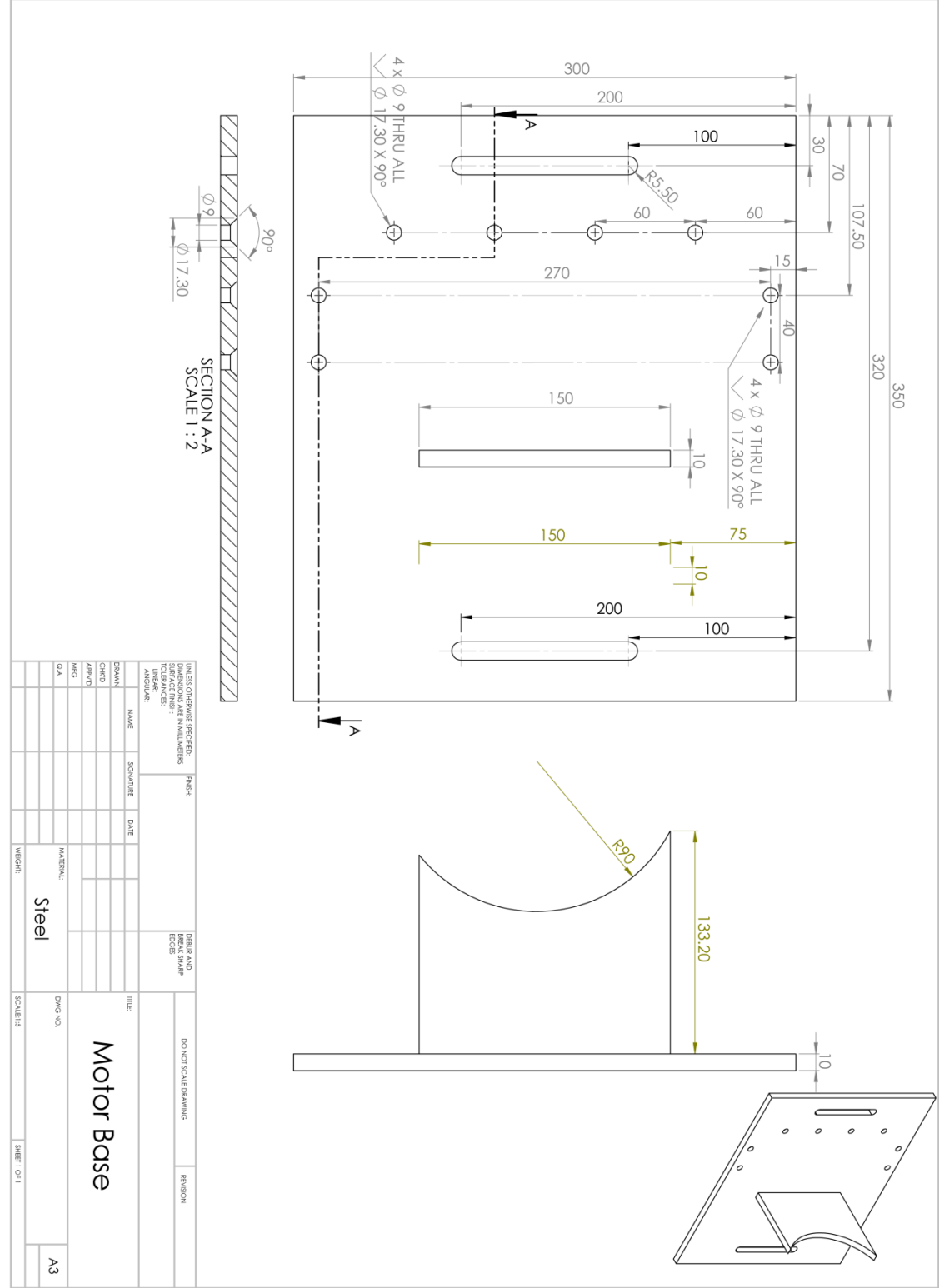
$$\begin{aligned} B_{fgv} &= \frac{1}{\pi} \int_0^{2\pi} B_{fg}(\theta) \cos v\theta d\theta \\ &= \frac{4}{v\pi} B_{g2} \sin \left(v \frac{p\alpha_2}{2} \right) + \frac{4}{v\pi} B_{g1} \left[\sin \left(v \frac{\pi - p\beta_1}{2} \right) - \sin \left(vp \frac{\alpha_2 + \beta_2}{2} \right) \right] \\ &\quad + \frac{4}{\pi} \frac{pg_c}{\pi R_{ro}} B_{g2} \int_{\frac{p\alpha_2}{2}}^{\frac{p\alpha_2 + p\beta_2}{2}} \frac{\cos v\theta}{v\theta + v \left(\frac{pg_c}{\pi R_{ro}} - \frac{p\alpha_2}{2} \right)} dv\theta + \frac{4}{\pi} \frac{pg_c}{\pi R_{ro}} B_{g2} \int_{\frac{p\alpha_2}{2}}^{\frac{p\alpha_2 + p\beta_2}{2}} \frac{\cos v\theta}{-v\theta + v \left(\frac{pg_c}{\pi R_{ro}} + p \frac{\alpha_2 + \beta_2}{2} \right)} dv\theta \quad (\text{A27}) \\ &\quad + \frac{4}{\pi} \frac{pg_c}{\pi R_{ro}} B_{g1} \int_{\frac{\pi - p\beta_1}{2}}^{\frac{\pi}{2}} \frac{\cos v\theta}{v\theta + v \left(\frac{pg_c}{\pi R_{ro}} - \frac{\pi}{2} + \frac{p\beta_1}{2} \right)} dv\theta \end{aligned}$$

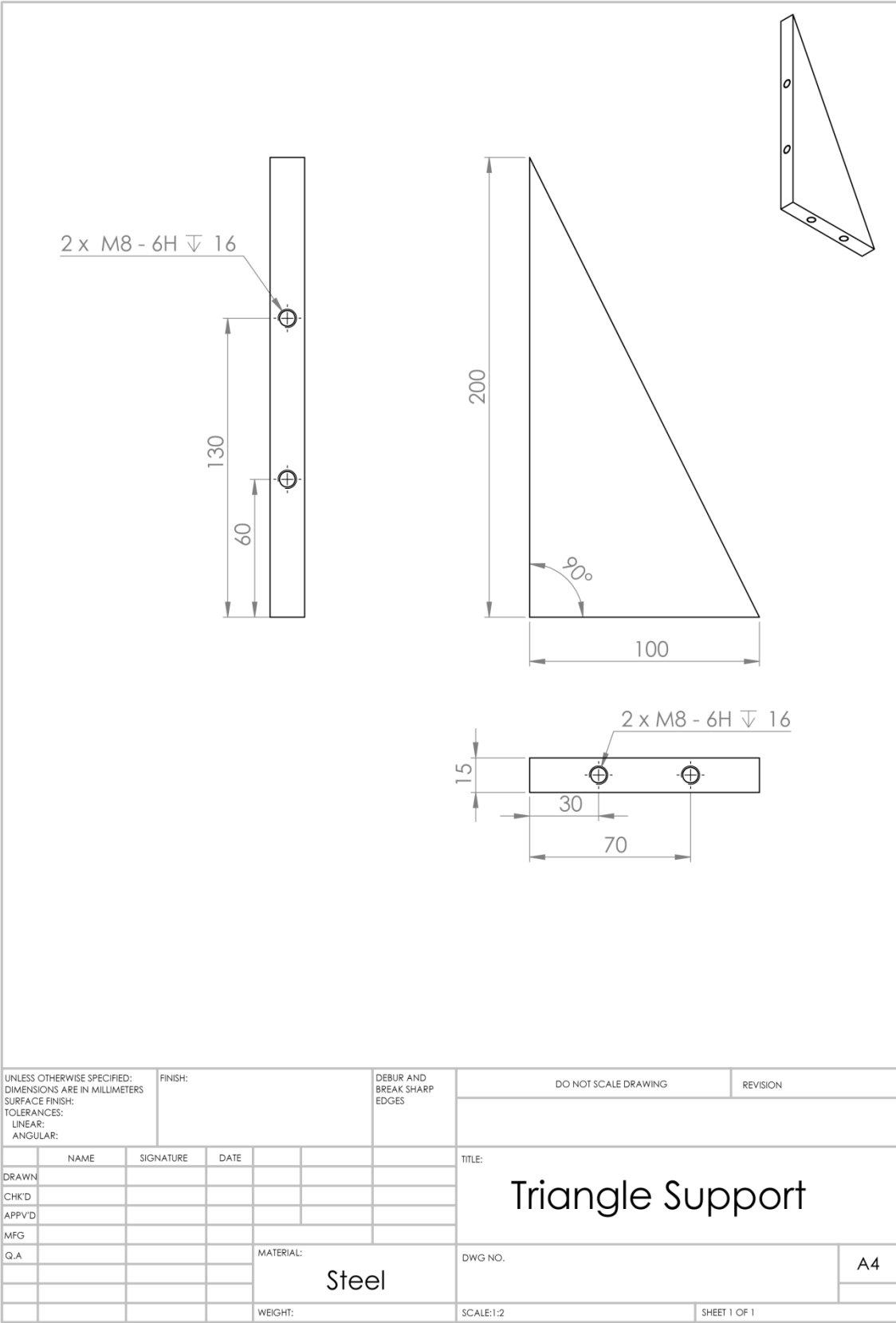


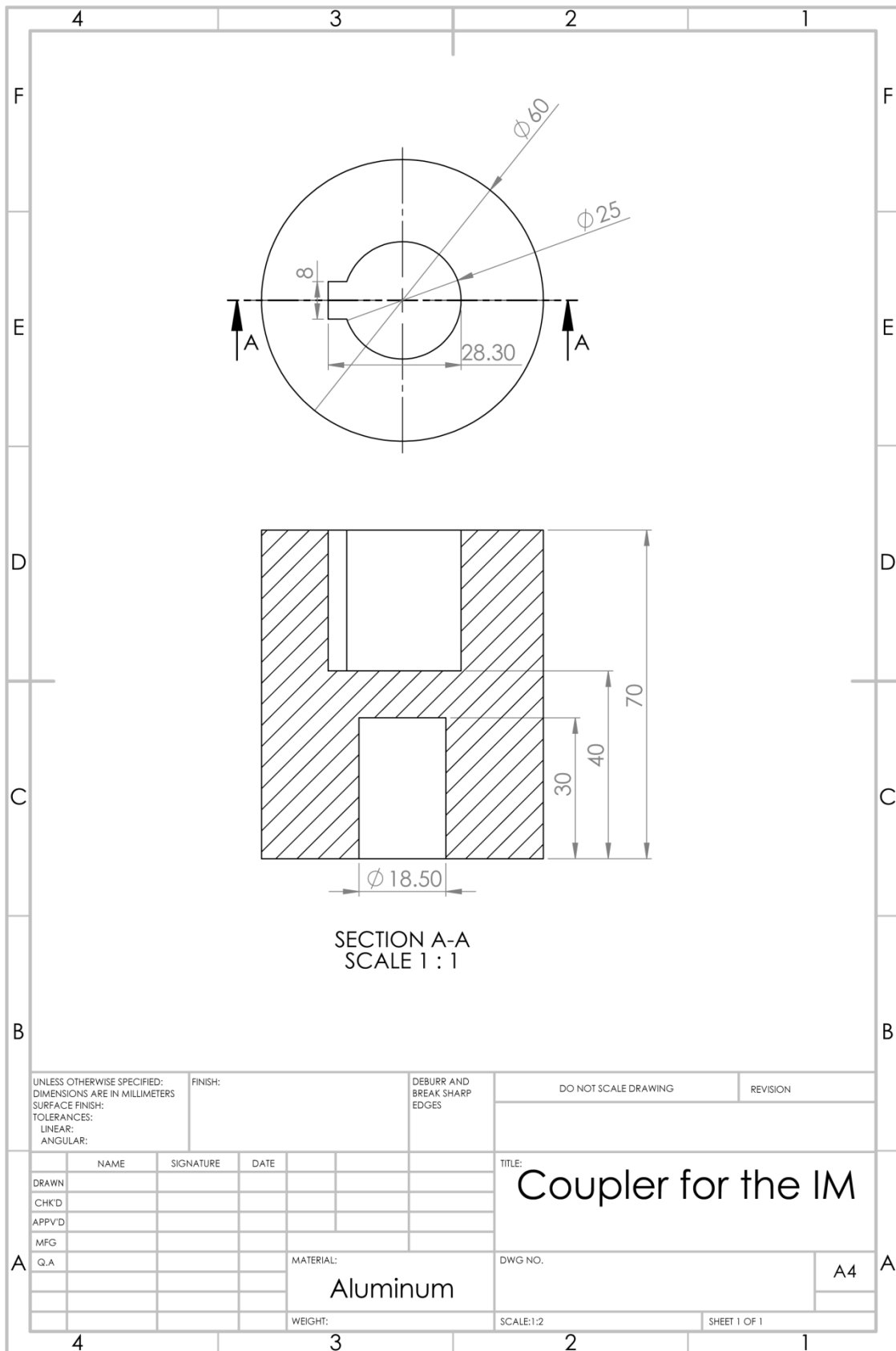




B.2 The Testing Rig







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SURFACE FINISH:									
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LINEAR:									
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NAME		SIGNATURE		DATE		TITLE:		Coupler for the IM	
DRAWN									
CHK'D									
APPV'D									
MFG									
Q.A						MATERIAL:		Aluminum	
						DWG NO.		A4	
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								SHEET 1 OF 1	

Appendix C TI High Voltage Motor Control and PFC Kit

C.1 Kit Feature

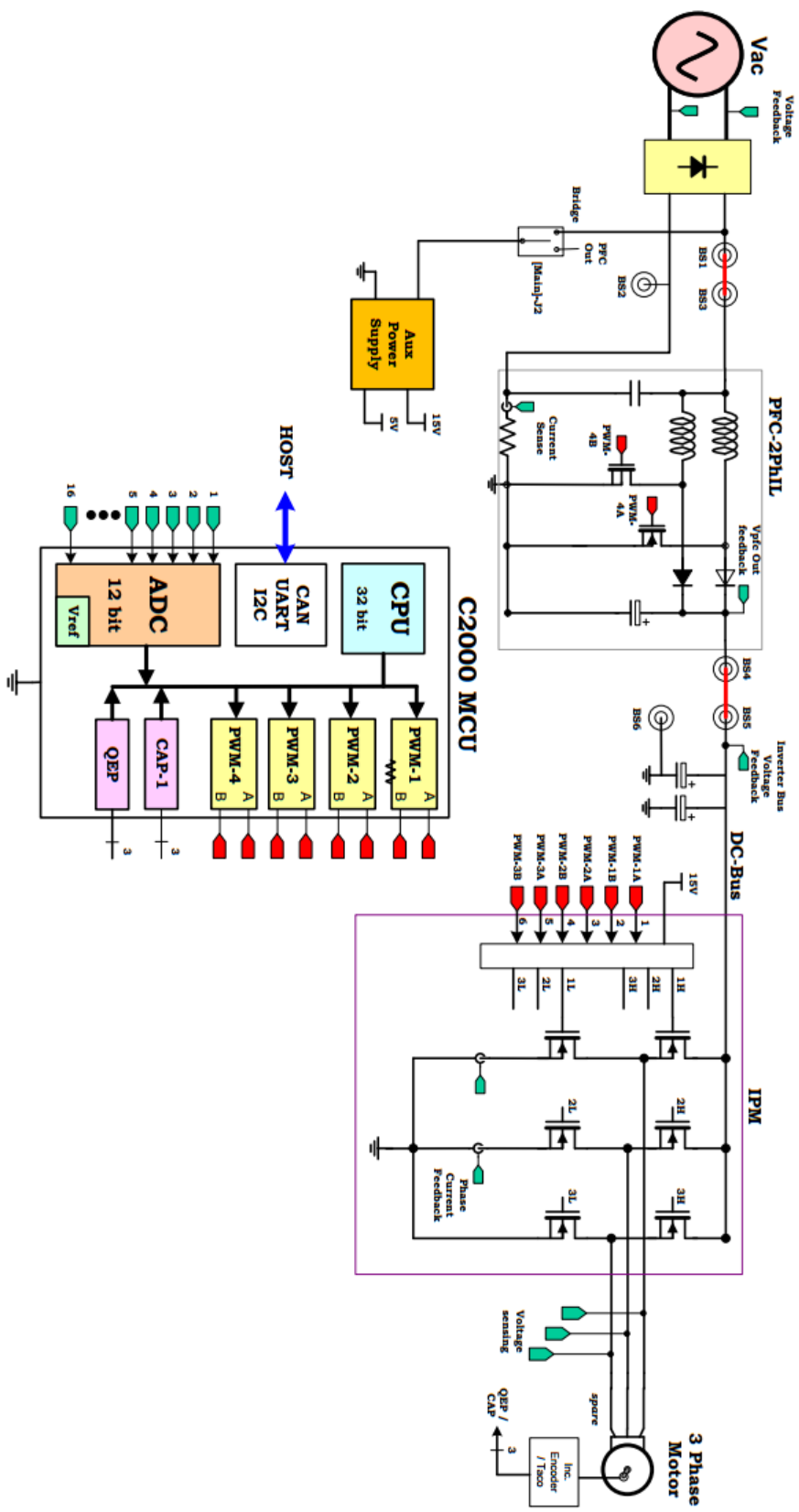
The high voltage motor control and PFC developer's kit is equipped with 32-bit floating-point microcontrollers TMS320F2803x/TMS320F2833x, which enable cost-effective design of intelligent controllers for three phase motors by reducing the system components and increase efficiency, as depicted in Figure_Apx 1. With these great advantages, it is possible to realize far more precise digital vector control algorithms.

- 350V DC max input voltage;
- 1KW/1.5KW maximum load;
- Sensorless and Sensored Field Oriented Control of PMSM Motor;
- QEP and CAP inputs available for speed and position measurement;
- High precision low-side current sensing using the C2000's high-performance ADC, TI OPA2350 high speed op-amps and Texas Instrument REF5025 high precision voltage reference chip.



Figure_Apx 1 The High Voltage Motor Control and PFC Developer's Kit

C.2 Control Kit Diagrams



Figure_Apx 2 Control kit hardware diagram

Appendix D The Programmable Dynamometer System

To validate the performances of the prototype machine under various working conditions, the programmable dynamometer system is necessary to provide required torque loadings under different rotational speeds. The whole testing system in the lab consists of the powder brake dynamometer, the controller, brake supply, torque transducer and cooling sub-system.

D.1 The Powder Brake Dynamometer

The Magtrol powder brake dynamometer (PB Series) contains magnetic powders inside. When electrical current is passing through the coils, the magnetic field is generated, and the property of the powders changes. Accordingly, the smooth braking torque is produced because of the friction between the powders and spinning rotor. Since the braking torque is generated by friction, relatively large torque can be achieved under low speed range, and the powder brake dynamometer is ideal for applications operating in the low to middle speed range or when operating in the middle to high torque range. The powder brake provides full torque at zero speed and requires water-cooling. The PB dynamometer has accuracy ratings of $\pm 0.3\%$ to 0.5% full scale with integrated optical speed sensor and is quite suitable for performance tests of electric motors.

D.2 The Programmable Dynamometer Controller

Magtrol programmable dynamometer controller employs state-of-the-art digital signal processing technology to provide superior motor testing capabilities. In the laboratory, the controller's high sample rate provides superior resolution for data acquisition and curve plotting. This allows for capturing more usable motor test data during switching, breakdown and other transitional areas of the motor test curve, with complete PC control via IEEE-488 or RS-232 interface. On the other hand, the controller can display torque, speed and power at all times, which can also be used as a manual standalone unit or part of a PC controlling system.

D.3 The Dynamometer Power Supply

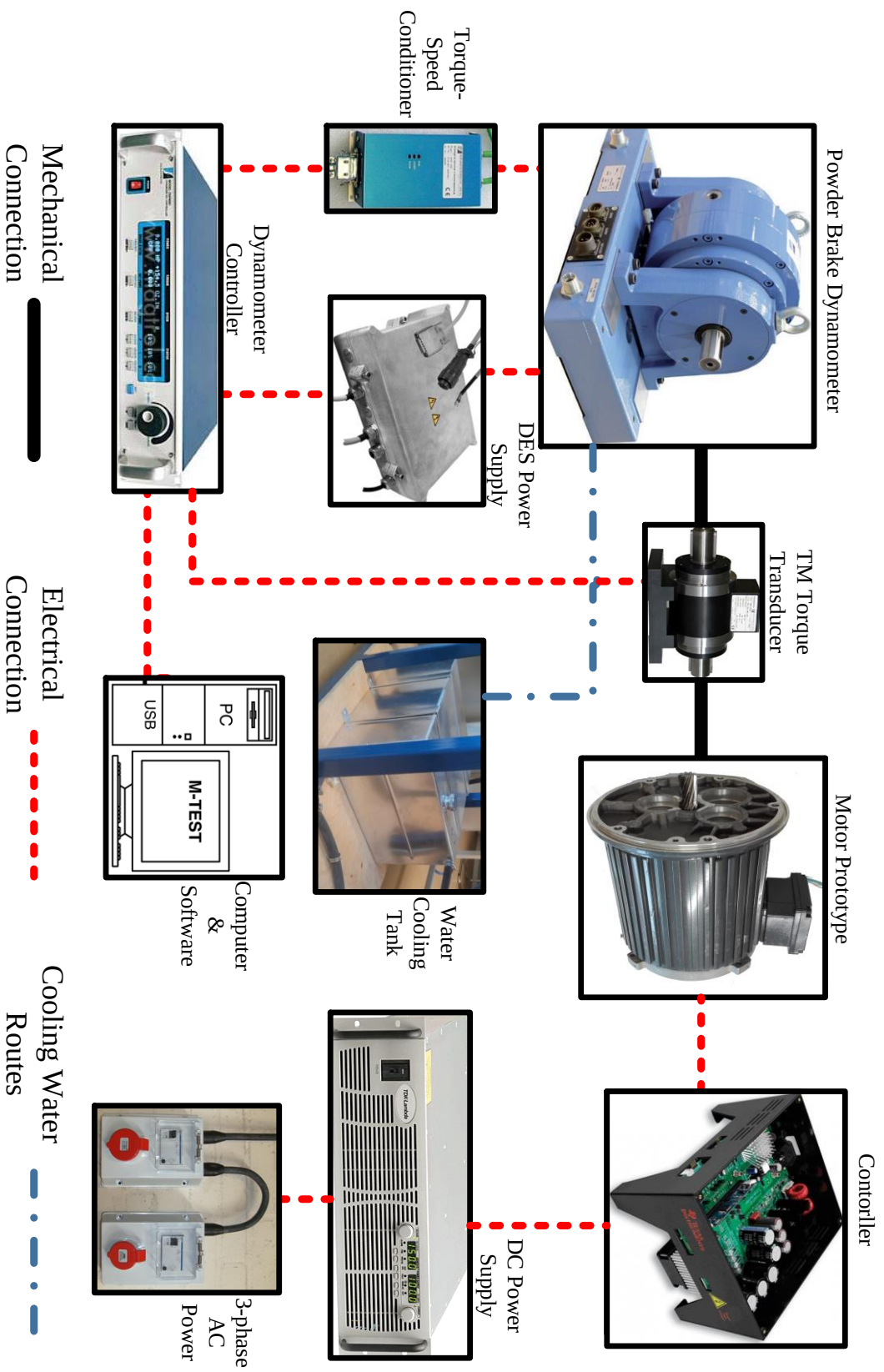
The dynamometer power supplies (DES 311) are suited to the entire range of Magtrol's powder brake dynamometers. The unit supplies the required current to the coils of the brake within the dynamometer to generate torque loadings, and the current can be controlled by analogue and digital set-points coming from an electronic peripheral, in this case, from the Magtrol's programmable dynamometer controller. To avoid any disruption of the surrounding electronic modules, the power supply is fitted in an industrial housing made of extruded cast aluminium.

D.4 The Controller In-Line Torque Transducers

The Magtrol's in-line torque transducer is a standalone measuring chain. The transducer provides extremely accurate torque and speed measurement over a very broad range. The torque transducer is very reliable, providing high overload protection, excellent long term stability and high noise immunity. Non-contact differential transformer torque measuring technology is employed for the transducer, and no electronic components rotate during operation. The transducer consists of a hardened stainless steel shaft with smooth shaft ends, an anodized aluminium housing containing the guide bearings and an electronic measurement conditioner. The integrated electronic circuit, supplied by a single DC voltage, provides torque and speed signals without any additional amplifier.

D.5 The Overall Diagram of the Testing System

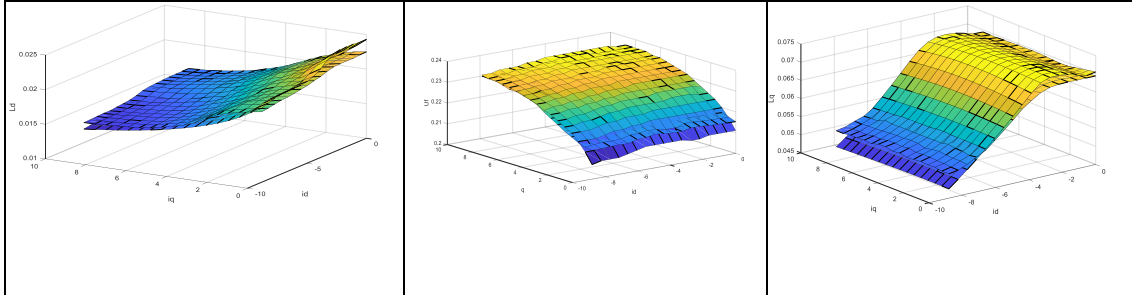
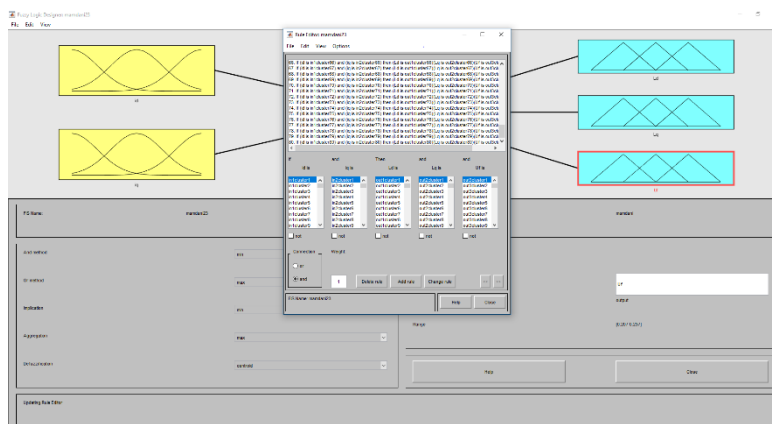
The overall diagram of the testing system is demonstrated in Figure_Apx 4. The connections among the equipment can be divided into three categories, mechanical, electrical and water cooling. The prototype machine is connected to the dynamometer through the torque transducer for mechanical power output testing. The controlling and measuring units are connected to provide the drive for the prototype machine and collect experimental data. To dissipate the heat generated in the power brake, water cooling pipes are necessary for continuous loading tests.



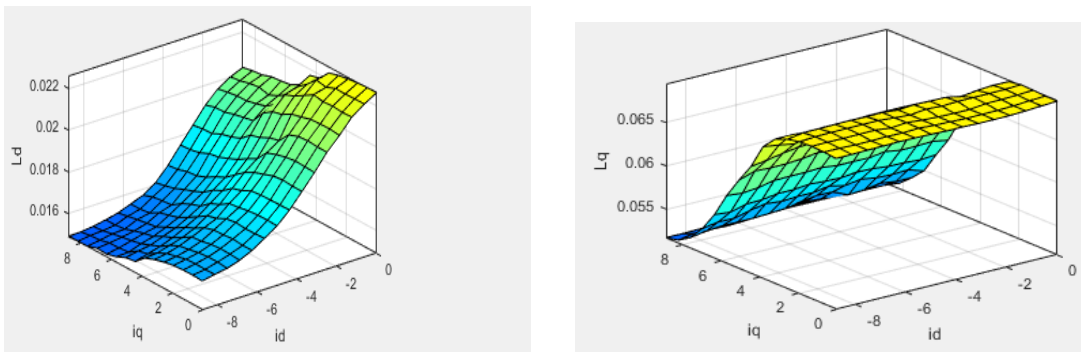
Figure_Apx 4 Control kit hardware diagram

Appendix E Universal Fuzzy Logic Approximator

Firstly, Universal Fuzzy Logic is used to find values of L_d , L_q and U_f all together in one step (two inputs: i_d & i_q and three output L_d , L_q and U_f).

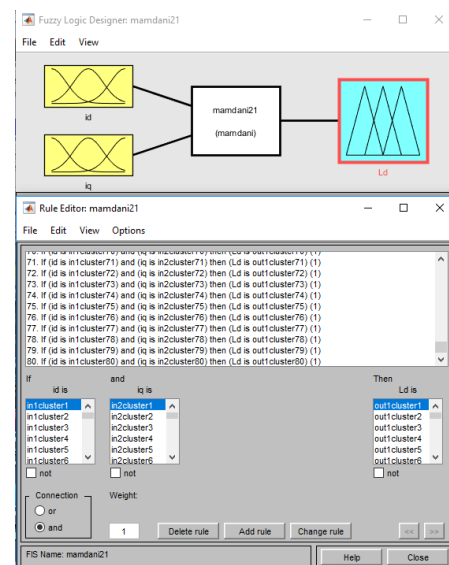
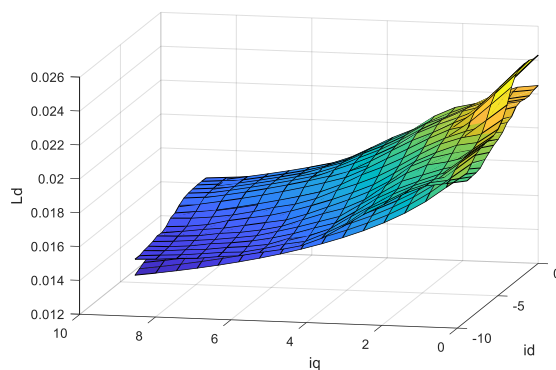


The number of clusters for all cases has been choosing 80. By changing this number there was only small change in the response figure

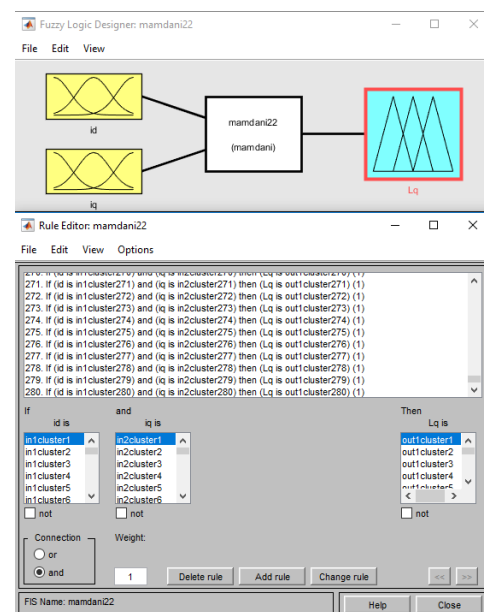
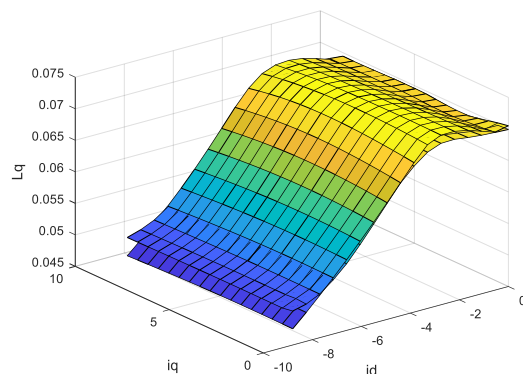


Secondly, Universal Fuzzy Logic is used to find values of L_d , L_q and U_f separately by designing three separate Fuzzy logic approximator each one with two inputs (i_d & i_q) and one output. The results show a similar to the previous case

L_d only



L_q only



Uf only

