Self-Sensing Permanent Magnet Machine

Tianhao Wang, BEng (Hons)

Thesis submitted to the University of Nottingham for the degree of Doctor of Philosophy

May, 2017

Abstract

This thesis looks at the saliency-based self-sensing control of permanent magnet synchronous machines (PMSM) and a novel machine configuration is proposed to improve the self-sensing performances. In recent time, PMSM drives have been steadily gaining popularity and have widespread applications in industry due to its benefits such as high power density, good dynamic performance and high efficiency. Self-sensing drives are superior to conventional drives in applications where the reliability and the cost are of important factors. Machine saliency is utilized for rotor position tracking during the start-up and the low speed operation when Back-EMF components are not detectable. For conventional PMSM machines, however, the saliency of Interior Permanent Magnet (IPM) machine is heavily affected by saturation effects under loaded operation; for the case of Surface Mounted Permanent Magnet (SMPM) machines, the saliency is not apparent and hard to detect. Hence the rotor position signals are relatively small or even undetectable at specific operation points, and these are the main challenges of PMSM drive self-sensing controlled at the low speed.

Addition of a novel saliency modulation rotor end (SMRE) structure to the end of a conventional PMSM rotor to improve the self-sensing capability is proposed. The SMRE provides an additional space anisotropic to the rotor. The saliency modulation of the rotor end is electrically asynchronous with the machine's rotating reference frame. Therefore, the machine saliency provided by SMRE is not affected by saturation effects under loaded operation when high frequency injection scheme is adopted in low speed ranges. In addition, for the medium and high speed range, the rotor position can be tracked without superposed injection as the saliency modulation can be achieved by taking the fundamental voltage as the carrier signal. A genetic algorithms (GA) optimization environment combined with the finite element analysis (FEA) enables to obtain optimized rotor end geometry for the maximum modulation signal and minimum total harmonic distortions (THD). The expected self-sensing performance is validated by a prototype machine and is compared with conventional PMSMs in experimental tests.

Acknowledgements

I would sincerely thank my supervisors Professor Chris Gerada, Dr. Zhuang Xu and Dr. He Zhang, who gave me the opportunity to work toward PHD. They were always supportive, patient and inspirational and I was benefited a lot from their advice, encouragement and wide vision.

I would also like to thank Dr. Michael Galea, Dr. Alberto Gaeta and Dr. Michele Degano who support me during my project, their patient guidance and advice helped me to come up with creative solutions for the project. Thanks must go to Dr Qiang Gao who worked on self-sensing control prior to me, he helped a lot on the revision of the thesis and related papers.

I would like to thank my examiners, Professor Paolo Bolognesi and Dr. Jing Li who provided constructive feedback and detailed advisory comments. I am appreciated on their careful work to help me to improve the thesis in the final stage.

I would also like express my gratitude to all those who helped and were part of the long journey towards realizing and implementing the work outlined in this thesis. Special thanks to my SEB 119 fellows, Dr. Weiduo Zhao, Bowen Shi, Yuli Bao, Anuvav Bardalai, Chuan Liu, Fengyu Zhang, Yinli Wang, Wei Dai, Tianru Zhang and Xuchen Wang for all the great time we shared.

LIST OF FIGURES

2.1 transformation from 3 phase to d/q rotating reference frame.

2.2 coupled equivalent circuits for the d and q axis of PMSM.

2.3 Block diagram of vector control system of PMSM drive.

2.4 Examples of PMSM rotor structure (a) Surface Mounted (b) Inset (c) Interior [4]

2.5 inductance locus in different current operating point

2.6. d/q differential inductance variation with d/q currents[11]

2.7. Self-sensing control feasible region[45]

2.8. Modulation current trajectory [13].

2.9. Different situations of Modulation current trajectory[13].

2.10 Relationship of differential inductance, cross-saturation inductance and Fundamental current trajectory [13].

2.11 Rotor structure of an Inset magnet motor and an IPM motor [55]

2.12 Permanent magnet dimensions of IPM motor [58]

2.13 Rotor geometry of IPM motor with different dimension of steel bridge and steel ring [59]

2.14 Dimensions optimization of the stator slot structure of [60].

2.15 Comparison between FIIPM machine and conventional IPM machine [69]

2.16 Optimizations of FIIPM based on baseline IPM machine [73]

2.17 Structure of the ringed-pole rotor with the short-circuited copper ring around the magnet [79]

2.18 Magnetic coupling of the rotor ring and stator winding [80]

2.19 Frequency response of the high frequency saliency [80]

2.20 Shor-circuited winding in an IPM machine [88].

3.1. Park transformation from d-q frame to the stationary α - β frame.

3.2. Diagram of the rotating voltage injection self-sensing scheme.

3.3. Detailed demodulation process of high frequency rotating voltage injection position estimation.

3.4. Equivalent average voltages of the reference

3.5. PWM sampling mode with triangular carrier modulation [90]

3.6. Delays introduced by discretizing process.

3.7. High frequency vector angle offset caused by DSP discretization

3.8. Current flow path during device switching in one phase leg of Voltage Source Inverter. (a) Positive load current (b) Negative load current

3.9. Ideal output voltage V_l^* (solid) and distorted voltage shape V_l (dashed) in one PWM period for positive and negative phase current i_l flow, accounting for power device voltage drop (b), turn-on and turn-off delay(c), dead-time (d) and parasitic capacitance effects (e).

3.10. Voltage error estimation procedure.

3.11. Estimated phase voltage error by (solid line) and the fitting line of the error compensation LUT (dashed line).

3.12. Diagram of the self-sensing control scheme with error compensations.

4.1 Schematic of Co-simulation between Matlab and FEM software Magnet

4.2 Phasor diagram of incremental d/q inductance estimation process.

4.3 Stator winding supply circuit for incremental d/q inductance measurement.

4.4 Phasor diagram Equivalent Incremental inductance trajectory estimation

4.5 Equivalent Incremental inductance trajectory

4.6 IPM model for self-sensing performance analysis

4.7 SMPM model for self-sensing performance analysis

4.8 IPM Machine Model FEM results of flux density distribution under different load. (a)

No-load (b) Rated-Load (c) Peak-Load

4.9 SMPM Machine Model FEM results of flux density distribution under different load.

(a) No-load (b) Rated-Load (c) Peak-Load

4.10 Incremental d/q inductance of the IPM machine model

4.11 Saliency ratio of the IPM machine model.

4.12. Saliency Phase shift of the IPM machine model.

4.13. Incremental d/q inductance of the SMPM machine model

4.14 Saliency ratio of the SMPM machine model.

4.15. Saliency Phase shift of the SMPM machine model.

5.1 Machine rotor configuration with the proposed saliency modulation rotor end (a) SMPM rotor (b) IPM rotor

5.2 Stator Winding Configurations: concentrated and distributed [92].

5.3 Stator Winding Configurations: double layer and single layer [92].

5.4. 12/10 and 12/14 combination of fractional slot concentrated winding

5.5. 18/8 Combination of fractional slot distributed winding

5.6. 18/6 Combination of integral slot distributed winding

5.7 Fundamental stator flux reflected in the rotor:12/10-12/14 combination

5.8 Fundamental stator flux reflected in the rotor: Distributed winding

5.9 High frequency stator flux modulation of the rotor saliency.

5.10 Flux path of 18/6 combination. (a) Iq excitation (b) Id excitation

5.11 Flux path of 18/30 combination. (a) Iq excitation (b) Id excitation

5.12 Flux path of 18/16 combination. (a) Iq excitation (b) Id excitation

5.13 Flux path of 12/10 combination. (a) Iq excitation (b) Id excitation

5.14 FFT of the stator flux linkage in stationary reference frame with 1k Hz HF current injection. (a)18/30 combination with integral stator slot distributed winding (b) 18/16 combination with fractional stator slot distributed winding (c) 12/10 combination with fractional stator slot concentrated winding

5.15 Cross section of the prototype motor. (a) 12-slot 14-pole PMSM (b) 12-slot 10-pole rotor end.

5.16. High frequency rotating voltage injection self-sensing scheme.

5.17 Frequency spectrum of high frequency injection operation mode

5.18 Fundamental voltage modulation self-sensing scheme.

5.19 Frequency spectrum of fundamental voltage modulation operation mode

6.1 Cross section of the saliency modulation rotor end per pole pitch.

6.2. Definition of the air-gap function factor K [92]. (a) rotor contour (b) resolver signal strength

6.3. Matlab and FEA joint-simulation process flow chart.

6.4. Schematic of Matlab and FEA co-simulation

6.5. Design optimization process flowchart.

6.6. Bubble Diagram of the optimization result: Air-gap Factor K

6.7. Bubble Diagram of the optimization result: Pole tooth height h1

6.8. Bubble Diagram of the optimization result: Pole slot depth h2

6.9. Bubble Diagram of the optimization result: Minimum Air-gap Lgmin and pole tooth width w1

6.10. Selected rotor structure designs from optimization results. (a) Rotor A (b) Rotor B

6.11.Frequency Spectrum of stator current with high frequency voltage injection at noload using FEA simulation. (a) Rotor A (b) Rotor B

6.12.Frequency Spectrum of stator current with high frequency voltage injection at rated current using FEA simulation. (a) Rotor A (b) Rotor B

6.13.Frequency Spectrum of stator current with high frequency voltage injection at peak current using FEA simulation. (a) Rotor A (b) Rotor B

6.14. Machine rotor with Saliency Modulation Rotor End (SMRE).

6.15. 3D FEA Model of the Prototype machine.

6.16. Schematic of the prototype machine.

6.17. Axial flux distribution of the rotor at no-load operation point. (a) Without the plastic ring (b) With plastic ring

6.18. Flux Density of the rotor and the stator of the rotor end section. (a) rated fundamental current operation (b) two times of the rated current operation (c) three times of the rated current operation.

7.1 Overall structure of the experimental test-rig.

7.2 Myway Inverter Platform (Model: MWINV-9R144)

7.3 PM machine under test and the load DC Machine.

7.4 Block diagram of the Myway inverter platform [94]

7.5 Rotor of the prototype machine.

7.6 Z signal alignment with stator Phase A

7.7. Experimental Result-position estimation current demodulation process

7.8. Experimental Result-Self-sensing estimated position at Motor starting

7.9. Vector Control Loop

7.10. Experimental results- reversed currents response.

7.11. Reverse torque current test.

7.12. Frequency Spectrum of measured stator current with high frequency voltage injection under rated fundamental current when motor speed is 60 RPM.

7.13. High frequency voltage injection self-sensing rotor saliency modulation position signal strength with increased load under different speeds.

7.14. High frequency voltage injection self-sensing rotor saliency modulation position signal strength with increased load under different speeds.

7.15.High frequency injection self-sensing position signal strength comparison of the prototype machine, IPM and Surface Mounted PM Machine.

7.16.High frequency injection self-sensing position signal phase shift comparison of the prototype machine, IPM and Surface Mounted PM Machine.

7.17. Frequency Spectrum of measured stator current with fundamental voltage modulation when motor speed is 2400 RPM.

7.18. Position signal Strength with increasing speed under different load by adopting fundamental voltage excitation.

7.19. Fundamental voltage saliency modulation self-sensing performance with increased load under different speed.(a) Position Signal Strength (b) Position Signal Phase Shift.

LIST OF TABLES

4.1 Main Parameters of the IPM machine

4.2 Main Parameters of the SMPM machine

5.1 Feasible slot/pole combinations for three-phase PM concentrated winding brushless machines[92]

7.1 Parameters of the PM machines under tests. (a) IPM (b) SMPM (c) prototype machine with the saliency modulation self-sensing rotor

В	Friction coefficient
h	Saliency modulation harmonic ratio
i _a	Phase A stator current (A)
i _b	Phase B stator current (A)
i _c	Phase C stator current (A)
i _d	d-axis stator current (A)
i _q	q-axis stator current (A)
ΔI_d	d-axis incremental current (A)
ΔI_q	q-axis incremental current (A)
iα	α -axis stator current (A)
iβ	β -axis stator current (A)
i'_{α}	Shifted α -axis stator current (A)
i'_{eta}	Shifted β -axis stator current (A)
i''_{lpha}	Filtered α -axis stator current (A)
$i_{eta}^{\prime\prime}$	Filtered α -axis stator current (A)
I _{inj}	High frequency injection current amplitude (A)
I_p	Positive sequence current amplitude (A)
I_m	PMSM saliency modulation current amplitude(A)
In	Negative sequence current amplitude (A)
Is	SMRE saliency modulation current amplitude(A)
J	Inertia
k_p	Proportional gain
k _i	Integral gain
Κ	Air-gap function factor

$\overline{K_d}$	Distribution factor
K_p	Coil pitch factor
K_{w10}	Winding factor of 12 slot 10 pole winding
K_{w14}	Winding factor of 12 slot 14 pole winding
K _t	Torque constant
L _d	d-axis stator inductance (H)
L_q	q-axis stator inductance (H)
L _{dq}	Mutual inductance between d/q axis (H)
L_{qd}	Mutual inductance between q/d axis (H)
L _{dinc}	d-axis incremental inductance (H)
L_{qinc}	q-axis incremental inductance (H)
L _{max}	major axis inductance (H)
L _{min}	minor axis inductance (H)
L_M	Mutual inductance between rotor ring and stator coil (H)
L_r	Rotor ring inductance (H)
L _{md}	d-axis magnetizing inductance (H)
L_{mq}	q-axis magnetizing inductance (H)
L _{lkd}	d-axis leakage inductance (H)
L _{lkq}	q-axis leakage inductance (H)
L_{lpha}	α -axis stator inductance (H)
L_{eta}	β -axis stator inductance (H)
Lgmin	Minimum air-gap length
ΣL_s	Sum inductance (H)
ΔL_s	Difference inductance (H)
ΔL_{dq}	Difference inductance of PMSM (H)
ΔL_{rt}	Difference inductance of SMRE (H)

m	Number of phases
Ν	Number of total slots
N _m	pole number of magnet
N _s	pole number of SMRE
Р	Number of poles
p	Differential operator
p	Number of pole pairs
q	Number of slots per pole per phase
R_r	Rotor ring resistance (Ω)
R _s	Stator Resistance (Ω)
SSR	Signal strength ratio
T_e	Electrical Torque (Nm)
T_L	Load Torque (Nm)
Va	Phase A stator voltage (V)
V_b	Phase B stator voltage (V)
V _c	Phase C stator voltage (V)
V _d	d-axis stator voltage (V)
V_q	q-axis stator voltage (V)
Vd	Stator voltage on estimated d-axis (V)
Vq	Stator voltage on estimated q-axis (V)
V_{lpha}	α -axis stator voltage (V)
V_{eta}	β -axis stator voltage (V)
V_h	High frequency voltage amplitude (V)
V_f	Fundamental voltage amplitude (V)
$V_{\alpha ref}$	Reference voltage on α -axis (V)
V_{error_phase}	Voltage error per phase (V)

Z_d	d-axis stator impedance (Ω)
Z_q	q-axis stator impedance (Ω)
$\Delta \psi_d$	d-axis flux-linkage variation (Wb)
$\Delta\psi_q$	q-axis flux-linkage variation (Wb)
ω_e	Fundamental vector electrical angular frequency (rad/s)
ω_h	High frequency vector electrical angular frequency (rad/s)
ω_n	Closed loop system bandwidth
ω_r	Mechanical angular frequency (rad/s)
θ	Mechanical angular rotor position (rad)
θ_{ds}	Saliency phase shift angle (rad)
$ heta_r$	Electrical angular rotor position (rad)
λ_M	No-load PM flux (mWb)
λ_{lpha}	α -axis flux linkage (mWb)
λ_{eta}	β -axis flux linkage (mWb)
ξ	Saliency ratio
ξ _d	Damping factor
ξ_{ab}	Absolute saliency ratio
δ	Air-gap length

Table of Contents

Chapter 1: Introduction	1
1.1- Background	1
1.2- Thesis structure overview	4
Chapter 2: Overview of Self-Sensing PMSM Drive	7
2.1- Vector Control of PMSM Drives	7
2.2- Rotor saliency tracking self-sensing scheme	11
2.2.1 Saliency of PM motor	12
2.2.2 High Frequency Signal Injection	15
2.2.3 PWM Transient Current Response	
2.3- Machine Saturation effects	
2.3.1 Differential inductance variation	
2.3.2 Saliency phase shift resulted by Cross-saturation	
2.4- Self-sensing capability oriented PM motor design	
2.4.1-Motor Design optimization for saliency tracking	
2.4.2-Machine design with intentionally introduced saliency features	
2.5-Summary	
Chapter 3 HF Voltage Injection Saliency Tracking Self-sensing Scheme	
3.1- Inductance saliency modulation of PMSM	39
3.2- High frequency voltage injection and current demodulation process	
3.3- Implementation issues of the high frequency rotating voltage injection self scheme	f-sensing 45
3.3.1 The discretized form of the DSP based processing algorithm	45
3.3.2 Inverter non-ideal aspects	49
3.4- Summary	55
Chapter 4 Self-sensing Performances Analysis of Conventional PM machine	e s 57
4.1- Joint-simulation of Matlab and FEM software	57
4.2- FEM estimation of Inductance Saliency	59
4.2.1 Incremental d/q inductance	59
4.2.2 Equivalent Incremental inductance trajectory	62
4.3- PMSM machines under FEM Analysis	66

CONTENTS

4.4- FEM results of the conventional PM machine saliency characteristics	71
4.5- Summary	76
Chapter 5 Design of the Saliency Modulation Rotor End	
5.1- Concept of Design considerations	
5.2- Fractional slot PMSM with non-overlapping concentrated winding	81
5.2.1 Superiority of the PMSM with fractional slot non-overlapping winding.	82
5.2.2 Selection of slot/pole combination	85
5.2.3 Equivalent stator excitation of 12/10 and 12/14	87
5.3- Comparision of the performances with different winding configuration	88
5.3.1 Stator excitation strength	89
5.3.2 Stator flux reflected in the rotor	91
5.3.3 FEA comparison results	96
5.4- Two Modes of Operation	102
5.4.1 High frequency rotating voltage injection self-sensing mode	104
5.4.2 Fundamental rotating voltage modulation self-sensing mode	107
5.5- Sumarry	108
Chapter 6 Optimization of Rotor End	111
6.1- Set-up of the GA optimization process	111
6.2- GA optimization Result	117
6.3- Selection of Rotor end Structure	121
6.4- 3D FEA Simulation	125
6.5- Summary	129
Chapter 7 Experimental Validation	131
7.1- Structure of the test-rig	131
7.2- Main Components of the Test-rig	134
7.2.1 Measurement and Protection of the Power circuit	134
7.2.2 PM machines for testing	135
7.2.3 Set-up of the Encoder	137
7.3- Self-sensing Operation of Saliency Based Rotor Position Tracking	138
7.3.1 Rotor position estimation	138
7.3.2 Vector control of the Self-sensing Drive	140

CONTENTS

7.4- Enhanced Self-sensing Feasibility of the prototype machine	. 144
7.4.1 Low speed operation with HF voltage injection scheme	. 144
7.4.2 High speed operation with fundamental voltage modulation scheme	. 150
7.5- Summary	. 154
Chapter 8 Conclusion and Discussion	. 156
8.1- Conclusions	. 156
8.2- Future work	. 157
Bibliography	. 159
Appendix	. 164

CHAPTER 1: INTRODUCTION

1.1-Background

The PMSM drives are becoming the main workhorse for the goal of "more-electric" world; they are attracting more and more attention in various applications like automotive traction, marine propulsion and aerospace applications. The main advantages of PMSM machines include high efficiency, simple structure, high power density and high reliability. The invention of vector control theory dramatically boomed the development of AC drives. While it was originally developed for inductions machines, the application of vector control theory on PMSM drive enables a fast and robust control which enables most of the PMSM machine potentials.

The development of digital controllers like digital signal processors (DSP) and Field Programmable Gate Array (FPGA), together with high performance fast-switching power devices like Insulated Gate Bipolar Translator IGBT and SiC mosfet further accelerated the progress of advanced control algorithms. Among them, the most important ones are maximum torque per ampere control (MTPA) and field weakening control (FW). Based on the vector control theory, the actions of these algorithms are mainly related to optimal current/voltage matching for best PMSM drive performances. While the rotor position signal is continuously required to realize the advanced control algorithms, the coordinate transformation and the speed estimation.

Conventional PMSM drives with position sensors require additional cabling, signal processing units and also high standard mechanical joint installation; which are not appropriate in applications with severe environments such as high temperature and strong vibration. The self-sensing control scheme, deriving the rotor position by means of specific machine features is more and more attractive thanks to its advantages including reduced cost, increased reliability and integration of mechanical and power electronic units.

The self-sensing control principles are classified into two modes: machine fundamental model methods and the saliency tracking techniques. The fundamental model based methods are suitable in medium and high speed ranges when sufficient back-EMF is detectable, and rotor position is estimated with various types of tracking observers. However, these algorithms rely on a precise machine mathematical model whereas the machine parameter variations and inverter non-linearity degrade the performance.

Saliency based position tracking techniques, normally using carrier signal injection are proved to be superior for the starting and the low speed operation. The inherent saliency of PMSM are provided either with rotor structural anisotropy for interior permanent magnet (IPM) machines or saturation induced saliency for surface mounted permanent (SMPM) machines; injection signal can be formed either rotating in stationary reference frame, or pulsating in the estimated reference frame. Another type of saliency tracking mechanism is by utilizing the pulse width modulation (PWM) pattern of the voltage source inverter (VSI). For these methods, the rotor position is calculated based on the machine dynamic equation with measured stator current derivatives at specific time instant in consequent PWM sampling periods. However, for these schemes, high speed and high accuracy derivative current sensors are required which are not commonly used in commercial drives. Theoretically, the saliency tracking methods are viable for whole speed range, concerns are commonly addressed relating to the audible noises and machine vibration that could arise due to the injected high frequency signal.

The main disadvantage of machine saliency is that it is heavily affected by machine saturation effect. The variation of the saliency with the increased load introduces rotor position estimation errors; more importantly, the saliency ratio becomes very small under heavy loads when the rotor position signal is undetectable and the self-sensing estimation fails. Hence the saturation effects will result in two major challenges for the self-sensing position estimation: to maintain the machine saliency for position tracking and to compensate the saliency phase shift caused by cross-saturation. For PMSM, the d-axis is aligned with the rotor magnet flux vector and the q-axis is perpendicular to it. For Interior

CHPATER 1: INRODUCTION

Permanent magnet (IPM) machines, the saliency is provided with rotor structural anisotropy as the permeability of the magnet on d-axis flux path is far smaller than that of the rotor iron on q-axis flux path. However, the q-axis inductance is easily affected with increased q-axis current as the stator induced flux on q-axis flux path is very effective; while for the cases of Surface Mounted Permanent magnet machines, the saliency is formed with the saturation on d-axis flux path due to the magnet flux. Although the stator induced flux is less dominant as the effective air-gap is larger, the difference of d/q axis inductances is small for whole current operation range. It is unavoidable that the saliency based self-sensing performances degrades for PMSMs with conventional rotor configurations and the feasibility region of self-sensing position estimation is limited.

Some machine design optimization solutions were proposed aiming at improved selfsensing capability. One idea is to increase the iron space of the q-axis flux path, and hence the q-axis current is less effective to result in saturation, both the saliency ratio and position signal phase shift are then less affected. But the performances are also changed by these modifications of machine structure. There is also a group of designs which introduce rotor saliency features intentionally to improve the PMSM self-sensing capability. It is proved that by mounting a copper ring around the poles, the anisotropy of the machine can be increased thanks to the induced current under high frequency excitation. Nevertheless, losses in such copper rings due to Magnetic Motive Force (MMF) harmonic and high frequency injection can be very high, especially in machines with fractional slot winding or in large machines. Field-intensified IPM (FIIPM) is also claimed to be superior in self-sensing capability by adding flux barriers in the rotor of an IPM. In principle, the saliency of IPM is reversed with the flux barriers on q-axis, which is of great benefit to saliency tracking techniques since motor saturation leads to a larger saliency ratio. However, compromises were made to sacrifice the torque production and power under field weakening operation; also the remaining space for magnets were restricted with the additional flux barriers. The self-sensing oriented design of PMSM machine always conflict with its fundamental performances and so a compromise should be made. Moreover, although the intentionally introduced rotor features increase the feasible region of self-sensing operation, the saturation effects are not completely eliminated since the saliency ratio of ringed-pole machine decreases under heavy load and cross-saturation phase shift is still very large for FIIPM.

In this thesis, the development of a new rotor configuration for better self-sensing capability of PMSM machine is described. The purpose of the design is to provide an additional saliency modulation signal with the modulation frequency different from machine fundamental components. Thanks to this asynchronous modulation characteristic, the injection based self-sensing control is not affected by saturation effects and a self-excited saliency modulation mode is established at higher speed.

1.2-Thesis structure overview

Chapter 2 reviews the existing research outcomes on saliency based self-sensing control. It covers the basic vector control scheme of PMSM drives and the saliency based self-sensing control methods including HF signal injection(current/voltage, rotating/pulsating) and PWM transient current response(INFORM/Fundamental excitation). The machine saturation effects on self-sensing control are discussed in terms of differential inductance variations and cross-saturation resulted saliency phase shift. Some efforts on self-sensing capability oriented PM motor design is also reviewed, this includes the optimization of machine to enhance the self- sensing performance; and also the machines with intentionally introduced saliency features for the position estimation.

Chapter 3 describes the saliency tracking self-sensing algorithm based on the high frequency rotating voltage injection. The inductance saliency modulation mechanism and the current demodulation process by adopting synchronous frame filtering (SRFF) are represented. Some implementation issues of the self-sensing algorithm in DSP controlled VSI platform are analyzed including the discretized pattern of DSP processing, inverter non-ideal behavior. In this chapter, the self-sensing control method used in this project is discussed and efforts have been taken to reduce errors resulted from the control algorithm and inverter.

Chapter 4 analyses the self-sensing performances of conventional PMSMs assisted by FEM simulation. The co-simulation set-up of Matlab and FEM software which significantly simplifies the simulation procedure is described. Two FEM methods of measuring the inductance saliency are introduced as the incremental d/q inductance measurement and the equivalent incremental inductance trajectory estimation. The methods are used to derive the saliency profile of an IPM machine and a SMPM machine with different current levels. The weakness of conventional PMSM in self-sensing position estimation is discussed based on simulation results.

Chapter 5 introduces the novel design of saliency modulation self-sensing rotor end, the primary design aspects together with the slot/pole combination selection considerations are discussed. The aim is to achieve strong stator flux excitation for both the PMSM rotor and the rotor end to guarantee torque production as well as saliency modulation signal strength. The self-sensing performances of SMRE with different winding configurations are compared. The optimal slot/pole combination of PMSM machine and the SMRE is derived in this chapter. In addition, the two operation modes for the novel design are introduced based on rotating voltage injection. For low speed operation, high frequency voltage injection scheme is adopted similar to the conventional scheme; at higher speed, a specific saliency tracking without superposed signal naming as the fundamental voltage modulation is described.

Chapter 6 describes the optimization process of the rotor end for optimal self-sensing performance. The set-up of the optimization process with genetic algorithms (GA) together with FEM simulation is introduced. The optimization aim is for better position signal quality in terms of larger signal strength and less total harmonic distortions (THD). The GA optimization results are presented and considerations for the selection of the rotor end structure are discussed. A 3D FEM analysis is also adopted that the axial magnet flux leakage through SMRE is avoided.

Chapter 7 presents the experimental results of the self-sensing performances. The test-rig

CHPATER 1: INRODUCTION

for the experiment tests is described. Three PMSM machines are used during the tests: the IPM and the SMPM machines analyzed in Chapter 4 and the prototype machine with SMRE. Self- sensing operation of saliency based control is examined, the enhanced self-sensing feasibility of the new design is proved by comparing its self-sensing capability with the conventional PMSM machines under the same evaluation standard

Chapter 8 concludes the thesis and provides suggestions for future research.

<u>Chapter 2: Overview of Self-Sensing</u> <u>PMSM Drive</u>

2.1-Vector Control of PMSM Drives

The principle of vector control theory, proposed by Felix Blaschke has been developed for the wide spread implementations in the control algorithms of high performance AC drives [1]. With the similarity to the control of DC drives, Vector Control, or Field Oriented Control ensures the dynamic performance of the AC drive system with its capability of the fast torque response. Taking the advantage of this, vector control theory becomes a standard of high performance AC drives in servo and traction applications for both induction motor and synchronous motors like permanent magnet synchronous motor (PMSM).

Based on the transformation between 3 phase vector and rotating direct/quadrature (d/q) reference frame, the principle of vector control is to transform the time dependent variables into a time invariant system [2]. For PMSM, the d-axis is aligned with the rotor magnet flux vector, and by adjusting the torque current iq and field current id, the control of instantaneous electrical quantities as well as the torque and the speed of the motor, both in transient and steady state, are realized by vector control with appropriate closed control loop (torque/current, speed, position) with respect to different applications [3].

In Fig 2.1, the two steps of the Park transformation from 3 phases to rotating d/q reference frame is illustrated. Any space vector of the motor (current, voltage, flux) can be represented in the d/q frame. Taking the stator current as an example, the resultant 3 phase current vector i_s is first represented in two phase stationary coordinates α and β , the value of the current components of $i_{s\alpha}$ and $i_{s\beta}$ is represented by (2.1), this is termed as 3 phase to α/β or Clark transformation (3/2); with the instantaneous value of the rotor angle θ_r , the current vector is then represented in rotating coordinates d/q frame and the value of current i_{sd} and i_{sq} is calculated as (2.2) and the transformation



Fig 2.1 transformation from 3 phase to d/q rotating reference frame.

is termed as α/β to d/q [4]. For PMSM, mostly, the d-axis is defined to be aligned with the rotor magnet flux, and the flux is controlled by the current component i_{sd} , the q-axis is perpendicular to d-axis and the current component i_{sq} is controlling the torque, the d/q coordinates is defined to rotating at the angular speed of ω_e which is synchronous with the rotor in electrical rotating speed.

$$i_{\alpha}(t) = \frac{3}{2}i_{a}(t)$$

$$i_{\beta}(t) = \frac{\sqrt{3}}{2}i_{b}(t) - \frac{\sqrt{3}}{2}i_{c}(t)$$
(2.1)

$$i_{d}(t) = i_{\alpha}(t) \cos\theta_{r} + i_{\beta}(t) \sin\theta_{r}$$

$$i_{q}(t) = -i_{\alpha}(t) \sin\theta_{r} + i_{\beta}(t) \cos\theta_{r}$$
(2.2)

The motor is seen as d/q system in the control process, while to realize the excitation of the motor by power inverter, it is required to transform the d/q values back into 3 phase values, the inverse transformation of the voltage vector from d/q to α/β and α/β to 3 phase

is given by (2.3) and (2.4) [4], the value of instantaneous rotor position θ_r is again necessary to realize the process.

$$V_{\alpha}(t) = V_{d}(t) \cos\theta_{r} - V_{q}(t) \sin\theta_{r}$$

$$V_{\beta}(t) = V_{d}(t) \sin\theta_{r} + V_{q}(t) \cos\theta_{r}$$
 (2.3)

$$V_{a}(t) = \frac{2}{3}V_{\alpha}(t)$$

$$V_{b}(t) = -\frac{1}{3}V_{\alpha}(t) + \frac{1}{\sqrt{3}}V_{\beta}(t)$$

$$V_{c}(t) = -\frac{1}{3}V_{\alpha}(t) - \frac{1}{\sqrt{3}}V_{\beta}(t)$$
(2.4)

As shown in Fig 2.2, the coupled equivalent circuits for the d and q axis of PMSM is illustrated, the instantaneous value of d/q axis voltages and currents and their relations are shown in (2.5) where L_d and L_q represent the inductances on d and q axis of the motor and the flux-linkage of the stator winding sourced by permanent magnet is represented by $\lambda_{\rm M}$. In equation (2.5), $\omega_{\rm e}\lambda_{\rm M}$ describes the induced voltage by rotating magnet on q-axis which is termed as back EMF; the term $\omega_e i_q L_q$ and $\omega_e i_d L_d$ describes the cross coupling voltage component by current between d and q axis [5]. Equation (2.6) describes the electromagnetic torque of the machine where P is the number of poles, it is seen that, neglecting the variations of the inductance caused by iron saturation effects, the output torque of a PMSM can be simply controlled by i_d and i_q . The term $i_q \lambda_M$ describes the magnetic alignment torque component, which is related to the value of i_q ; the term $(L_d - L_q)i_d i_q$ describes the reluctance torque component, which is related to both i_d and i_q . For PMSM with saliencies, that is the d-axis inductance L_d is different from that of q-axis L_q (normally $L_q > L_d$), i_d is preferably controlled to be negative to achieve maximum torque, for a given current, a maximum torque angle or a current trajectory is used to achieve maximum torque per ampere (MTPA) control [6].



Fig 2.2 coupled equivalent circuits for the d and q axis of PMSM.

$$V_{d} = i_{d}R_{s} + L_{d}\frac{di_{d}}{dt} - \omega_{e}i_{q}L_{q}$$

$$V_{q} = i_{q}R_{s} + L_{q}\frac{di_{q}}{dt} + \omega_{e}i_{d}L_{d} + \omega_{e}\lambda_{M}$$
(2.5)

$$T_e = P * \frac{3}{2} (i_q \lambda_M + (L_d - L_q) i_d i_q)$$
(2.6)

In Fig 2.3, the block diagram of the PMSM drive vector control system is shown. The basic control structure of speed loop and torque/current loop is included. In the speed loop, the torque demand represented by I_{dq}^* is derived by basing on the difference between ω_e^* , the reference speed, defined by user or higher potion loop in servo application and the machine rotor speed ω_e derived by differential of the electrical rotor position. In the current loop, the machine torque is controlled by adjusting the current i_{sd} and i_{sq} , where the reference voltage V_{dq}^* is calculated with the error between I_{dq}^* and the measured stator current I_{dq} by PI regulators. In the control loop, the motor is taking into d/q system, 3 phase to d/q and its inverse transformation is adopted to feedback current and command reference voltage. For PMSM drive system, the control algorithms are implemented by microprocessors like Digital Signal Processor (DSP) or Field —



Fig 2.3 Block diagram of vector control system of PMSM drive.

Programmable Gate Array (FPGA). The calculated 3 phase reference voltages are transformed into switching duty cycles by Pulse Width Modulation (PWM) to drive the voltage source inverter. To realize the closed loop control of the drive system, 2 or 3 stator phase currents are measured by current sensors; another important parameter to realize the control function is the rotor position which is measured by a position sensor in conventional PMSM drives.

2.2-Rotor saliency tracking self-sensing scheme

As the back-EMF component of PMSM motor becomes insufficient for the model based rotor position estimation, the techniques of saliency tracking self-sensing operation have been attracting significant attentions by various research groups. By tracking the anisotropic property of the machine, self-sensing scheme becomes achievable for low and even zero speed; also, the estimation error introduced by the variation of motor mathematical model, mainly the change of stator resistance of motor under temperature variations is avoided in comparison with model based self-sensing scheme. The anisotropic property of PMSM machines is mainly expressed as stator inductance difference between L_d and L_q , the saliency can be classified into structural saliency arising from the rotor geometric structure and saturation saliency due to the magnetic



Fig 2.4 Examples of PMSM rotor structure (a) Surface Mounted (b) Inset (c) Interior [4]

saturation caused by the magnet flux on d-axis flux path. Some designs of PMSM rotor intentionally introduce inductance variations in the stator winding, this will also be one of the focus of the research project.

2.2.1 Saliency of PM motor

The structure of a salient rotor causes a difference of the d and q axis flux path reluctance, as the permeability of permanent magnet is close to unity, the d-axis flux path through the magnets passes longer effective net air-gap compared to that of the q-axis flux with longer iron path, the result is the lager stator q-axis inductance L_q in comparison with L_d . The 3 basic rotor structures of PMSM motors are as shown in Fig 2.4. Fig 2.4 (a) is the type of surface mounted PM Machine, for this type of rotor, the structural saliency is very small as a constant effective air-gap is seen by both d and q axis flux. Fig 2.4 (b) is the type of inset PM machine with the magnets placed in rotor shallow slots, L_q is larger than L_d for this type of rotor as the q-axis flux passes through a longer iron path than that of d-axis flux. In Fig 2.4 (c), the rotor of Interior PM machine is shown, the magnets are placed inside the rotor to strengthen the rotor mechanical construction, structural saliency for this type of rotor is very large and this feature makes the IPM a proper candidate for both self-sensing control and wide power range application.

The d/q axis synchronous inductance can be expressed as (2.7) [7]:

$$L_d = L_{md} + L_{lkd}$$

$$L_q = L_{mq} + L_{lkq} \tag{2.7}$$

where the inductance L_d and L_q are combined of the magnetizing inductance L_{md} and L_{mq} and the leakage inductance L_{lk} . L_{md} and L_{mq} are directly related to the permeability of the path for main flux-linkage produced by the coils, the synchronous magnetizing inductances are resulted by motor structural dimensions, L_q is larger than L_d for most of salient type rotors [8]. The leakage inductance is correlated to the leakage flux at inside slots, tooth tip and around end winding. Both the magnetizing and leakage inductance are affected by iron saturation caused by the magnet flux, for the case of surface mounted PM machine with inefficient structural saliency, the saturation inductance saliency caused by saturation effect is utilized for position tracking [9].

The inductance in stationary reference frame can be expressed as (2.8) [10]:

$$\begin{bmatrix} L_{\alpha} \\ L_{\beta} \end{bmatrix} = \begin{bmatrix} \Sigma L_{s} - \Delta L_{s} \cos(2\theta_{r}) & -\Delta L_{s} \sin(2\theta_{r}) \\ -\Delta L_{s} \sin(2\theta_{r}) & \Sigma L_{s} + \Delta L_{s} \cos(2\theta_{r}) \end{bmatrix}$$

$$\Sigma L_s = \frac{L_q + L_d}{2}, \, \Delta L_s = \frac{L_q - L_d}{2} \tag{2.8}$$

where θ_r is the rotor position in electrical angle and the inductance saliency modulation frequency is twice of the rotor electrical frequency as the machine saliency is symmetrical between positive and negative d-axis.

In figure 2.5, the inductance equivalent locus in different current operating point shows the variations of the synchronous stator inductance L_d and L_q in accordance to i_d and i_q . These variations are caused by the iron saturation of the generated winding flux-linkage with increasing stator current, the rotor geometry structure is also affects this

CHAPTER 2

phenomenon [11]. For IPM motor with salient rotor, the inductance L_q is larger than L_d for no load as expressed by the current point i_{dq1} in figure 2.5, this coincides with the above analyses on structural saliency where the saliency ratio is related to the rotor geometry. For proper PM motor design, the iron core of the motor is limited to be small as possible for consideration of overall weight and cost, and hence with the increasing torque current i_q , the q-axis inductance L_q decreases significantly due to the saturated iron path of the exciting flux-linkage [12]. For the current point of i_{dq2} in Fig 2.5, the saliency becomes insufficient to detect when L_q is equal to L_d : this situation usually occurs when current working point moves near to the rated current level and saliency tracking self-sensing scheme fails in this situation. A peak current is sometimes required, this is illustrated by operation point i_{dq3} in Fig 2.5 where the q-axis inductance is further saturated thus reverse the saliency. Saliency tracking in this case is theoretically effective, nevertheless it requires an adaptive technique and the algorithm is more complicated in



Fig 2.5 inductance equivalent locus in different current operating point

accounting for parameter uncertainty. Another situation due to the motor cross saturation effect is shown as the current point i_{dq4} . The rotation of the inductance equivalent locus is caused by cross-saturation effect, this will further lead to the deviation of rotor estimation θ_{err} [13]. The error needs to be compensated in self-sensing algorithms especially for the current operation point when negative d-axis current is introduced to utilize the reluctance torque of salient PM motor.

2.2.2 High Frequency Signal Injection

High frequency signal injection scheme is based on the superimposition of a test signal on the fundamental supply. The test HF signal is modulated by the saliency of PMSM machine and the rotor position can be derived by the demodulation of the measured HF response. The test injected signal can be either current [14] or voltage [15] and the corresponding voltage or current response is demodulated. Frequency of the injection is selected to vary from several hundreds to several thousand Hz, normally higher than the current loop bandwidth so as to reduce the torque ripple introduced by injection; as a result, voltage injection is usually preferred due to the limitation of current control bandwidth which makes the high frequency injection more difficult to achieve with cost effective inverters.

One of the voltage injection techniques is by applying a high frequency rotating voltage vector on the fundamental voltage reference in stationary α/β reference frame [16]. For easy calculation of stator currents, the voltage vector with constant amplitude V_h and the frequency of ω_h can be expressed as:

$$\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} = V_h \begin{bmatrix} -\sin(\omega_h t) \\ \cos(\omega_h t) \end{bmatrix}$$
(2.9)

By neglecting the effect of the stator resistance as the high frequency modeling of the machine becomes inductive, the current response due to the rotating voltage injection by putting the inductance expressed in (2.8) into the equation is as (2.10) [17]:

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = I_{p}e^{j\omega_{h}t} + I_{n}e^{j(2\theta_{r}-\omega_{h}t)} = I_{p}\begin{bmatrix} \cos(\omega_{h}t) \\ \sin(\omega_{h}t) \end{bmatrix} + I_{n}\begin{bmatrix} \cos(2\theta_{r}-\omega_{h}t) \\ \sin(2\theta_{r}-\omega_{h}t) \end{bmatrix}$$

$$I_p = \frac{\Sigma L_s V_h}{\omega_h (\Sigma L_s^2 - \Delta L_s^2)} I_n = \frac{\Delta L_s V_h}{\omega_h (\Sigma L_s^2 - \Delta L_s^2)}$$
(2.10)

In equation (2.10), the high frequency current response includes a positive sequence and a negative sequence component with the amplitude I_p and I_n defined by the injection voltage amplitude V_h , the injection frequency ω_h and the stator inductance ΣL_s and ΔL_s as defined in (2.8). Several methods are proposed for the demodulation process of the current response to extract the rotor position angle θ_r in the negative sequence component [18, 19]. A heterodyne demodulation scheme is used in [17] to separate the rotor position signal from the high frequency current response and the fundamental excitation signals. Another option is by introducing the synchronous frame filtering process [20]. The measured stator current with all frequency components is first shifted by the injection frequency ω_h that the current is synchronous with the injection voltage vector; the negative sequence component is turned into DC term and the rotor position angle can be derived by low pass filtering and an atan equation [21]. Details of the procedure will be illustrated in Chapter 3.

Another approach termed as synchronous frame testing signal injection is realized by a pulsating carrier signal injected along an estimated synchronous axis [22, 23]. The oscillating voltage signal is injected on estimated d-axis in most conditions; it can be also applied on q-axis [24], but the torque ripple is larger. For d-axis injection, the following vector on estimated axis is applied on fundamental excitation:

$$\begin{bmatrix} \widetilde{Vd} \\ \widetilde{Vq} \end{bmatrix} = V_h \begin{bmatrix} \sin(\omega_h t) \\ 0 \end{bmatrix}$$
(2.11)

For position estimation, the current response on the estimated q-axis is utilized as it is proportional to the angle deviation of the estimated axis to the real d and q axis of the machine. Such condition is assured because of the saliency that L_d is different from L_q , thus the position tracking algorithm is continuously adjusting the estimated axis angle so that the rotor angle is estimated indirectly when the current response on estimated q-axis approaches zero[25, 26].

In[27], a measurement frame rotates by 45° from the estimated d and q axis is proposed for self-sensing control of induction machine. As the inductance saliency is assumed symmetrical along d and q axis, the amplitude of the pulsating current response on the orthogonal axis should be the same when the estimated frame is coincided with the machine d/q frame. Thus the measured amplitude difference can be used as the position error for the estimation of the rotor position by a tracking observer. However for the case of PM machine, due to magnetic saturation on the d-axis, the current response of positive and negative d-axis components will be different, this will introduces estimation errors of the position angle.

The high frequency carrier signal injection methods are relatively simple to apply on DSP controlled voltage source inverters as the algorithms are based on superposition of the carrier signal on fundamental excitation and the demodulation algorithm is applied in a regular sampling process. However, both the rotating voltage injection and synchronous frame testing signal injection are influenced by several non-ideal physical attributes which may degrade the self-sensing performance significantly. The effects have been studied in [28] for rotating injection and for pulsating signal injection [29]. There are two major causes of the position estimation error, one is termed as inverter non-linearity [30], and the other one is the tracked saliency variations caused by machine load dependent saturation [11]. The comparison between the two carrier injection methods of the non-ideal effects on their performance is studied in [31]; in [32] the comparison is focused on the effects introduced by saturation induced saliencies.

2.2.3 PWM Transient Current Response

Another saliency tracking technique termed as PWM current transient response is introduced as the machine is driven by PWM voltage source inverter, which provides the intrinsic high frequency excitations to exploit the saliency. This technique can be classified into two types, one is the INFORM (Indirect Flux detection by Online Reactance Measurement) method based on the use of additional test pulses [33]; the other one is the fundamental PWM excitation scheme that the voltage vector applied to the machine remains unchanged [34]. Saliency tracking is achieved for both methods by measuring the current derivate at specific sampling point.

The INFORM method as descried in [35] uses additional test pulses to achieve the saliency tracking. The voltage vectors applied on the machine during one PWM period are defined as six active vectors and two null vectors when all the upper or lower switches are on. For one sample of rotor position, three pairs of test pulses are applied during the null vector in order to cancel the effects of back EMF and stator resistance voltage drop: the pulses in each pair have equal durations but opposite directions so that the overall average output voltage remains the same [36]. The stator current transient response of the machine for each phase is related to the stator voltage V_s , the stator resistance R_s , the phase inductance L_s and the flux-linkage induced voltage E_s as expressed:

$$V_s = i_s R_s + L_s \frac{di_s}{dt} + E_s \tag{2.12}$$

The stator voltage V_s is the average output voltage of the inverter which is depending on the switching state and the DC link voltage. Assuming the DC link voltage being constant and the components of stator resistance voltage drop and back EMF is cancelled by three pairs of test pulses, the stator inductance can be solved by measured current derivate. The self and mutual inductances can be expressed as follows [37]:

$$L_{aa} = L_0 + L_1 cos(2\theta_r)$$

$$L_{bb} = L_0 + L_1 \cos(2\theta_r + \frac{2}{3}\pi)$$

$$L_{cc} = L_0 + L_1 \cos(2\theta_r - \frac{2}{3}\pi)$$

$$L_{ab} = L_{ba} = \frac{1}{2}L_0 + L_1 \cos(2\theta_r - \frac{2}{3}\pi)$$

$$L_{ac} = L_{ca} = \frac{1}{2}L_0 + L_1 \cos(2\theta_r + \frac{2}{3}\pi)$$

$$L_{bc} = L_{cb} = \frac{1}{2}L_0 + L_1 \cos(2\theta_r + 2\pi)$$
(2.13)

where L_0 is the average value and L_1 is the amplitude of inductance variations related to the electrical rotor position angle θ_r . For low speed operation, by assuming that the back emf, resistance voltage drop and rotor position remains unchanged during one PWM period, the rotor position θ_r is solved with the phase inductance by compensating the voltage components and measuring current derivative values realized by the test pulses [38].

Another technique known as fundamental PWM excitation is applied for saliency tracking without additional test pulses. In comparison of the INFORM method, by integrating the transient excitations into fundamental PWM vectors, the harmonics introduced by high frequency test signals are eliminated so that the scheme is superior in terms of torque ripple and audible noise [39]. The scheme is based on measuring the modulation of current derivative due to the inductance saliency at different excitation voltage vectors in PWM periods.

Theoretically, the approach is able to work for full speed range of the PM machine. The current derivatives are sampled under specific null and active voltage vectors, due to the current oscillations after the instant of switching; the sampling point should be set when the stator current stop oscillating. The resulting current derivatives are then calculated to compensate the back EMF and the resistance voltage drop, rotor position angle is derived with the inductance modulation [40].

The INFORM method is only applicable for low speed operation of self-sensing scheme in theory as the null vector in the PWM switching period becomes narrower for high speed operation of the PM machine. The opposite pulse for the compensation of the testing pulses are required to be applied for the same time duration to ensure the same amount of the output voltage vector, hence a minimum pulse width is limited that the three pairs of test pulses are not always applicable in one PWM period. For full speed range, combination of the model based self-sensing scheme for higher speed operation is proposed in [41]. The additional testing pulses of the INFOM method also introduce additional distortion to the current so its total harmonic distortion is increased: this leads to a higher torque ripple and audible noise, and the switching losses are increased as well. The lifetime of the machine is also affected since the stator winding voltage stress increases up to two times of the DC bus voltage due to the fast changing voltage at terminals [42].

The main challenge for fundamental PWM excitation is termed as narrow vector problem, when the applied voltage vector pulse width is not enough for the derivative measurement before the current oscillations settle down during PWM fundamental excitation, the extended modulation method is used [34]. Nevertheless, the PWM transient current response technique, both in the IMFORM method and in the fundamental PWM excitation, needs to alter the normal pattern of PWM excitation and sampling, which complicates the algorithms and increase the required computing power of the DSP; in addition, high standard differential current sensors are necessary for fast, accurate and linear sensing to guarantee the performance of the self-sensing scheme [43].

2.3- Machine Saturation effects

The main challenges when adopting the saliency based self-sensing scheme are related to machine saturation effects under loaded operation. This is due to the fact that the driving current is applied mainly on the q-axis, which is perpendicular to the direction of magnet flux (d-axis) in the rotor rotating reference frame. The most severe situation always
occurs when a peak torque current is required. One of the challenges is to maintain a proper saliency along the current trajectory; another is related to the angular error caused by saliency shift referred to as cross-saturation [13]. Some terms have been used to define the self-sensing operation region, they are: anisotropy ratio [44], feasible region [45] and sensorless safety operation area (SSOA) [46]. However, the current operation point is restricted to guarantee the self-sensing performance. Analyses of cross-saturation error are also widely investigated [47, 48], some solutions are provided regarding the machine design optimization [49-51], and compensation techniques in the control algorithms [52, 53]. However, the procedure of modelling the cross-saturation error for compensation is complicated and a Look-Up-Table (LUT) is necessary for each machine [53]. In addition, the self-sensing oriented optimizations, such as thinner magnets and thicker iron bridge size [50], changes the original designs allowed to obtain optimal machine general performances [51]. Unfortunately, with conventional PMSM rotor configuration, there is no specific technique eliminating the saturation effects on self-sensing control completely to avoid differential inductance disappearing and to ensure linear saliency shift.

2.3.1 Differential inductance variation

The saliency of the machine under each current operation point is directly related to the differential inductances, which is the equivalent of incremental inductances on synchronous reference frame [11]. As shown in Fig. 2.6, the q-axis and d-axis differential inductances of an IPM machine are shown as the curves A and curves B versus the q-axis and d-axis currents respectively, in each group of curves, the variations of the inductance caused by the currents on opposite axis (id, iq) are also shown. It can be noticed that the q-axis inductance decreases significantly with increased q-axis torque current, this is the inevitable result of the machine saturation effect that the q-axis flux saturates the corresponding flux path. In terms of the d-axis inductance, the variation is however much smaller, as the d-axis reluctance is large due to the existence of the magnets and the no-load d-axis flux path is nearly saturated caused by the magnet flux. Due to these differential inductance characteristics, there is a region of the two axis are very close to each



Fig 2.6. d/q differential inductance variation with d/q currents[11]

other. It can be also noticed that the cross coupling effects lead to further variations of the differential inductances especially at low d/q axis currents. At high current values when the machine is saturated, cross-coupling effects are less dominant [11].

To define the feasible region of the self-sensing operation, the differential inductances of the two axis are ploted in the d/q current plane as shown in Fig.2.7. The q-axis and d-axis inductances are shown as the dashed line and the solid line respectively. The maximum current vector amplitude is also expressed as the dotted circle. It is noticed that the q-axis inductance dramatically decreases with increased q-axis current; the d-axis inductance is also decreased with increased value of d-axis current (negative value) while the variation is relatively small. The curve L_{dif} is defined by a bold line when the inductances are equal to each other. For this situation, the saliency based self-sensing scheme fails to detect rotor position and the feasible region is defined for the current points not on this curve or close to it [45].



Fig 2.7. Self-sensing control feasible region[45]

2.3.2 Saliency phase shift resulted by Cross-saturation

Apart from inductance variations on the respected d and q axis, the machine inductance saliency is also affected by cross-saturation effects that the angle of the equivalent inductance trajectory is shifted [13]. As shown in Fig.2.8, for general current operating points, the high frequency current trajectory caused by rotating voltage injection is an ellipse. It can be noticed that for most situations, the maximum and minimum current ΔI_{max} and ΔI_{min} defined by the major and minor axes of the ellipse is not aligned with the q and d axes where the differential currents ΔI_{qh} and ΔI_{dh} are measured. Hence, the phase shift angle ϵ is defined for the deviation between the estimated ellipse and the real rotor angle.

The saliency profile represented as the HF current trajectory and also the phase error is directly related to the values of d/q inductances L_d and L_q and the cross saturation



Fig 2.8. Modulation current trajectory [13].

inductance L_{dq} , and these values vary with different current operating points [13]. Three typical situations of the HF current trajectory are illustrated as in Fig 2.9. For the first case, the cross saturation inductance is equal to zero, this is the optimal situation as the phase shift angle ϵ is zero that the major and minor ellipse axes are aligned with the d and q axes and a strong inductance saliency remains. For the second situation, the differential inductances L_d and L_q are equal to each other while a saliency is still detectable due to cross saturation. The current trajectory ellipse is shifted by 45°. The worst case occurs when both L_d and L_q and L_{min} and L_{max} are equal to each other: in such case the HF current trajectory becomes a circle for the last situation in Fig 2.9. Under such situation, rotor position detection is impossible [13].

Shown in Fig.2.10, the curves of $L_{dif} = 0$, $L_{dq} = 0$ and the MTPA trajectory are plotted for and IPM machine based on FEM simulation. The critical point at the cross of the curve $L_{dif} = 0$ and $L_{dq} = 0$ defines current operating point at which self-sensing detection fails, corresponding to the last situation in Fig.2.9 when the HF current trajectory becomes a circle. Normally, the position signal becomes too small to detect when the fundamental current is operating close to the critical point. Unfortunately, it is

2



Fig 2.9. Different situations of Modulation current trajectory[13].



Fig 2.10 Relationship of differential inductance, cross-saturation inductance and Fundamental current trajectory [13].

preferable that the current move close to this point along the current trajectory for more torque production but the self-sensing performance is affected [13].

2.4-Self-sensing capability oriented PM motor design

As discussed previously, the high frequency injection self-sensing scheme strongly relies on the saliency of the machine: the saturation effect is of great impact on this concept [11-13]. The major concern of machine under loaded condition is the zero saliency situations when the incremental inductances on d and q axis are equal, making the saliency tracking self-sensing scheme infeasible. Common approach in machine design procedure is to shift the zero saliency operating point outside the envelope of the current operation range. However, the zero saliency condition occurs at start-up of the machine when a current lager than rated is often used for higher acceleration.

The design of electrical machine is always applied with several design specifications to be met, the self-sensing capability oriented design can be delivered under similar pattern but the additional enhancement of saliency characteristics should not degrade the fundamental performance of the machine. The research on design optimizations to guarantee both the self-sensing capability and good machine design has been explored in recent years, highlighting that, compromises are always required. An alternative approach is developed focused on novel rotor design features. One design is termed as the fieldintensifying IPM machine where the inductance on d-axis is always larger than that of qaxis, so the zero saliency condition is avoided under load conditions; another design is by increasing the high frequency response of saliency by applying copper coils in the rotor: the so called ringed-pole machine is developed to enhance the self-sensing feasibility region.

2.4.1-Motor Design optimization for saliency tracking.

A lot of efforts have been undertaken to enhance the self-sensing capability in the machine design procedure, the inductance saliency is optimized to guarantee the position estimation under heavy load operation and minimize the estimation error introduced by cross-saturation effect.

The geometrical structure of rotor is of significant impact on machine inductance: in [54, 55], the performance is compared between an inset magnet rotor as shown in Fig 2.11 a and an IPM rotor with multiple flux- barriers as in Fig 2.11 b. The two motors with identical stators were tested with high frequency voltage signal injection, using both Finite Element Analysis (FEA) and experimental test demonstration: the Inset magnet motor is proved to be better with wider operation range even at heavy load. The advantage of the inset magnet motor structure is the larger iron path of the rotor, while the iron bridges are more easily saturated by excitation load current for the case of IPM. It is also pointed out that in consideration of the main flux paths in the motor, it is preferable



Fig 2.11 Rotor structure of an Inset magnet motor and an IPM motor [55]

that the stator saturates before the rotor so that the saliency remains and the rotor position signal remains detectable.

The self-sensing capability with respect to permanent magnet dimensions is studied in [56-58] as well as the location in rotor laminations. As shown in Fig 2.12, the optimization is based on the parameters defined for permeant magnet as the width w_m , the thickness l_m , the distance of depth from the rotor surface d_r and the span of the bridge θ_{rt} [58]. Finite elements analysis was used to evaluate the saliency ratio and the



Fig 2.12 Permenant magnet diemensions of IPM motor[58]

optimization was applied for better self-sensing capability without sacrificing significantly the fundamental performance of machine in terms of average torque and torque ripple. It can be noted that for better torque production, the size of permanent magnet is preferred to be larger; however, the iron path becomes narrower and the cross saturation becomes worse, so this is not favorable for the sake of self-sensing control.

The effect of different dimensions of the rotor steel bridge between adjacent magnets and the steel ring as shown in Fig 2.13 on saliency characteristics of the motor and self-sensing capability are studied in [59]. With numerical simulations and experiments, it was proven that comparing with the baseline rotor as shown in Fig 2.13 a, the saliency ratio and the position estimation accuracy is improved by increasing the size of steel bridge as in Fig 2.13 b. Furthermore, the position estimation error is decreased with thinner rotor ring near the air gap as demonstrated by the motor in Fig 2.13 c. This is similar by shorten the magnet width and increasing the size of steel bridge, the iron path on q-axis is wider and hence the q-axis inductance is increased without the change of the inductance on d-axis. By increasing the size of steel ring, both the inductances on d and q axis are



Fig 2.13 Rotor geometry of IPM motor with different dimension of steel bridge and steel ring [59]

increased yet the d-inductance increases more. This is because of the increasing of leakage flux between adjacent permanent magnets with thicker steel ring, comparing with the thinner ring where the flux path of d-axis is more saturated. Furthermore, the leakage of the magnet flux in the ring reduces the total amount of flux penetrates the stator winding, which leads to a lower output torque of the machine.

Attention has been also put into analyzing the influence of stator slot profile on selfsensing capability [60]. As shown in Fig 2.14, the optimization was carried out focusing on the size slot opening and tooth tip. The investigation in [60] indicates that a reduction of these two sizes results in a smaller estimation error. However, the author suggested that a generalized optimization is preferable, taking into considerations of the stator structure by using these parameters. The analysis results for different type of motor proves that the saliency characteristics depend on the rotor configuration so the slot structure effects cannot be easily defined.

The effects of tooth tip were also studied in [61]: in this paper, it is also proposed that the self-sensing capability can be significantly changed by closing the slot with iron bridges, the harmonics and the signal to noise ratio were taken into account in this research. The authors in [61] also state that it is not as easy to change the saliency profile by changing the stator only as if by changing the rotor. Nevertheless, self-sensing capability oriented optimizations of the machine design are now applied more comprehensively considering



Fig 2.14 Dimensions optimization of the stator slot structure of [60].

the machine overall performance. The optimization can be carried out similarly to that of a machine design with general specifications using genetic algorithms [62]: the capability of self-sensing can be defined as several objectives with particular requirements [63]. Recently, more and more efforts are put into research on the area of self-sensing oriented design of machine and the process is challenging to balance with the fundamental machine performance [64, 65].

2.4.2-Machine design with intentionally introduced saliency features.

Different from that of the conventional PM machine rotor structures, a special type of design is introduced in [66-68] for improving the field wakening performance. As shown in Fig 2.15, flux barriers with very high reluctance are located on the flux path of q-axis in the rotor of these types of machine, as a result, the inductance saliency is reversed that L_q is smaller than L_d even at no-load. In terms of the control strategy, as the behavior of the saliency becomes negative during whole operation range, a positive d-axis current is applied for the utilization of the reluctance torque to achieve maximum output torque as



Fig 2.15 Comparison between FIIPM machine and conventional IPM machine [69]

CHAPTER 2

shown in Fig 2.15 a. The positive d-axis current strengthens flux field rather than that of the negative d-axis current for conventional IPM as in Fig 2.15 b, where the magnet field is weakened, hence the name flux-intensifying IPM (FIIPM). It is claimed by the inventors of FIIPM that this type of machine suffers less armature reaction effects and the demagnetization performance is better than conventional IPM machines. Thanks to the reversed saliency, the stator MMF is shifted with positive d-axis current component that trying to increase the magnetization: as a result, the area of permanent magnet under opposite stator excitation direction is reduced, in addition, the demagnetization effects caused by q-axis current component is also decreased with smaller inductance on q-axis. It is stated in [70] that by using FIIPM, the rare earth material can be saved with smaller permanent magnet volume and low-coercive force magnets. The design of FIIPM is further optimized in [69]. A leakage path between adjacent magnets for variable flux is introduced to extend the operation speed range of FIIPM machine [71]. For low speed high torque operation, the path is highly saturated by q-axis current and most of the magnet flux penetrates the winding coil for torque production; for case of high speed operation, as the torque current is reduced, some portion of the magnet flux is "shortcircuited" through the leakage path that the back emf is reduced. A design of FIIPM for outer rotor application is introduced in [72] with better flux-weakening capability.

Another benefit of FIIPM is the extraordinary self-sensing capability compared to that of normal IPM machines [73]. With the unique feature that L_q is smaller than L_d , the FIIPM machine will not lose saliency under loaded condition. As a matter of fact, the saliency ratio increases under torque current saturation, hence in theory, the rotor position signal is easier to be captured with saliency tracking self-sensing scheme.

Investigations of the FIIPM self-sensing performance have been undertaken in [73-75]: the analysis is based on the comparison of the saliency profile between proposed novel design of FIIPMs and conventional IPM machines as shown in Fig 2.16. In these papers, finite elements simulations along with experimental tests on prototypes are used to evaluate the self-sensing performance in terms of the position signal strength and estimation accuracy. It is concluded that FIIPM is superior to conventional IPMs for self-



Fig 2.16 Optimizations of FIIPM based on baseline IPM machine [73]

sensing control: the feasible range is increased significantly and the estimation errors are reduced due to less effects of saturation on saliency properties and secondary saliencies.

The study in [76] was carried out with a concurrent design in consideration of both selfsensing and power conversion performance. It is found that besides this good self-sensing capability, by reducing the q-axis inductance with design of the q-axis flux barriers, the iron loss is also reduced thanks to less armature flux and saturation effect by load current. In [77], several injection based saliency tracking techniques were implemented to FIIPM, the square-wave injection on d-axis was selected as it gives the best overall performances.

In a different approach, an intentionally designed saliency is introduced by means of additional hardware features. The use of a short-circuited winding on the rotor of a PM machine is first introduced in [78] to improve the detection of initial rotor position. The technique has been further applied on surface mounted PM machines in [79-81] due to the limited intrinsic saliency of the rotor for position detection as shown in Fig 2.17. The short-circuited winding is configured as a ring around the magnet of each pole, naming at ringed-pole machine. Improvement of self-sensing capability is achieved thanks to the mutual inductance between the rotor winding and stator winding. The principle is based on that as the ring is wound on d-axis, a current will be induced when the d-axis high frequency field linking the coil variates, fluxes are generated from the coil resisting the flux generated from the stator winding, this is equivalent to the reduce of stator inductance on d-axis. On the contrary, the q-axis inductance is not changed under high frequency excitation as the rotor ring is not linked with stator q-axis.



Fig 2.17 Structure of the ringed-pole rotor with the short-circuited copper ring around the magnet [79]

The additional rotor coil winding with inductance L_r is coupled with the stator winding on d-axis and a mutual inductance L_M occurs between them. Current is introduced in this ring with high frequency variation of the stator excitation field, however the voltage in the short-circuited ring is zero. Shown in Fig 2.18, the magnetic coupling of the machine between the rotor and stator winding in rotating d-q reference frame, the main flux, the rotor as well as the ring rotates synchronously. Under high frequency analyzation, the daxis flux-linkage is summed up with the rotor coil flux-linkage and the flux produced by magnet, the q-axis is unaffected by the ring coil without the magnetic coupling. The high frequency saliency ratio $\xi(\omega_h)$ of the high frequency impedance $Z_q(\omega_h)$ over $Z_d(\omega_h)$ is derived in [80]

$$\xi(\omega_h) = \frac{Z_q(\omega_h)}{Z_d(\omega_h)} = \frac{R_r + j\omega_h L_r}{R_r + j\omega_h L_{rt}}$$

$$L_{rt} = L_r - \frac{L_M^2}{L_d}$$
(2.14)



Fig 2.18 Magnetic coupling of the rotor ring and stator winding [80]

In equation (2.14) the transient inductance L_{rt} is a function of the rotor ring inductance L_r , the mutual inductance L_M and the stator winding inductance on d-axis L_d . For high frequency injection, ξ becomes more inductive and a saliency ratio of $\frac{L_r}{L_{rt}}$ can be achieved and the contribution of the ring resistance R_r can be neglected as shown in Fig 2.19. Thus, the design consideration for the rotor ring is related to achieve a good coupling, this is studied in [82-84] for different configuration of the ring coil and its location on the rotor.



Fig 2.19 Frequency response of the high frequency saliency [80]

CHAPTER 2

The aim of the authors is to create a rotor saliency in such a way that the self-sensing control can be applied similarly to an IPM machine. The main benefits of this particular technique is that the anisotropy is created by electrical coupling between stator and rotor windings rather than structural iron saliency as for IPM machines, hence the magnetic saturation effect is less dominant. In addition, as long as the machine fundamental performance is concerned, the impact introduced by the copper ring is rather marginal since it rotates synchronously with the machine main flux. A FEM simulation strategy is proposed in [85] to analysis the self-sensing capability basing on high frequency injection: the evaluation is applied to an IPM machine, an Inset PM machine and a ringed-pole surface mounted PM machine. In [86, 87], the ringed-pole technique is utilized in an outer rotor machine to guarantee the self-sensing capability for an application as stator/alternator. The ring coil has been also put into an IPM machine as shown in Fig 2.20: Although the self-sensing control suffers from the effect of decreasing saliency ratio by iron saturation, the feasible region is significantly widespread by introducing the ring coil [88]. The losses of the rotor ring are analyzed in [89], sources of the copper loss includes the MMF harmonics, slot opening effects and also the high frequency signal induced current in the rotor ring. The author states that the loss is not important for small machines with distributed winding; but for machines with large-size or fractional-slot winding, such losses can be very high making the adoption of the ring coil not appropriate.



Fig 2.20 Shor-circuited winding in an IPM machine [88].

2.5-Summary

The existing literatures on self-sensing control of PMSM drives are reviewed in this Chapter. The overview covers the conventional vector control theory for PMSM machines which relies on the continuous knowledge of rotor position information. Self-sensing algorithms were proposed to enhance the overall performance of the PMSM drive for lower overall cost and better system reliability especially in applications with severe environment conditions. The main subject addressed in this PHD project is the saliency tracking based self-sensing technique for low-speed and initial position estimation when the model-based methods lose efficacy due to lacking of back EMF component.

Several saliency based self-sensing control methods are evaluated, based on high frequency signal injection as well as on measurement of the PWM transient current response. Two forms of the high frequency signal can be injected into the machine for position estimation, the rotating injection in stationary reference frame and the pulsating injection in synchronous frame. Outcomes of these two methods are similar in terms of estimation accuracy, yet the high frequency rotating voltage injection is adopted in this project for more intuitive saliency observation. Schemes based on measuring of the PWM transient current response termed as the IMFORM and the fundamental PWM excitation method are not adopted in this project due to their high requirements on hardware configuration, due to the need of changing the PWM excitation/sampling pattern and current derivative measurement.

The inductance saliency distortion due to machine saturation effects are of primary concern in this thesis. The worst case occurs at the current operation point when the machine loses saliency and the self-sensing scheme fails; in addition, the cross-saturation effects are resulting in the deviation of the estimated position angle. Many papers were focused on the optimization process during the machine design process to improve the self-sensing capability for wider feasible operational range. However, compromises between the machine general and self-sensing performances have to be made. Beyond that, researches have been applied in an optional way by introducing intentionally designed saliency features for self-sensing position tracking. Two candidates naming at the FIIPM and the Ringed pole machines are proved to be suitable for self-sensing control. Nevertheless, compromises were made invisibly under these kinds of design. For FIIPM, the space for magnet is restricted strictly by flux barriers and the saliency ratio is hard to increase as the path of d-axis flux passes through the magnet, which makes the d-axis inductance very small; while for the case of ringed pole machine, the drawbacks are related to the losses in the rotor ring which is unavoidable.

Up to now, for conventional machine structure, there is no specific technique to completely eliminate the machine saturation effects on self-sensing control, the problems are expressed as lose of saliency under heavy load operation and positon angle non-linear deviation caused by cross saturation with increased torque current. Also for the intentionally self-sensing oriented machines, it is unavoidable to degrade the machine general performances to a certain extent. Hence in this project, the problems of PM machine related to self-sensing operation will be analyzed, in addition, a solution will be provided for self-sensing PM machine with enhanced self-sensing capability.

<u>Chapter 3 HF Voltage Injection</u> <u>Saliency Tracking Self-sensing</u> <u>Scheme</u>

The adopted algorithm for self-sensing control in this research project is based on saliency tracking of the PMSM machine. The saliency is introduced by either rotor geometry anisotropy as for IPM machines or by the saturation induced saliency caused by magnet flux linkage for the case of SMPM machines. Despite of the specific machine structure, the d-q flux path reluctance saliency is actually reflected as the d-q inductance difference of the stator winding. The high frequency rotating voltage injection scheme is selected as the basic method for saliency tracking, thanks to its reliability, high accuracy, and also the straightforward angle calculation.

The rotor position is estimated by processing the saliency modulation current responds produced by the injection of high frequency rotating voltage as the carrier signal. Since in the demodulation process, the angle of the injection carrier signal vector is used, it is necessary to ensure that such is accurate. The errors that affect the voltage vector certainty caused by the pattern of DSP sample and calculation process and the inverter non-linearity are compensated in the demodulation process.

3.1- Inductance saliency modulation of PMSM

For inductance saliency tracking self-sensing control scheme, an additional carrier signal superimposed on the fundamental voltage is needed to provide the inductance saliency modulation: in fact the saliency cannot be extracted from the machine fundamental model itself synchronous with the rotating reference frame, and an asynchronous carrier signal is needed for the "scanning" of the inductance saliency. Voltage injection is superior to current injection as the injection frequency is not limited by the current control bandwidth [14], [15]. In addition, the higher the frequency of the carrier voltage, the better the performance of the self-sensing drive is able to achieve. With higher injection frequency, the machine circuit model is inductive more and hence

the effect of the stator winding resistance can be neglected; also if the induced high frequency current is out of the current control loop bandwidth, the vector control algorithm is not responding to the high frequency noise component caused by injection and the overall drive system is then more stable.

To derive the rotor absolute angle position, the machine dynamic model equation in rotating d-q reference frame is transformed in to the stationary α - β frame as in Fig.3.1. The machine voltage equation is given by:

$$\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} = \begin{bmatrix} pi_{d}L_{d}\cos\theta_{r} - pi_{q}L_{q}\sin\theta_{r} \\ pi_{q}L_{q}\cos\theta_{r} + pi_{d}L_{d}\sin\theta_{r} \end{bmatrix}$$
(3.1)

where V_{α} and V_{β} are the stator voltage in stationary reference frame and its relation to the rotating d-q frame currents i_d/i_q and inductances L_d/L_q are given. p is the differential operator and θ_r is the rotor electrical position angle defined by the angle between the d-axis and the α axis as shown in Fig. 3.1. It can be noticed that in (3.1), only the voltage component related to stator inductance is considered, it is assumed that the injection frequency is high enough that the machine stator is almost purely inductive such that the



Fig. 3.1. Park transformation from d-q frame to the stationary α - β frame.

resistance voltage drop can be neglected. In addition, the back emf component is not included as for low speed operation, its amplitude is relatively small and its frequency content is different from the saliency modulation component which can be cancelled by filtering in the demodulation process. Hence at this point, the relation between high frequency voltage and current is assumed to be defined only by inductance to simplify the analysis assuming that the position estimation is based on inductance saliency modulation.

Since the stator currents in stationary reference frame can be expressed from the d-q frame currents i_d/i_q as:

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = \begin{bmatrix} i_{d} \cos\theta_{r} - i_{q} \sin\theta_{r} \\ i_{d} \sin\theta_{r} + i_{q} \cos\theta_{r} \end{bmatrix}$$
(3.2)

The voltage equation in stationary reference frame can be represented with the stationary reference frame current i_{α} and i_{β} as:

$$\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} = \begin{bmatrix} \Sigma L_s - \Delta L_s \cos(2\theta_r) & -\Delta L_s \sin(2\theta_r) \\ -\Delta L_s \sin(2\theta_r) & \Sigma L_s + \Delta L_s \cos(2\theta_r) \end{bmatrix} \cdot \begin{bmatrix} pi_{\alpha} \\ pi_{\beta} \end{bmatrix}$$
(3.3)

$$\Sigma L_s = \frac{L_q + L_d}{2}, \, \Delta L_s = \frac{L_q - L_d}{2}$$

where ΣL_s is the average inductance and ΔL_s is the mean differential inductance. The detailed mathematical derivation process is provided in the appendix. It can be noticed from the equation (3.3) that the rotor position angle is related only with the mean differential inductance ΔL_s which is directly defined by the machine saliency ratio.

Optionally, the voltage equation in stationary reference frame can be represented by defining the space inductance as:

$$\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} = \begin{bmatrix} L_{\alpha} & -L_{\alpha\beta} \\ -L_{\alpha\beta} & L_{\beta} \end{bmatrix} * \begin{bmatrix} pi_{\alpha} \\ pi_{\beta} \end{bmatrix}$$
(3.4)

 $L_{\alpha} = \Sigma L_{s} - \Delta L_{s} \cos(2\theta_{r})$ $L_{\beta} = \Sigma L_{s} + \Delta L_{s} \cos(2\theta_{r})$ $L_{\alpha\beta} = \Delta L_{s} \sin(2\theta_{r})$

 L_{α} , L_{β} and $L_{\alpha\beta}$ stand for the space inductances as a function of stator inductances ΣL_s , ΔL_s and the electrical rotor position angle θ_r . It is the result of asynchronous voltage modulation that the equation is only meaningful when an asynchronous carrier voltage is injected and the inductance saliency modulation can be extracted.

3.2- High frequency voltage injection and current demodulation process

As in Fig.3.2, the high frequency voltage vector is injected superposed on the reference fundamental voltage vector $V_{\alpha\beta}^*$ calculated from the current control loop. The rotor position angle θ_r used for the transformation between the d-q frame and the $\alpha - \beta$ frame is derived from the current demodulation process in which the high frequency response of the stator current due to injection is demodulated.



Fig. 3.2. Diagram of the rotating voltage injection self-sensing scheme.

The injection rotating voltage can be represented as below:

$$\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} = V_h \begin{bmatrix} -\sin(\omega_h t) \\ \cos(\omega_h t) \end{bmatrix}$$
(3.5)

where V_h and ω_h are the injection voltage amplitude and angular frequency respectively. Substituting the voltage in (3.3) by the defined high frequency voltage vector, the high frequency current response due to injection can be represented as shown:

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = \int \frac{1}{\Sigma L_s^2 - \Delta L_s^2} \begin{bmatrix} \Sigma L_s + \Delta L_s \cos(2\theta_r) & \Delta L_s \sin(2\theta_r) \\ \Delta L_s \sin(2\theta_r) & \Sigma L_s - \Delta L_s \cos(2\theta_r) \end{bmatrix} \begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} dt$$
(3.6)

By integration, the current in $\alpha - \beta$ frame is derived as:

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = I_p \begin{bmatrix} \cos(\omega_h t) \\ \sin(\omega_h t) \end{bmatrix} + I_n \begin{bmatrix} \cos(2\theta_r - \omega_h t) \\ \sin(2\theta_r - \omega_h t) \end{bmatrix}$$
(3.7)

$$I_p = \frac{\Sigma L_s V_h}{\omega_h (\Sigma L_s^2 - \Delta L_s^2)}, I_n = \frac{\Delta L_s V_h}{\omega_h (\Sigma L_s^2 - \Delta L_s^2)}$$

Where I_p and I_n are the current amplitudes of positive and negative sequence current components, they are directly related to the average inductance ΣL_s and the mean differential inductance ΔL_s . The estimated θ_r is in the negative sequence current component and the detailed calculation process is as in Fig. 3.3.

As in the Fig 3.3, the procedure of position angle estimation is termed as synchronous frame filtering (SRFF). The measured stator current in $\alpha - \beta$ frame is firstly transformed with the injection vector angle θ_h which is derived from the injection vector: $\theta_h = \omega_h t$



Fig. 3.3. Detailed demodulation process of high frequency rotating voltage injection position estimation.

$$C(-\theta_h) = \begin{bmatrix} \cos(-\omega_h t) & \sin(-\omega_h t) \\ -\sin(-\omega_h t) & \cos(-\omega_h t) \end{bmatrix}$$
(3.8)

the transformed current is as:

$$\begin{bmatrix} i'_{\alpha} \\ i'_{\beta} \end{bmatrix} = C(-\theta_h) \begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = I_p \begin{bmatrix} \cos(2\omega_h t) \\ \sin(2\omega_h t) \end{bmatrix} + I_n \begin{bmatrix} \cos(2\theta_r) \\ \sin(2\theta_r) \end{bmatrix}$$
(3.9)

The whole frequency components are shifted forward with the injection frequency. Positive sequence component is located at twice of the injection frequency, and the position signal contained in the negative sequence current component is transformed into the fundamental component. Hence the inductance saliency modulation current component can be extracted with low pass filtering (LPF):

$$\begin{bmatrix} i_{\alpha}^{\prime\prime} \\ i_{\beta}^{\prime\prime} \end{bmatrix} = I_n \begin{bmatrix} \cos(2\theta_r) \\ \sin(2\theta_r) \end{bmatrix}$$
(3.10)

The rotor angle can be derived as:

$$\theta_r = \frac{\tan^{-1}(\frac{i_{\beta}''}{i_{\alpha}''})}{2} \tag{3.11}$$

3.3- Implementation issues of the high frequency rotating voltage injection self-sensing scheme

3.3.1 The discretized form of the DSP based processing algorithm

Since in the DSP system the data is processed in digital form, the continuous physical quantities of the machine are sampled and converted into discretized quantities with analogue to digital (A/D) conversion and the calculated reference command phase voltage is injected into machine with PWM implementation, and the equivalent average voltages applied to the machine are as shown in Fig. 3.4. Therefore, in addition to theoretical



Fig. 3.4. Equivalent average voltages of the reference

calculation, some implementation issues of the self-sensing algorithm in the DSP based platform are necessary to be considered.

The effects of DSP discretizing are more apparent for the self-sensing algorithm application as the operation frequency is far higher than that of the conventional vector control application. As higher injection frequency is desired, there is some unavoidable degradation as shown in Fig. 3.4. The switching frequency of the IGBT drive system used in this project is 10k Hz, and hence the average equivalent voltage output to the machine stator is hold for each switching period T_s (0.1ms). Also, the maximum injection frequency is directly related to the switching frequency, so that if high switching speed devices like the silicon carbide semiconductor are used, the injection frequency can be increased and the injected signal quality can be significantly improved. When a multi-level converter is employed, a better approximation of the output voltage can be obtained.



Fig. 3.5. PWM sampling mode with triangular carrier modulation [90]

In Fig.3.5, the progress for generating the inverter Gate signals according to the sampled PWM process is illustrated: the switching instants are calculated by intersection between the triangle carrier signal c(t) and the modulation signal m(t) which is derived from the reference voltage command. When the modulation signal m(t) is larger than the carrier c(t), the upper side switch is turned on; the lower side switch is turned on when m(t) is smaller than c(t). It can be noticed that the value of the modulation signal m(t) is sampled at the start of each switching period and the sampled modulation signal $m_s(t)$ is used for calculation of the switch turn-on/off duration. Hence the inverter equivalent average voltage is according to the sampled reference value at the start of each switching period, this is the cause of the conversion delay.

There are two major delays introduced by the DSP discretizing process as shown in Fig. 3.6. As the calculated reference voltage command in the current sampling point is



Fig. 3.6. Delays introduced by discretizing process.

CHAPTER 3



Fig. 3.7. High frequency vector angle offset caused by DSP discretization

outputted to the machine stator at the next PWM period, the voltage signal *Vref* is first delayed by one sampling period to the red curve representing *Vrefd*. Then, due to the delay of PWM conversion, the voltage signal is further delayed averagely by half of the switching period.

The overall effect of the DSP discretization process can be characterized as a deviation of the high frequency vector angle as shown in Fig. 3.7. The equivalent average voltage is delayed by one and a half switching period and hence the overall offset angle due to discretization is calculated as:

$$\theta_{offsetd} = 1.5 * T_s * \omega_{inj} \tag{3.12}$$

3.3.2 Inverter non-ideal aspects

In addition to the deviation of high frequency voltage vector caused by DSP discretization process, the inverter non-ideal behaviors also lead to the distortion of the stator voltage seen by the motor. For High Frequency injection self-sensing application, the distortions of the inverter output voltages are caused by the combined non-ideal behaviors of the





Fig.3.8. Current flow path during device switching in one phase leg of Voltage Source Inverter. (a) Positive load current (b) Negative load current

switching devices and the voltage errors are the differences between reference commands and the equivalent average output voltage. In Fig. 3.8, output characteristics of one phase leg of the inverter are examined to illustrate the power devices non-ideal switch behaviors. The most important ones are device voltage drop, turn-on and turn-off delay, dead-time and parasitic capacitance effects.



Fig.3.9. Ideal output voltage V_l^* (solid) and distorted voltage shape V_l (dashed) in one PWM period for positive and negative phase current i_l flow, accounting for power devices voltage drop (b), turn-on and turn-off delay(c), dead-time (d) and parasitic capacitance effects (e).

For the case of positive load current flowing as shown in Fig. 3.8 (a), during the dead time, both when switching from lower side to upper side and vice versa, the lower side diode is conducting to permit current flow. Hence for this single phase, the stator winding terminal is conducted to the negative pole of the DC bus. For the case of negative load current flowing, the upper diode is conducting when the phase leg is switching, hence the stator winding terminal is conducted with the positive pole of the DC bus. If only the dead-time is considered when modelling the inverter non-ideality, the output voltage error can be simply modeled as voltage loss and gain for positive and negative load current respectively. While during dead-time, the inverter output voltage is not directly changed in accordance to the ideal pulse shape. The real inverter output voltage characteristics are more complicated as shown in Fig.3.9 and the main nonlinear behaviors are as following:

A. Power Device Voltage Drop

Voltage drops of the Insulated Gate Bipolar Translator (IGBT) switch Vi and diodes Vd under conduction shift the inverter output voltage level according to the sign of load current i_l as shown in Fig.3.9 (b). The output voltage level is shifted lower if the load current is positive and higher when the load current is negative.

B. Turn-on and Turn-off delay

The delay consists of non-ideal switching characteristic of finite switching time of active components and the propagation delay of the gate driver circuit as shown in Fig.3.9 (c). Normally the value of switch turn-off time tdoff is larger compared to switch turn-on time tdon.

C. Dead-time

As the IGBTs require finite time to switch, the dead-time td is introduced to avoid current shoot through of the inverter DC bus, for the case when both the upper and lower IGBTs are in conduction. As shown in Fig.3 (c), during the dead-time, when load current i_l is positive, D2 in Fig 3.8 (a) is conducting resulting in average voltage loss in the PWM period; an average voltage gain is introduced during dead-time if the load current is negative since D1 in Fig 3.8 (b) is conducting. The distortion of voltage is related to both the DC bus voltage level and the pre-set dead-time which is set at design time according to device switching speed.

D. Parasitic Capacitance Effects

The rise/fall time of the inverter output voltage is affected by parasitic capacitance together with the load current level. As shown in Fig 3.8 (d), the rise time for case of positive load current and the fall time for case of negative load current are very short, this is because the parasitic capacitance is charging through DC bus during the instant when power switch turns. While for the case of device switching-off, the parasitic capacitance is charging with the load current, and hence the fall time for positive load current and the rise time for negative load current are relatively larger. Also the charging time is related to the current amplitude. The smaller the load current level, the larger the voltage error resulted by parasitic capacitance effects, while the voltage gain/loss pattern is opposite to dead-time errors and the error is inherently compensated when current is near zero.

In [8] and [17], the estimation methods of phase voltage error with respect to different current values were documented, the related self-commissioning procedures of DSP controlled voltage source inverter were also proposed. The voltage error is dominant by dead-time for large current values; for small phase current levels, the parasitic capacitance effect arises and the voltage error characteristics become more non-linear. Measurement of voltage error is based on current control of PMSM Machines. DC currents are injected on the α -axis, the rotor angle is locked that the magnet flux direction is aligned to α -axis and phase A of the stator winding. As shown in Fig.3.10, for the first step, two large DC current levels (6A, 8A) were injected for the first step, the equivalent resistance *Req* is derived according to

$$V_{\alpha ref} = \frac{4}{3} * V_{error_phase} + Req * i_{\alpha}$$
(3.13)

where $V_{\alpha ref}$ is the reference voltage on α -axis, and the injected current levels should be large enough that the test point is out of the non-linear region.

CHAPTER 3



Fig.3.10. Voltage error estimation procedure.

For the second step, current steps with interval of 0.1A for small currents and 0.2A for larger currents were injected and the phase voltage error for different phase currents are calculated as follow

$$V_{error_phase} = \frac{3}{4} * (V_{\alpha ref} - Req * i_{\alpha})$$
(3.14)

That the voltage error per phase V_{error_phase} is derived according to the difference between the command reference voltage and the excited stator voltage which is represented by the phase current and the equivalent resistance calculated from the first step.

Shown in Fig.3.11, the measurement result matches with the analysis, the voltage error is constant for larger phase current which is dominant by dead time; while due to the inherent error compensation of the parasitic capacitance effect, the voltage errors become



Fig.3.11. Estimated phase voltage error by (solid line) and the fitting line of the error compensation LUT (dashed line).

quite non-linear when phase current is small. A fitting line is calculated according to the measured results and the data is saved as a look-up-table (LUT) to compensate the voltage error caused by inverter non-linearity. In the control algorithm, the command reference voltages were superimposed by documented voltage errors with respect different current levels, so as to improve the reflection of the command voltage on the actual excited stator voltage.

Shown in Fig. 3.12, the overall self-sensing PM drive control scheme is provided with the error compensation techniques. The discretization error is compensated for the command high frequency voltage vector, so that the high frequency vector angle deviation is cancelled. In terms of inverter non-linearity, the compensation is achieved by using the estimated voltage error LUT, that before sending the voltage command reference value to the inverter, the voltage error is compensated with respect to the measured current value. Hence the position estimation errors arise from the control algorithm and power electronic limitations are minimized.



Fig. 3.12. Diagram of the self-sensing control scheme with error compensations.

3.4- Summary

In this chapter, the algorithm of high frequency voltage injection self-sensing control scheme is described. The PMSM saliency is utilized to track the rotor position angle with rotating voltage vector superposed on the fundamental voltage. The mechanism of inductance saliency modulation is first analyzed, and the demodulation process of the high frequency responding current is illustrated, in which the synchronous frame filtering (SRFF) process is used for the estimation of the rotor position angle.

Several implementation issues are discussed. Different from ideal calculation procedure, the estimated rotor position angle is affected by DSP discretizing process and inverter non-linearity behavior. The high frequency injected voltage signal is degraded by the PWM sampling pattern that the ideal sinusoidal voltage signal is converted into voltage pulses in each switching period. Two main delays of the signal are introduced: the sample delay arising as the voltage command calculated in the current sampling point is applied to the stator winding at the next PWM period; the conversion delay introduced as the voltage pulse for each PWM period is according to the voltage value at the start point of the PWM period so that the real reflected voltage signal is delayed by half of the switching period. The overall effect on self-sensing operation is that the high frequency

CHAPTER 3

vector is delayed by one and a half of the switching period and the voltage vector angle is deviated. The angle offset due to discretization process is calculated with the relation between injection frequency and the inverter switching frequency and is used to derive the accurate angle. The inverter non-linear behavior can also cause uncertainty of the real stator voltage injection, hence the voltage error is compensated according to the measured current value with respect to different phase current levels.

The works discussed in this chapter are mainly focused on the self-sensing control scheme adopted during the project, the HF rotating voltage injections method is used for PMSM machine rotor position estimation provided with direct estimation capability and operation simplicity. In addition, efforts were taken to restrict the estimation error resulted from the control algorithm and the inverter, so that the analysis can be brought into correspondence of the main concept that the PMSM drive self-sensing performances are related only to the machine itself.
<u>Chapter 4 Self-sensing Performances</u> <u>Analysis of Conventional PM</u>

Machines

The self-sensing performances of PMSM machines with conventional rotor configuration are investigated in this chapter with the assistance of FEM software. The use of the cosimulation between Matlab and the FEM software is introduced. The methods to examine the self-sensing performance of the machines by examine their incremental inductances are also described. The capabilities of self-sensing operation for an IPM and a SMPM machine are investigated by examining their inductance saliency. The mechanisms determining machine saliency are quite complicated to compare for different machines with different stator/rotor configuration, while by examining the incremental inductances of the machines, a unique procedure is used in this chapter to verify the machine inductance saliency which is used for self-sensing position estimation.

4.1- Joint-simulation of Matlab and FEM software

The FEM software Infolytica Magnet is used in this project for the computer based simulations of machine models. The scripting capability of the software enables the conjunction between Magnet and the data processing software Matlab. Hence the joint-simulation of Matlab and the FEM software can join their advantages so that the machine electromagnetic performances can be verified very quickly and the simulation time consumed can be significantly reduced. Throughout the project, the process as shown in Fig.4.1 is utilized to examine the machine self-sensing performance.

The interface of Matlab and Magnet is realized by Visual Basic commands and the scripts programmed in Matlab can be applied as the actions of the FEM software. The scripts in Matlab contain both the functions of data processing and simulation host control, also the data is received from Magnet and the commands are sent from Matlab. As illustrated in



Fig.4.1 Schematic of Joint-simulation between Matlab and FEM software Magnet.

the figure, for the specific case of machine self-sensing property analysis, the machine model under test is fixed for each simulation, since machine configurations remain unchanged different from the case of development of new machine designs. Simulation parameters like the machine speed, simulation time/step and stator excitation current/voltage values are primarily set in Matlab and sent to Magnet, then the simulation starts based on the selected analysis mode and the derived simulation results are sent back to Matlab for post–processing. The superiority of realizing the host control in Matlab scripts is that repetitive simulations can be achieved and the simulation process is automatic.

4.2- FEM estimation of Inductance Saliency

4.2.1 Incremental d/q inductance

A conventional and intuitive way to examine the saliency based self-sensing performance of a PMSM machine is by analyzing the machine incremental inductances. For this method, the incremental d/q inductance is calculated that the machine inductance saliency is examined with respect to load variations. By means of software, the machine inductance characteristics can be derived and the machine self-sensing property can be examined during the design stage. This is an important approach: when the drive system is aiming at self-sensing operation, acquisition of the machine self-sensing characteristics during the machine design process is of great help for the self-sensing control algorithm. In this way, the feasible operating region can be defined in which a reasonable machine saliency is detectable and also the saliency shift error caused by cross-saturation effect can be previously derived.

The simulation process to determine the machine incremental inductance is as illustrated in Fig.4.2. A multi-step simulation approach is utilized for each current operating point



Fig.4.2 Phasor diagram of incremental d/q inductance estimation process.

CHAPTER 4

with respect to different load. The simulation is applied in rotating d-q reference frame with the rotor rotating at low speed. Transient 2D simulation is adopted during the simulations, this can reduce the simulation time and is appropriate as the machine inductance saliency is mainly related to the machine 2D geometry and the radial flux distribution, the influences related to end winding can be neglected. Since the current control trajectory is selected that only q-axis current is utilized, the current operating point is moving along the q-axis according to load variations. The current test vector for the first step contains only the torque current component; the small test current increments are applied on the q-axis and d-axis respectively for the subsequent simulations. Incremental inductances are then calculated according to the variation of the flux-linkage caused by incremental d/q currents during post processing.

The current supply circuit as in Fig. 4.3 is configured to realize the incremental d/q inductance measurement process. For each phase winding, the d/q current components can be set separately with the supply current sources divided into d and q alignments: the d and q torque current sources and the d and q incremental current sources. The phase shift between d and q current sources is 90° and the 120° electrical displacements are set between the different phases. The initial phase rotor position angle is applied to guarantee the q current components to be aligned with the machine fundamental back-EMF



Fig.4.3 Stator winding supply circuit for incremental d/q inductance measurement.

component; this is according to a previous simulation of machine model under no-load operation. Then the q current components are aligned with the machine q-axis and the d currents are set to lag the q currents by 90° . The configuration of the circuit with separate current sources significantly simplifies the set-up of d/q currents.

The d/q flux linkages are increased due to the increment test d/q current as shown in Fig. 4.2. It should be noticed that for each d/q current incremental test step, the incremental currents on the other axis are set as zero, i.e. the d-axis current values are zero for each phase under the test step of incremental q current. This is to ensure the increment of stator flux linkage is related only to the d or q axis current respectively so that the inductances can be calculated accordingly. Selection of the increment test current amplitude is also important, the test currents are set relatively small so that the total flux linkage are close to the situation of the operation point regarding to each load level, and the measurement of the inductance is then obtained accurately.

The incremental inductance is calculated according to the variation of the flux-linkage and the related value of the increment current as:

$$L_{dinc} = \frac{\Delta \psi_d}{\Delta I_d}, L_{qinc} = \frac{\Delta \psi_q}{\Delta I_q}$$
(4.1)

where $\Delta \psi_d$ and $\Delta \psi_q$ are the variations of the d and q flux-linkage with respect to the increased d and q current amplitude ΔI_d and ΔI_q . Using the same simulation, the mutual inductances can be also calculated as:

$$L_{dq} = \frac{\Delta \psi_d}{\Delta I_q} , L_{qd} = \frac{\Delta \psi_q}{\Delta I_d}$$
(4.2)

I.E. the mutual inductances L_{dq} and L_{qd} are derived with the variation of the fluxlinkages $\Delta \psi_d$ and $\Delta \psi_q$ and the variant current on the opposite axis ΔI_q and ΔI_d respectively. The angle of the saleicny phase shift due to cross-saturation effect can be calculated with the mutual inductance and the incremental d/q inductances [91]:

$$\theta_{ds} = \frac{1}{2} * \tan^{-1}\left(\frac{2L_{dq}(L_{dinc} + L_{qinc})}{L_{qinc}^2 - L_{dinc}^2}\right)$$
(4.3)

The simulations of the incremental inductance estimation are fast as for each step, only one electrical period is analyzed to derive the average inductance value. Also the inductances for each operating point are derived with average values of the flux-linkages, hence the number of sampling points are relatively small.

4.2.2 Equivalent Incremental inductance trajectory

An optional approach to examine the PMSM machine self-sensing performance is by measuring the equivalent incremental inductance trajectory to derive the machine inductance saliency. For this particular method, the equivalent incremental inductance trajectories regarding load variations are calculated with high frequency rotating incremental current tests. Different from the incremental d/q inductances, the absolute inductance saliency derived from equivalent increment inductance trajectory is more accurate when the saliency is shifted due to cross-saturation effect. During machine operation, cross-saturation is unavoidable, hence in rotating reference frame, the maximum and minimum inductances are not always aligned with q and d axis of the machine. The incremental d/q inductances are therefore not appropriate to represent machine saliency under particular situations. For example, when L_{dinc} equals to L_{qinc} some saliency exists but it is shifted, yet the inductance difference is detectable. The concern for incremental inductance trajectory measurement is that the simulation time is relatively longer. Thus this approach is used as a supplement to d/q incremental inductance measurement that the simulation is applied to specific current operating points which are of primary importance like the rated torque operating point and the peak torque operating point.



Fig.4.4 Phasor diagram Equivalent Incremental inductance trajectory estimation As shown in Fig.4.4, to determine the equivalent incremental inductance trajectory of the machine, the simulation is applied into two steps for each current operating point with respect to load variations. Different from the previous simulation, the rotor is stationary and the current vector is formed by DC currents in each phase. Hence the d/q reference frame is in static mode and the rotor angle is adjusted that the d and q axis are coincided with the stationary α and β axis.

The machine operating situation in the FEM simulation is similar to that of the real situation when the machine operates in self-sensing mode. The torque current vector represented by the red arrow forms up the fundamental current component, for the first step with test vector 1, the flux-linkage in each phase is derived and transformed into α/β form; while with superposed high frequency injection signal for saliency modulation, the combined current vector is as the test vector 2 represented by the dashed black arrow, hence in the second step, the increased value of the flux-linkage due to additional rotating current injection is derived, and the equivalent inductance trajectory in α/β (d/q) can be calculated, according to the flux-linkage variation and the amplitude of the high

frequency rotating current. The rotor is locked at the particular position to ensure the measured equivalent incremental inductance trajectory in stationary reference frame to be same to that of the d/q frame, this is similar to the real case, as the injection frequency is far higher than the machine fundamental frequency and the rotor speed can be neglected.

The supply circuit of the 3 phase windings with separate current sources is based on the same mechanism of the previous simulation. The fundamental current is formed by d and q current components, while in this simulation, as the fundamental electrical frequency is set to zero for locked rotor, the fundament currents in each phase are DC currents with their values related to their phase angles. It should be noticed that the rotor alignment procedure is same to previous simulation I.E. the direction of the magnet flux-linkage is aligned with the α axis of the machine. The high frequency current vector is formed by the high frequency current sources in each phase. The current source frequency is set to be the same of the injection frequency (1kHZ) and the amplitude is kept small (0.1A) that the combined current is not deviate too much from the operating point. Different from the previous simulation where the incremental d/q inductances are derived from the average values, the number of sampling points is far lager to make the equivalent incremental inductance trajectory more close to the real situation as in Fig .4.5 (100 per step and 200 per simulation).

A specific case of the machine inductance saliency characteristic is as shown in Fig.4.5, due to the cross-saturation effect caused by the torque current, the ellipse trajectory of the machine equivalent inductance deviates from its original vertical position. The angle of the phase shift is equal to the result derived in (4.3). It can be noticed that due to the inductance trajectory deviation, the maximum and minimum inductances are different from the q and d axis inductances which are normally considered to calculate the machine saliency ratio defined as:

$$\xi = \frac{L_q}{L_d} = \frac{L_{qinc}}{L_{dinc}} \tag{4.4}$$



Fig.4.5 Equivalent Incremental inductance trajectory

In the self-sensing scheme, the saliency modulating signal strength is related to the absolute machine saliency ratio:

$$\xi_{ab} = \frac{L_{max}}{L_{min}} \tag{4.5}$$

By fitting the measured equivalent inductance locus with a matching ellipse, the values of L_{max} and L_{min} are derived which are half of the major axis and the minor axis of the ellipse; deviation angle θ_{ds} is also derived easily. In this chapter, both of the incremental inductance measuring schemes are used to examine the self-sensing performance of the machine models under analysis. Although the d/q incremental inductance results may not be appropriate to be used to define the position signal strength, it is however the fastest

way in FEM simulation to understand the rough self-sensing operation feasible region and to calculate the angle deviation caused by cross-saturation effect.

4.3- PMSM machines under FEM Analysis

An IPM machine and a SMPM machine are used as examples to analyze the self-sensing performance of general PMSM machines. As the rotor of IPM and SMPM machines are the most representative configurations for conventional PMSM machines, the study of their saliency characteristics is convincible to examine the feasibility and property of saliency based self-sensing position estimation method for PMSM machines with normal rotor configurations.

The machine model and the main parameters of the IPM machine under FEM analysis are shown in Figure 4.6 and Table 4.1. The 8 pole rotor is fitted into a stator with 18 slots. The winding of the machine is distributed with double layer configuration. Normally, rotor saliency is more apparent in machines with distributed windings where the d/q axis flux paths are more separate. The inductance saliency is resulted by the rotor geometry



Fig.4.6 IPM model for self-sensing performance analysis

Parameters	Value(units)		
Stator OD	120mm		
Rotor OD	76mm		
Air-Gap Length	1mm		
Slot Number	18		
Pole Pairs	4		
Magnet Thickness	3mm		
Magnet Span	85%/pole pitch		
Average Flux Density at Tooth	1.2T		

Table.4.1 Main Parameters of the IPM machine

characteristics that the reluctance of the q axis flux path is larger than d-axis, as the relative permeability of the magnet material is much smaller than that of iron. It can be observed that the rotor iron outer contour is optimized to quincunx shape, this is for the aim of better machine electro-magnetic performance that the air-gap flux density distribution is more sinusoidal and less harmonics are introduced. However, the machine inductance saliency is affected by this specific design that the effective air-gap of the q-axis flux path is larger. The contradiction between machine general and self-sensing performance is a common fact in design optimization process so compromises are unavoidable.

The structure of the example SMPM machine is as shown in Fig.4.7 with the main parameters listed in Table. 4.2. The rotor geometry saliency of SMPM machines is not obvious as the reluctances of d and q axis flux paths are nearly the same, as the permeability of the magnet material is similar to that of air. In addition, as the winding is concentrated configuration, the rotor anisotropy is less apparent as the d axis and q axis flux paths are overlapped. The stator inductance saliency for SMPM machines is mainly related to the iron saturation introduced by the permanent magnet. Due to the existence of magnet flux in d-axis, the flux path is more saturated compared to that of the q-axis, and hence the reluctance of d-axis is relatively larger. However, as the reluctance difference is



Fig.4.7 SMPM model for self-sensing performance analysis

Parameters	Value(units)		
Stator OD	96mm		
Rotor OD	52mm		
Air-Gap Length	1mm		
Slot Number	12		
Pole Pairs	10		
Magnet Thickness	3mm		
Magnet Span	95%/pole pitch		
Average Flux Density at Tooth	1.2T		

Table.4.2 Main Parameters of the SMPM machine

formed only when the machine is heavily saturated, the inductance saliency of SMPM machines is normally very small.

FEM simulations are applied to both machine models by varying the torque current *Iq* from zero(no-load) to 3 times of the rated current value(peak torque). The FEM results of machine flux density distributions for IPM machine model under no-load, rated load and

CHAPTER 4

peak load are shown in Fig. 4.8. Under no-load operation, the flux is only produced by the magnets. It can be seen that the field lines are symmetrical vs. the magnet axis so the flux linkage vector consists only d-axis component. The amplitude of overall flux distribution is fairly small so most of the machine parts are not saturated except for the iron bridges between the magnets which are intentionally made as thin as possible to avoid flux short-circuit related wastes of magnet usage. As the flux density of the stator teeth, back iron and rotor iron core is less than 1.2 T, the machine saliency introduced by rotor geometry is unaffected and is very apparent. However, with increased torque current, the amplitude



Fig.4.8 IPM Machine Model FEM results of flux density distribution under different load. (a) No-load (b) Rated-Load (c) Peak-Load

of the overall flux-distribution is increased and hence the machine is more saturated as in Fig.4.8 (b) and Fig.4.8 (c). Saturation is more remarkable in the stator than in the rotor as the width of the flux paths (teeth, back iron) is smaller. It can be noticed that the flux become symmetrical vs. inter-pole axis so the flux vector approaches the q-axis with increased stator currents. Hence the q-axis inductance is more affected.

For the SMPM machine model, the flux density distributions under no-load, rated load and peak-load are shown in Fig.4.9. Under no-load operation, the flux densities along the flux path including the stator teeth, back iron and rotor core iron are smaller than 1.2 T,



Fig.4.9 SMPM Machine Model FEM results of flux density distribution under different load. (a) No-load (b) Rated-Load (c) Peak-Load

similar to the case for IPM model. As the flux density distribution is controllable by defining the machine construction dimensions, it is a general design consideration to limit the no-load flux density to be smaller than the saturation density (normally 1.6T for common lamination material). Also, the flux contour is even symmetrical to the magnet axis, so the total flux is in d-axis. However, for loaded machine operation as in Fig 4.9 (b) and Fig 4.9 (c), the situation is slightly different compared to that of IPM machine. As the permanent magnets are mounted on the surface of the rotor and their permeability is similar to air, the effective machine air-gap is approximated to the thickness of the magnet and the overall reluctance of the flux path is much larger. As a result, the stator inductance is relatively small and the q-axis flux induced by stator winding is less dominant. Hence the main parts of the machine are not heavily saturated and the related decrease of the inductances on both axis are less significant.

4.4-FEM results of the conventional PM machine saliency characteristics.

The FEM simulation results for the saliency characteristics of the above IPM machine model and SMPM machine model are provided in this section. As shown in Fig.4.10 for



Fig. 4.10 Incremental d/q inductance of the IPM machine model

the IPM machine, both the d and q axis inductances and the mutual inductances are plotted with increased torque current Iq from zero to three times of the rated current. The mutual inductances Ldq and Lqd coincide so the rotor magnet flux direction is aligned well with the d-axis of the stator winding. Both the d and q axis inductances decreased with increased torque current as the machine tends to saturated. The decrease of the q axis inductance is more significant as the saturation of the q-axis flux path is directly resulted by the stator excited flux. It is also shown in the figure that Lq drops to be smaller than Ld when current is larger than two times of the rated current, hence the machine saliency is reversed beyond this operation point.

Shown in Fig. 4.11, the saliency ratio of the d and q inductances is calculated according to (4.4), while the absolute saliency ratio is also included with respect to (4.5). The two ratios are slightly different since the ellipse of the equivalent incremental inductance trajectory is deviated due to cross saturation effect. In the self-sensing estimation process, the absolute saliency ratio is directly related to the modulating position signal strength if



Fig 4.11 Saliency ratio of the IPM machine model.



Fig. 4.12. Saliency Phase shift of the IPM machine model.

the cross saturation phase shift is fully compensated. While the rotor position is undetectable when the saliency ratio drops to one at 2.5 times of the rated current, also the estimated d/q axis is reversed when the saliency ratio is smaller than one. This characteristic is the major disadvantage of saliency based self-sensing control since the self-sensing performance degrades at heavy load operation. And the position estimation becomes unfeasible for a specific current operation range.

The rotor position estimation error is directly related to the phase shift of the machine saliency, shown in Fig 4.12 the shift angle is derived in both angle that the black line is according to (4.3) for the incremental d/q inductance method and the red line is according to the deviation angle of the equivalent incremental inductance ellipse. The variation of the phase shift is quite large and non-linear: this makes it hard to apply the compensation mechanism since a detailed modelling of the estimation error is necessary.



Fig. 4.13. Incremental d/q inductance of the SMPM machine model



74

For the case of SMPM machine model, the variation of the inductance is much smaller because the effective air-gap is relatively large so the stator winding excited flux is less influential. However, the value of d and q axis inductances are quite close to each other and the situation of Ld Lq crossover still exists. Hence the saliency ratio of the machine is always small along the whole current operation range as in Fig. 4.14 which makes the detection process of inductance saliency hard to achieve.

In terms of the position estimation error, the overall phase shift of the SMPM machine model is smaller compared to the IPM machine. While the error also behaves non-linear along the torque current, this requires similar compensation effort.

For conventional PMSM machines, IPM machine is preferable for saliency based selfsensing control as the reflected inductance saliency is larger, thanks to its geometry anisotropy. However, as the effective air-gap is smaller, this type of machine is more affected by saturation effect since the variation of inductances is significant. For SMPM machines, although the saturation effect is less influential, the difference of the d/qinductances is however too small.



Fig .4.15. Saliency Phase shift of the SMPM machine model.

4.5-Summary

In this chapter, the analysis of conventional PMSM machines self-sensing performance is applied with the assistance of FEM simulation. The co-simulation between Matlab and the FEM software Magnet is described. Repetitive simulation procedures are automated by realizing the host control of the simulation with Matlab scripts, and the rapidity of the FEM software is fully utilized. Two methods to examine the machine inductance saliency characteristics are introduced, terming as the incremental d/q inductance measurement and the equivalent incremental inductance trajectory estimation. The principles of the methods are essentially similar: the incremental inductance is calculated according to the variation of flux-linkage caused by the increment of a small test current. While the incremental d/q inductance method can be applied straightforward which is suited for a rapid and rough estimation; the ellipse construction of the equivalent inductance trajectory method however requires much more simulation points for better accuracy. The latter process is then used as a supplement for the detailed analysis of specific operating points which are of major interest, like the rated current point and the peak current point.

Two typical PMSM machines are analyzed to examine their inductance saliency characteristics. The considered IPM machine with the magnets embedded in the rotor is provided with structure saliency as the permeability of the magnets is far lower than the iron; the saliency of the SMPM machine is due to the iron saturation caused by rotor magnets. In common machine design approaches, the no-load flux density of the main flux-paths is controlled to be somewhat smaller than the saturation point, hence the saturation induced saliency in conventional SMPM machines is normally very small, sometimes insufficient for rotor position detection. On the other hand, for the case of IPM machines, as the effective air-gap is smaller, the flux permeability and the stator winding inductance is relatively large, and the MMF due to the stator winding is more influential than SMPM machines, hence the inductance variation in IPM machine under loaded operation is much more significant.

In the simulation results, the saliency ratio and the phase shift derived with the two incremental inductance methods are used to represent the self-sensing capability for the two machines. The phase shift results coincides with each other and the saliency ratios are slightly different, while they provide similar conclusion that the IPM machine is more suited for saliency based self-sensing control at light load. However there is no ideal candidate in PMSM machines permitting to use the self-sensing control along the whole load range.

It can be concluded that for conventional PM machine structures, the saliency based selfsensing capability is limited: the main drawbacks highlighted in this chapter are the decrease of saliency ratio and non-linear cross-saturation phase shift. These characteristics are unavoidable unless the operation current is restricted to a small range that would not permit to exploit the full potential of machine. Hence for better selfsensing capability and proper fundamental machine performances, the machine conventional structure needs to be changed.

<u>Chapter 5 Design of the Saliency</u> <u>Modulation Rotor End</u>

In this thesis, a novel rotor configuration of permanent magnet synchronous machine (PMSM) is proposed to enhance the self-sensing capability based on machine saturation effect as discussed in previous chapter. A saliency modulation rotor end (SMRE) with pole number different from that of the magnet poles is added to the end of the machine rotor to provide an additional space anisotropic to the PMSM. Based on principle of rotating voltage injection, the new machine is provided with better self-sensing performance by tracking the position of SMRE. The position signal current caused by saliency modulation of the rotor end behaves as a harmonic component. Also, with the particular fractional pole/slot combination, saliency modulation of the rotor end is electrically asynchronous with the machine rotating reference frame, therefore, when high frequency injection scheme is adopted in low speed range, self-sensing performance is not affected by saturation effects under loaded operation. In addition, for medium and high speed ranges, when fundamental voltage vector is sufficient for excitation, rotor position can be tracked without superposed injection as the rotor end saliency modulation can be extracted by utilizing the fundamental voltage as carrier signal. In contrast with model based methods, this scheme is more reliable under machine parameter variations benefiting from the asynchronous saliency modulation of the rotor end.

5.1-Concept of Design considerations

The novel machine design has been applied by modifying the rotor of a PMSM machine. The saliency modulation rotor end is installed on the shaft at one end of the PMSM rotor so that the two parts are mechanically synchronized; stator core length is increased to fit the modified rotor. The total length of saliency modulation rotor end is kept to be same as a commercial encoder/resolver (18mm) for fair comparison. For position tracking, the mechanism is similar to that of variable reluctance (VR) resolvers, but the exciting coil is constituted by the same three phase machine winding and the end rotor is designed to provide reluctance variations for saliency modulation. The difference is that the position signal is derived within the main driving power electronic circuit and digital signal processing (DSP) units, additional electronic devices are not required for excitation and demodulation.

The design of the novel machine is analyzed by taking the following points into consideration:

 Pole number of the rotor end is selected different from that of the machine's main rotor to provide an electrical saliency modulation harmonic. the saliency modulation harmonic ratio h is defined for the prototype motor:

$$h = \frac{N_s}{N_m} \tag{5.1}$$

Where N_s is the pole number of the rotor end and N_m is the pole number of the PMSM rotor.

- The pole/slot combination and winding layout needs to be selected to ensure strong stator interaction for the pole numbers of both PMSM rotor and rotor end. This is to guarantee sufficient motor magnetic torque production and reasonable saliency modulation signal strength,
- 3. A particular set of fractional slot stator winding configuration is selected thanks to its strong MMF excitation for a couple of pole numbers: the phase sequences related to these two pole numbers are opposite to each other. Therefore, the fundamental flux vector for torque production is rotating asynchronously with the rotor end, and the saturation effect caused by the fundamental flux is avoided for high frequency injection at low speed operation; in addition, at high speed operation, the carrier signal can be formed by the fundamental driving vector that no superposed injection is required for self-sensing operation.

- 4. The stator structure and the winding configuration will be based on the selected baseline machine satisfies the above conditions, this will simplify the process of prototype machine manufacture.
- 5. The pole number of the rotor end and its geometry structure are the main design variables to provide air-gap reluctance variations.

Consequently, for the design of prototype machine to verify the proposed technique, the tasks can be summarized into two stages:

- 1. Selection of the baseline PMSM machine and the pole number of the rotor end with the consideration of pole/slot combination and winding configuration as mentioned above
- Since the design of stator and winding configuration is based on the baseline motor which is fixed, the rotor end geometry structure and its effects on reluctance of the stator excited flux path are analyzed to provide air-gap reluctance variations.

In this chapter, the selections of pole/slot combination and rotor geometry design are analyzed aiming at optimal saliency modulation position signal. The main decisions about the prototype machine including selection of baseline motor, rotor end pole number and also the feasible geometry of the rotor end are made in this chapter, whereas a further optimization process is applied in the next chapter.

Shown in Fig 5.1, the proposed saliency modulation rotor end configuration can be adopted for both SMPM machine and IPM machine. As the pole numbers of the main rotor and the rotor end are carefully selected according to above analysis, the position current signal resulted from the rotor end saliency modulation is isolated from the machine fundamental current and the current component related to the main rotor saliency. In this project, the main rotor is selected with surface mounted magnets as in Fig.5.1, due to the reason that the machine original saliency is very small and it is more challengeable to realize position estimation.



Fig 5.1 Machine rotor configuration with the proposed saliency modulation rotor end (a) SMPM rotor (b) IPM rotor

5.2- Fractional slot PMSM with non-overlapping concentrated winding

The fractional slot concentrated non-overlapping winding attracted research interests over few years, also they have been widely applied in different applications such as hybrid electric vehicles and servo systems. The advantages by utilizing fractional slot number includes: low-torque ripple, high efficiency, high-torque density, good fault-tolerance and flux-weakening performance [92]. The main drawbacks regard to the high stator excited MMF harmonics, these harmonics may potentially cause high rotor eddy current loss and audible noise. While on the other hand, the technique of the saliency modulation rotor end for self-sensing control is benefited from a strong low order MMF harmonic, the stator excitation for the rotor end with the pole number different from that of the main machine pole can be very strong and hence a strong position signal can be achieved. The superiority of fractional slot PMSM will be discussed. More than that, for the design of the prototype machine with the introduced saliency modulation rotor end, selection of pole/slot combination of the prototype machine is made in this stage, and that the bassline PMSM motor including stator winding configuration, the rotor type as well as the rotor end pole number is determined based on analysis of the concentrated winding with fractional slot number.

CHAPTER 5

5.2.1 Superiority of the PMSM with fractional slot non-overlapping winding

The definition of fractional slot winding refers to the average number of slots per pole per phase:

$$q = \frac{N}{2*m*pp} \tag{5.2}$$

where m is the number of phases, N is the number of total slots and pp, the number of pole pairs. Fractional slot winding features a number of slot per pole per phase q of the machine which is fractional rather than integer.

The machines featuring a concentrated non-overlapping winding with fractional number of slots per pole per phase have inherent benefits from short end windings, low cogging torque and hence higher efficiency and power density. Fractional slot windings were widely adopted in low speed water turbine generator stators. As for the requirement of grid frequency, the pole number of the generator is usually high so the pole pitch is small and the number q is limited as for smaller machine size and ease of manufacture, which is



Fig.5.2 Stator Winding Configurations: concentrated and distributed[92].

hard to achieve by distributed windings with integer slot number as in Fig.5.2: Moreover, for distributed windings, the back EMF is more distorted due to the slot/tooth harmonic when q is small [92]. Thus, non-overlapping fractional slot winding configuration is adopted thanks to its relatively simple manufacture process and ability to reduce cogging torque, as the phase differences of the tooth that EMF harmonics are interacted in series conductors in each phase.

In general, compared to machines with integer slot winding, the fractional slot nonoverlapping winding configuration is superior in the following aspects:

- 1. The total slot number is dramatically reduced, small slots are replaced with fewer big slots, and this reduces the total area required from slot insulation so that the slot copper fill factor can be increased. The smallest q for integral slot machine is 1, while the general feasible selection of q for fractional slot machine is within the range of 1/4 1/2, that the total slot number is at least halved.
- 2. A more sinusoidal back EMF waveform is achieved without employing the short pitching method as for integral slot windings in which the minimum number of q is 2 (normally 3 to achieve the same winding factor of fractional slot machine).
- 3. With concentrated windings, usually coil pitch defines the slot pitch, i.e each coil is around a single teeth. This is of great benefit to reduce the end winding length and simplify manufacture process. In addition, as the end windings of different phases are non-overlapping, no insulation is needed at the end area and the phase to phase fault tolerant is increased.
- 4. Easily reduced cogging torque and torque ripple. In fact, for machines with integral slot number, either the stator core or the rotor is often skewed to reduce cogging torque. For fractional slot machines, the harmonic order of cogging torque is several times the number of teeth and the amplitude is way smaller than that of the integral slot machines. Hence, no skewing is needed and the performance in terms of vibration and audible noise is greatly improved by using fractional slot windings.



Fig.5.3 Stator Winding Configurations: double layer and single layer [92].

The type of fractional slot winding can be either single layer or double layer depending the copper coils are wound either on alternate teeth or on all of the teeth respectively as shown in Fig.5.3. For the case of single layer winding, the selection of number of stator slots is limited as it must be even number to have multiples of 6, for 3 phase machines (multiples of 3 for double layer). In addition, the overhang of the end winding is larger so more copper material is required and the copper loss is increased; hence, double layer winding is preferred for PMSM machines.

Machine performances are improved by adopting fractional slot with non-overlapping concentrated stator winding configuration thanks to the increased slot copper fill factor and shorter end winding, the overall machine resistance is lower that the copper losses are reduced, and then the efficiency and torque density is increased. Above all, by fixed fractional slot winding configuration of the prototype machine, the PMSM machine general performances are benefited in terms of energy and material savings, smaller size and automation of manufacture process. The overall cost of the machine is reduced and its competitiveness is increased from commercial point of view.

The major drawbacks of fractional slot machines are related to the relatively higher space stator MMF spatial harmonics. While to extract the saliency modulation provided by the rotor end, it turns out that higher stator excitation harmonic is necessary, as the pole number of rotor end is different from that of the PMSM rotor. Also, the strong harmonic order should be lower so that the rotor end structural saliency can be well reflected in stator winding as inductance saliency. This is inherently provided by PMSM machines with fractional slot number and will be discussed by hereafter.

5.2.2 Selection of slot/pole combination

There are many slot/pole combinations for PMSM machines which are possible for applying the proposed technique. As shown in Table.5.1, the most popular slot/pole combinations are listed for 3 phase machine when different numbers of slot per phase per pole q are applied to each pole number. When q is selected as 1, it is the most general baseline slot/pole combination that for each pole pitch, a slot of each phase are applied. However, it usually offers the poorest performance in terms of back EMF harmonics and optimizations like short pitching can not be applied. For fractional slot PMSM machines with non-overlapping concentrated windings, the conventional slot/pole combination is when q is 0.5 so that the total slot number is 1.5 times the pole number. This type of combination is widely used in constant speed applications, the advantage is that no sub-harmonic appears for space MMF. On the other hand, the winding factor for fundamental excitation is relatively low (0.866) whereas it is desirable that the pole pitch is close to the coil pitch to increase such value.

In order to maximize the utilization of magnet flux for better torque density, winding factor is kept larger than 0.866 with the additional definition that q is selected in such a way that N_s is very close to pole number 2p, ie: $N_s = 2p \pm 1$ or $N_s = 2p \pm 2$. It should be noticed that these types of combinations cannot be applied to machines with number of pole pairs p which is multiples of 3 when a 3 phase machines is used, since the total slot number N_s needs to be divisible by 3 to allocate the coils evenly for balanced excitation. As the least common multiple between the total slot number N_s and the number of poles 2p is large, the cogging torque of machines with these slot/pole combinations is small,

Number of poles, 2p	Feasible slot number				
	q = 1	q = 0.5	$N_s{=}2p\pm 1$	$N_s = 2p \pm 2$	
2	6	3	3	_	
4	12	6	3	6	
6	18	9	_	_	
8	24	12	9	6	
10	30	15	9	12	
12	36	18	_	_	
14	42	21	15	12	
16	48	24	15	18	
18	54	27	_	_	
20	60	30	21	18	
22	66	33	21	24	
24	72	36	_	_	
26	78	39	27	24	
28	84	42	27	30	
30	90	45	_	_	
32	96	48	33	30	
34	102	51	33	36	
36	108	54	_	_	
38	114	57	39	36	
40	120	60	39	4	
42	126	63	_	_	
44	132	66	45	4	
46	138	69	45	48	

Table.5.1 Feasible slot/pole combinations for three-phase PM concentrated winding brushless machines[92]

usually with high harmonic order. However, for machines with odd number of slots, the main disadvantage is that the stator space MMF may be unbalanced as the stator slots are asymmetric disposed, which may result in excessive vibration and noise, while the bearing lifetime is also affected. For these reasons, the slot/pole combination of the prototype machine is selected with $N_s = 2p \pm 2$ as for the last column in Table.5.1.

In addition to general PMSM machine performances, for self-sensing control by tracking the saliency of the rotor end, the selected slot/pole combination is of great benefit for stator excitation under both pole numbers of the PMSM rotor and the rotor end. As in the last column in Table.5.1, the feasible slot/pole combinations appear as pairs, i.e. 15/14, 15/16 for odd N_s and 12/10, 12/14 for even N_s . The same stator winding configuration can then be used for machines with 2 different pole numbers, but the phase sequence is opposite for these two pole numbers. This feature is a perfect condition for the proposed saliency modulation rotor end design, when the pole numbers of the PMSM magnet rotor and the rotor end are selected as a pair of the same stator; both parts of machine are under strong excitation while they are electrically asynchronous. For high frequency injection scheme at low speed operation, the fundamental driving flux vector for torque production is rotating relatively to the rotor end, hence the saliency provided by the rotor end structural anisotropy is less affected by the main flux vector under saturation; when machine speed is high enough, as the fundamental driving flux vector is scanning the rotor end, saliency modulation can be formed by taking the fundamental voltage as the injected carrier signal. In addition, thanks to this particular winding configuration, the winding factor for both pole numbers are the same as will be analyzed next: both the torque production and saliency modulation sections can be then well suited for the stator winding using this particular configuration.

5.2.3- Equivalent stator excitation of 12/10 and 12/14

The pole numbers of the prototype machine are selected as 10 and 14 for rotor end and PMSM rotor respectively with a 12 slot concentrated stator winding: the respective stator excitation are as shown in Fig 5.4. As the total slot number is 12, the slot per pole per phase q = 2/5 for 10 pole rotor end and q = 2/7 for 14 pole PMSM rotor according to (5.2). For the prototype machine with the pole slot combinations of 12/10 and 12/14, the winding factor is calculated:

$$K_{w10} = K_d K_p = \frac{0.5}{2sin(\frac{30^\circ}{2})} * sin(75^\circ) = 0.933$$

And

$$K_{w14} = K_d K_p = \frac{0.5}{2sin(\frac{30^\circ}{2})} * sin(105^\circ) = 0.933$$

Hence the two sections of the machine have the same winding factor since the two slot/pole combinations have the same distribution factor and coil pitch factor. Due to the winding factor as well as the electric loading are quite high for both the rotor pole pairs of



Fig.5.4. 12/10 and 12/14 combination of fractional slot concentrated winding

5 and 7 as in Fig.5.4, the 12 slot concentrated stator winding of the prototype machine is a suitable choice to validate the proposed technique as mentioned before, to guarantee relative high stator excitation for both torque production and saliency modulation. In this thesis, the selection is made on 12/10 and 12/14 combination, although the proposed technique is also applicable to other slot/pole combinations. As a matter of fact, the fundamental inductance saliency related to rotor structure is more apparent in distributed stator windings. However, decision was made for a fractional slot winding as the wanted saliency modulation behaves as a harmonic component, strong stator excitation is very hard to achieve for integer slot windings that the resulted harmonic saliency modulation is barely detectable.

5.3-Comparision of the performances with different winding configuration

Aiming to achieve the best self-sensing performance so that the position signal strength caused by the rotor end saliency modulation is as strong as possible, different machine structures were considered for the prototype machine to validate the proposed technique. As discussed above, the fractional slot concentrated winding with strong excitation for a

pair of number of poles appears more suitable for the rotor end saliency modulation technique, due to its ability to provide strong stator excitation for both the PMSM rotor and the rotor end. The detailed analysis of the stator excitation flux reflected in the rotor is given in this section, self-sensing performance in terms of position signal strength is compared between the prototype machine (12 slots/14 pole PM rotor, 10 pole rotor end), a machine with integral slot distributed winding (18 slots/6 pole PM rotor, 30 pole rotor end) and a machine with fractional slot distributed winding (18 slots/8 pole PM rotor, 16 pole rotor end). The major winding types are included for comparison.

5.3.1 Stator excitation strength

Firstly, the strength of stator excitation for different number of pole pairs is compared as in Fig 5.4-5.6 by plotting the winding factor and the electric loading for the prototype machine, for the machine of 18/8 combination with fractional slot distributed winding and also for the machine of 18/6 combination with integer slot distributed winding. As discussed previously, with the 12 slot fractional concentrated winding that the stator is available for a pair of rotor pole numbers, the stator winding excitation is quite strong as in Fig.5.4 for both the 10 pole rotor end for saliency modulation and for the 14 pole PMSM rotor for torque production. It can be noticed from the figure that the winding



Fig.5.5. 18/8 Combination of fractional slot distributed winding



Fig.5.6. 18/6 Combination of integral slot distributed winding

factor as well as the electric loading are same for both the pole numbers, which is of great benefit for the rotor end saliency modulation and machine fundamental torque production.

As a matter of fact, the pole number of rotor end and the PMSM rotor can be exchanged since the stator winding excited flux are reflected on both the rotors as fundamental excitation with opposite rotating direction. Selection of the 10 pole rotor end is due to the consideration that the rotor structure saliency is more apparent in the stator winding with smaller pole number. In fact, for fractional slot concentrated windings, the rotor structure saliency is more difficult to be reflected as the stator winding inductance saliency compared to machines with distributed windings. However, thanks to the strong excitation factor available for two different pole numbers, the position signal strength is still guaranteed with a certain value of stator inductance saliency, on the contrary, stator excitation of distributed windings with fractional slot number are quite weak for rotors with pole numbers different from that of the fundamental rotor pole numbers as in Fig 5.5.

For the 18 slot 8 pole machine, the slot per pole per phase q is 0.75, so a reasonable pole number of the rotor end can be selected as 16 and 20 with relative strong stator excitation, yet the excitation is very weak. For the 18 slot 6 pole machine with integer slot distributed winding is in Fig.5.6, the slot per pole per phase q is 1, reasonable selection of

rotor end pole number is 30, although the stator excitation is very strong for this harmonic number, in a real machine, the position signal strength is limited due to the flux leakage as will be analyzed with FEA.

5.3.2 Stator flux reflected in the rotor

As shown in Fig.5.7, for the prototype machine with 12/14 combination of the PMSM rotor section and 12/10 combination of the rotor end section, the situations of fundamental stator excited flux reflected in the rotor are different for these two parts. For the PMSM rotor, the stator excited flux is in 90° phase difference compared to the



Fig.5.7 Fundamental stator flux reflected in the rotor:12/10-12/14 combination

magnet flux. In fact, since the rotor is with surface mounted magnets, the control strategy consists in applying the q-axis current component only since no reluctance torque is utilized; for high speed operation, when field weakening is required, the stator flux is shifted towards larger phase difference to that of the magnet flux. However, in steady state, the fundamental stator excited flux reflected on the rotor is constant in the PMSM section, since the flux rotation is synchronous with the PMSM rotor moving in same direction.

The saliency of the PMSM section is mainly the iron saturation saliency that the magnet flux saturates the iron parts along the flux path. However, for the prototype machine, such saturation saliency is avoided since the proposed technique is just based on track the saliency modulation provided by rotor end. Also it can be noticed in Fig 5.7 that for the PMSM saturation saliency, the situation is similar to that of rotor structure saliency since the reluctance of the flux paths of the d and q axis are different, hence the stator winding inductance in rotating reference frame Lq is larger than Ld.

For the PMSM section, the machine inherent saliency is not as ideal as expected since it varies due to the saturation effect of iron at large load as a result of the increased torque current Iq. This is the reason why this saliency is unreliable for position estimation: the worst situation occurs when the machine saliency disappears, also the saliency shift caused by cross-saturation effect also leads to position estimation error which is very hard to compensate. In addition, with fundamental voltage component only, modulation of the machine inherent saliency cannot be achieved as the stator excited flux is synchronous with the PMSM rotor: in this case, additional signal injection is necessary if saliency based self-sensing scheme is applied.

In contrast, the fundamental stator excited flux reflected in the 10 pole rotor end is rotating in opposite direction as shown in Fig 5.6, due to the fact that the phase sequence of 12/10 combination is opposite to that of the 12/14 combination, and the fundamental flux is synchronized with the 14 pole PMSM rotor for torque production. With this typical characteristics, saliency modulation of the rotor end can be achieved without
additional carrier signal as the fundamental stator excited flux is sweeping the rotor end, in which case the fundamental voltage itself can be seen as the a rotating carrier signal and the scheme is similar to that of HF rotating voltage injection. This scheme can be applied during machine medium and high speed operation when the fundamental voltage is high enough as a carrier signal. Compared to conventional adopted model based scheme in higher machine speed range, the technique is superior as the rotor position is calculated directly from the rotor end saliency modulation current signal, no mathematical modelling of the machine is needed in the estimation algorithms, hence the scheme is more reliable versus machine parameters variations.

For the case of machines with distributed winding, corresponding to the pole/slot combination of a 18/8 and a 18/6 machine as examples, the situations of fundamental stator excited flux reflected in the rotor are different from that of the prototype machine



Direction of the Whole rotor

Fig.5.8 Fundamental stator flux reflected in the rotor: Distributed winding

as shown in Fig 5.8. The fundamental stator flux is rotating synchronously with the PMSM rotor as well as the rotor end, hence both the machine inherent saliency and the saliency resulted by higher pole number rotor end are not modulated by fundamental signal. It should be also noticed that the saliency of high pole number rotor end is affected by saturation due to the synchronization of the fundamental flux vector which is fixed on certain pole teeth.

Similar to that of conventional machines, for the prototype motor with 12/10 rotor end section and 12/14 PMSM section, when machine is operating at low speed or during startup, the self-sensing position estimation still requires additional high frequency voltage injection as the fundamental voltage is not sufficient for rotor end saliency modulation. As shown in Fig. 5.9, the high frequency excited stator flux reflected in the rotor represented for different cases of conventional PMSM rotor saliency including both rotor structure saliency of IPM machines and magnet saturation saliency of SMPM machines, the saliency modulation of the prototype machine with 12/10 rotor end section and 12/14winding for PMSM section, and also the high pole number rotor end of machines with distributed winding used for comparison. For high frequency injection self-sensing scheme of conventional PMSM machines, as for the first case in Fig 5.9, the excited HF stator flux is modulating the machine saliency resulting either from rotor structure anisotropy or from saturation caused by magnets. The fundamental flux vector is synchronous with the rotating reference frame and is fixed on the rotating rotor. Hence, high frequency signal is injected to form an excited flux moving relative to the rotor to modulate the machine saliency, in this case, the shape of the HF flux path is similar to that of the fundamental flux while it is moving "faster" than the rotor in same direction. For low speed operation, the rotation speed of the rotor can be neglected so the relative speed between the HF flux and the rotor saliency is high enough that the HF machine model is nearly pure inductive. Due to the fact that the fundamental flux caused by torque current Iq is fixed on the rotor, saturation effects are very influential that the q-axis flux path is normally saturated during machine accelerating when PMSM inherent saliency is very hard to detect.



Fig.5.9 High frequency stator flux modulation of the rotor saliency.

For the second case where the machine is provided with distributed winding ie 18/8 and 18/6 combination, rotor end is selected to have larger pole numbers ie 16 and 30 for the 18/8 and 18/6 machine respectively. The situation of the stator excited flux is similar to that of the conventional PMSM, and the fundamental flux is synchronous with the rotor while the high frequency excited flux achieves relative motion for saliency modulation.

While in this case, the saliency number is increased such that the fundamental fluxes are not reflected in certain number of rotor end teeth. It is a possible solution that for low speed operation, those rotor end teeth without fundamental flux path can guarantee a certain degree of saliency under saturation, while the position signal strength is quite weak due to several practical reasons as will be analyzed next.

For the prototype machine as in the last case, the excited fundamental flux is moving in opposite direction of the rotor end, hence the saturation effect is not reflected in the rotor end. With the high frequency flux modulation, the rotor end saliency is always detectable and the position signal strength is guaranteed.

5.3.3 FEA comparison results

The FEA results for the selected slot/pole combination are compared in this section where the flux path and the flux density of the machine end part with the saliency modulation rotor end under Iq and Id stator current excitation is plotted by using FEA software. Analysis is focused on the stator excited flux path that the situations of different slot/pole combination are compared; hence the saliencies provided by rotor end of different combinations are compared. As the situation of the stator excited flux is related to air-gap length and rotor geometry structure, to eliminate these effects on saliency, the rotor geometry is set for all the combinations such that the air-gaps are same (0.5mm) and the rotor tooth width is half of the pole pitch.

The ideal situation of rotor structure saliency reflected in stator winding inductance is as shown in Fig 5.10, where the stator winding is distributed and the number of slot per pole per phase q is integer, this case is similar to a PMSM machine with salient rotor, ie by putting magnets in rotor slots as for inset magnet machines. Neglecting the fundamental excitation as for no-load operation, with the injected high frequency carrier signal, the stator excited flux is rotating relatively to the rotor end. When the carrier current vector is on q-axis of the rotor rotating reference frame, the situation is as in Fig 5.10 (a): the stator excited flux penetrates the rotor through the rotor teeth, this is the condition when the stator inductance is maximum (ie Lmax) as the reluctance of the magnetic path is



Fig.5.10 Flux path of 18/6 combination. (a) Iq excitation (b) Id excitation

minimum: For the case of Fig 5.10 (b) when the carrier current is on d-axis of the rotor rotating reference frame, the stator excited fluxes are supposed to cross the rotor slots, while in reality, they "leak" through the rotor teeth as in Fig 5.10 (b): the reluctance is a maximum in this case, so the stator winding inductance reaches a minimum as for *Lmin*.

Since the pole number of the rotor end is selected as 30 for the 18/6 machine, the situations of the flux are as shown in Fig 5.11: the field maps under *Iq* and *Id* stator current excitation are shown in Fig 5.11 (a) and Fig 5.11 (b) respectively. Different from the previous case, the "leakage" of stator excited flux is much more serious when carrier current vector is on d-axis of the reference frame defined with rotor end when a maximum overall reluctance is desired aiming for minimum inductance *Lmin*. As can be observed in Fig 5.11 (b), the flux path is expected to be across the rotor slot, however, as the rotor pole number is large, the pole tooth width is too small, so the stator flux leaks through the adjacent teeth. The corresponding result is that the overall flux linkage under d-axis current vector is on q-axis as in Fig 5.11 (a). Hence the rotor end structure saliency is not apparently reflected in stator winding flux linkage, so the saliency modulation signal strength would be very weak, as the pole number of rotor end is too large.



Fig.5.11 Flux path of 18/30 combination. (a) Iq excitation (b) Id excitation

For the cases of machines with fractional slot number, the magnetic paths of the stator excited flux are not evenly distributed in the rotor as shown in Fig 5.12 and Fig 5.13. In Fig 5.12, the 18 slot distributed winding is for the PMSM machine with 8 magnetic poles and the number of slot per pole per phase q is 0.75. The pole number of the rotor end is selected as 16, at which the stator excitation MMF harmonic is located. The number of the flux loops is still 8 as the stator winding is aiming for 8 poles for torque production. As can be observed in Fig 5.12 (a), with Iq excitation so that the stator inductance is maximum, the flux path is not distributed as ideal as for the previous case where the slot number is integer. Some rotor teeth are fully located under the stator teeth with flux flowing through, however, some are partially located under the stator teeth, so the flux flowing is affected. When the stator flux is excited with Id current as in Fig 5.12 (b), although some flux paths have high reluctance since the rotor slots blocking the flux, the other paths still have low reluctance with rotor teeth located exactly under the stator teeth. As a matter of fact, the rotor structure saliency is not apparently reflected as stator winding inductance saliency because of the non-even distribution of the flux loops, as for the situations of machines with fractional slot number. Moreover, the pole number of the



Fig.5.12 Flux path of 18/16 combination. (a) Iq excitation (b) Id excitation

rotor end is selected larger than that of the machine PMSM rotor, as a result, the position signal strength caused by saliency modulation of the rotor end is weak.

An optional approach for machines with fractional slot number is to have the rotor end pole number smaller than that of the PMSM rotor magnetic poles. For the case of the prototype machine, the stator winding is a 12 slot concentrated winding, which is suitable for excitation of both 10 pole rotor and 14 pole: for the aim of achieving a larger saliency ratio of rotor structure to be reflected as stator winding inductance variability, the rotor end is hence designed with 10 pole teeth as shown in Fig 5.13, with 14 magnetic poles for the PMSM rotor. As the number of slot per pole per phase q is 0.4 which is a fraction, the flux loops are also unevenly distributed. While it is different from the previous as the pole number of the rotor end is smaller, the overall reluctance of the flux path is relatively small under Iq stator excitation as in Fig 5.13 (a), most part of the rotor pole teeth are under the conducting stator teeth and the flux flowing is relatively smooth; for Id stator current excitation, although half of the flux are conducted as for Iq excitation, while the reluctance of the other flux paths is larger because of the rotor slots as shown in Fig 5.13 (b). In this case, the stator winding inductance saliency is relatively larger, so the position signal can be detected easily.



Fig.5.13 Flux path of 12/10 combination. (a) Iq excitation (b) Id excitation

To examine the inductance saliency related to rotor end in terms of position signal strength, the above machines with different pole slot combinations are simulated with FEA software. A high frequency current vector is injected into the stator winding defined as below:

$$\begin{bmatrix} I_{\alpha} \\ I_{\beta} \end{bmatrix} = \begin{bmatrix} I_{inj} \sin(\omega_h t) \\ I_{inj} \cos(\omega_h t) \end{bmatrix}$$
(5.3)

where the amplitude I_{inj} is 1A and the angular frequency ω_h is defined for 1kHz injection. The machine stator inductance defined in stationary α - β reference frame is as:

$$\begin{bmatrix} L_{\alpha} \\ L_{\beta} \end{bmatrix} = \begin{bmatrix} \Sigma L_{s} - \Delta L_{s} \cos(2h\theta_{r}) & -\Delta L_{s} \sin(2h\theta_{r}) \\ -\Delta L_{s} \sin(2h\theta_{r}) & \Sigma L_{s} + \Delta L_{s} \cos(2h\theta_{r}) \end{bmatrix}$$

$$\Sigma L_s = \frac{L_{max} + L_{min}}{2}, \Delta L_s = \frac{L_{max} - L_{min}}{2}$$
(5.4)

100

where θ_r is the electrical rotor position, *h* is the harmonic ratio of the saliency which is 1 in this case, ΣL_s and ΔL_s are the average and differential inductance defined by the minimum L_{min} and maximum inductance L_{max} in rotating reference frame. The respected flux-linkage in stationary reference frame is therefore as in (5.5)

$$\begin{bmatrix} \lambda_{\alpha} \\ \lambda_{\beta} \end{bmatrix} = \begin{bmatrix} i_{inj} \Sigma L_s(\sin(\omega_h t)) + i_{inj} \Delta L_s(\sin(\omega_h t \pm 2h\omega_r t)) \\ i_{inj} \Sigma L_s(\cos(\omega_h t)) + i_{inj} \Delta L_s(\cos(\omega_h t \pm 2h\omega_r t)) \end{bmatrix}$$
(5.5)



Fig.5.14 FFT of the stator flux linkage in stationary reference frame with 1k Hz HF current injection. (a)18/30 combination with integral stator slot distributed winding (b) 18/16 combination with fractional stator slot distributed winding (c) 12/10 combination with fractional stator slot concentrated winding

Such expression includes the component related to the carrier current with amplitude $i_{inj}\Sigma L_s$, and the saliency modulation component whose amplitude is directly related to the stator inductance saliency ΔL_s .

The FEA simulation results of flux-linkage are shown in Fig 5.14: the data is processed with discrete Fourier transformation (DFT) and the signal components are analyzed in frequency domain. To compare the position signal strength in terms of saliency modulation ability among the 3 different slot/pole combinations, the rotor mechanical speed is adjusted to provide same electrical modulation frequency ω_r which is 200 Hz; a relative value *SCR* is also defined by the ratio of saliency modulation component over the component related to the carrier current injection:

$$SCR = \frac{i_{inj}\Delta L_s}{i_{inj}\Sigma L_s} = \frac{L_{max} - L_{min}}{L_{max} + L_{min}}$$
(5.6)

which can be used to define the inductance saliency. The ratio SCR for the 18/30 combination with integral stator slot distributed winding, 18/16 combination with fractional stator slot distributed winding and 12/10 combination with fractional stator slot concentrated winding are 2%, 4.68% and 13.75% respectively. The results validate the previous conclusion that a better position signal from rotor end saliency modulation can be derived from the machine type with 12/10 concentrated winding.

5.4-Two Modes of Operation

As discussed before, For the stator MMF reflected in these two rotors, although the 10 pole rotor end and the 14 pole main rotor share the same winding, the respective phase sequences are opposite as shown in Fig.5.15, in which the two parts are observed from same direction The main stator MMF is rotating synchronously with the 14 pole rotor for torque production as in Fig.5.15 (a); while for the 10 pole rotor end, as the pole pitch is different, the resulting stator excited flux is asynchronously rotating in opposite direction as in Fig. 5.15 (b). Therefore, under the same stator winding, the relative phase sequences are opposite for the two parts and the saliency variation of the rotor end can be



Fig 5.15 Cross section of the prototype motor. (a) 12-slot 14-pole PMSM (b) 12-slot 10-pole rotor end.

modulated by the main stator flux. With this specific characteristic, during medium and high speed operation, fundamental voltage can be considered as carrier signal modulating the rotor end saliency so the rotating voltage injection self-sensing scheme can be realized without high frequency injection. Compared to model based position estimation technique in which the fundamental voltage vector is directly used to estimate machine back-EMF, the scheme is in principle saliency based, which is less affected by machine parameter change as no mathematical modelling of the machine is required.

During low speed operation when fundamental voltage is not sufficient for excitation, the self-sensing position estimation is achieved with the carrier signal formed by superposed high frequency voltage. It is the same technique applied to conventional machines with saliency, while the machine saturation has less impact on self-sensing performance. Since for the prototype machine, the main flux excited by stator winding is reflected asynchronously on the rotor end, the position signal is always detectable with increased driving current because the rotor end saliency is not weakened. In this case, the self-sensing capability is significantly improved compared to conventional PMSM whose saliency is greatly affected by machine saturation when large torque current is applied.

The technique is applicable to other pole/slot combinations, selecting the rotor end pole number different from that of the machine main rotor. And also choosing an appropriate pole/slot combination to ensure relatively strong stator flux excitation for both torque production and saliency modulation.

5.4.1 High frequency rotating voltage injection self-sensing mode

As shown in Fig 5.15, when a high frequency rotating voltage vector defined in (5.7) is applied to the machine, the induced stator current is derived as in (5.8). For high frequency injection operation, the fundamental current component related to voltage drop of stator resistance and the back EMF is neglected.

$$\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} = V_h \begin{bmatrix} -\sin(\omega_h t) \\ \cos(\omega_h t) \end{bmatrix}$$
(5.7)

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = I_{p} \begin{bmatrix} \cos(\omega_{h}t) \\ \sin(\omega_{h}t) \end{bmatrix} + I_{m} \begin{bmatrix} \cos(2\theta_{r} - \omega_{h}t) \\ \sin(2\theta_{r} - \omega_{h}t) \end{bmatrix} + I_{s} \begin{bmatrix} \cos(-2h\theta_{r} - \omega_{h}t) \\ \sin(-2h\theta_{r} - \omega_{h}t) \end{bmatrix}$$

$$I_{m} = \frac{\Delta L_{dq} V_{h}}{\omega_{h} (\Sigma L_{s}^{2} - \Delta L_{dq}^{2})} I_{s} = \frac{\Delta L_{rt} V_{h}}{\omega_{h} (\Sigma L_{s}^{2} - \Delta L_{rt}^{2})}$$
(5.8)

As in (5.8), the stator current in $\alpha - \beta$ frame contains a positive sequence component which is not related to rotor position and two negative sequence components which are induced due to rotor saliency modulation. The second current component I_m is the modulation result of saturation saliency caused by magnets and the number of saliency is twice of the PMSM rotor pole number; the third component I_s is the rotor end saliency modulation current, which is used to estimate the rotor position θ_r . As discussed before, the phase sequence reflected on the rotor end is opposite to that of the PMSM rotor, position modulation direction is reversed. The harmonic ratio h is defined as in (5.1), which are the ratio of pole number of the rotor end and the main rotor respectively. In this paper, rotor position estimation is based on the saliency modulation of rotor end and hence the second component is considered as disturbance and its amplitude needs to be small. For the prototype machine, this is achieved as the sizes of stator teeth and back iron are relatively large, so the flux paths are not saturated and the saturation induced saliency ΔL_{dq} is very small.

Demodulation process is applied to estimate the saliency position θ_{re} by means of Synchronous Frame Filtering (SRFF), as shown in Fig.5.16, measured stator current is shifted forward with the carrier vector frequency ω_c , position signal contained in negative sequence current component is extracted with a bandpass filter (BPF), the filtered currents i''_{α} and i''_{β} in (5.9) contain then only the modulation component of the rotor end saliency.



Fig 5.16. High frequency rotating voltage injection self-sensing scheme.

So the estimated PMSM rotor position θ_{re} is finally calculated from the saliency modulation signal:

$$\theta_r = -\tan^{-1}(i_{\beta}^{"}/i_{\alpha}^{"})/2h$$
(5.10)

The theoretical frequency spectrum of the high frequency voltage injection operation model is as in Fig 5.17. The amplitude and frequency (1k Hz) of injection signal is fixed to ensure sufficient position signal is detectable. The stator current in $\alpha - \beta$ frame is shifted forward by the injection frequency: according to (5.8), all the current components are then shifted forward by the carrier frequency (1k Hz). The rotor position is estimated basing on the rotor end saliency modulation current component I_s .



Fig 5.17 Frequency spectrum of high frequency injection operation mode

5.4.2 Fundamental rotating voltage modulation self-sensing mode

For medium and high speed-operation, the rotor end saliency modulation scheme can be realized without high frequency carrier voltage injection. The respected stator current is:

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = I_{f} \begin{bmatrix} \cos(\omega_{e}t) \\ \sin(\omega_{e}t) \end{bmatrix} + I_{s} \begin{bmatrix} \cos(-2h\theta_{r} - \omega_{e}t) \\ \sin(-2h\theta_{r} - \omega_{e}t) \end{bmatrix}$$
$$I_{s} = \frac{\Delta L_{rt}V_{f}}{\omega_{e}(\Sigma L_{s}^{2} - \Delta L_{rt}^{2})}$$
(5.11)

where the fundamental current I_f is the fundamental synchronous current vector related to the machine terminal voltage, back emf and stator impedance; The second component I_{sf} is the saliency modulation current and its amplitude is directly related to the fundamental voltage V_f , such that the position signal is stronger at higher machine speed.



Fig 5.18 Fundamental voltage modulation self-sensing scheme.



Fig 5.19 Frequency spectrum of fundamental voltage modulation operation mode

In medium and high speed range, when fundamental voltage is considered as the carrier signal, the second term I_m in (5.8) no longer exists due to synchronization yet the rotor end saliency modulation term I_{sf} still exists as the rotor end is electrically asynchronous with the fundamental vector. Shown in Fig 5.18, the demodulation scheme is similar: only the filter parameters and carrier vector frequency are changed to the fundamental frequency ω_f for SRFF as in (5.9) and rotor position is derived based on (5.10). In Fig 5.19, the theoretical frequency spectrum of the fundamental voltage excitation operation model is shown. The amplitude and frequency of the carrier signal is flexible related to machine speed and torque. According to (5.11), all the current components are shifted forward by the fundamental frequency. The rotor position is estimated based on the rotor end saliency modulation current component I_{sf} .

5.5-Sumarry

In this Chapter, a novel rotor configuration of a permanent magnet synchronous machine (PMSM) have been investigated, in which a saliency modulation rotor end (SMRE) is added to the conventional rotor to improve self-sensing capability of PMSM. The SMRE provides an additional space anisotropy feature to the rotor. Saliency modulation of the

rotor end is electrically asynchronous with the machine rotating reference frame. Therefore, when high frequency injection scheme is adopted in low speed range, the selfsensing capability is enhanced as the saliency ratio of rotor end is not affected by saturation. In addition, for medium and high speed ranges, rotor position can be tracked without superposed injection as the rotor end saliency modulation can be extracted by utilizing the fundamental voltage as carrier signal.

The main design concepts have been derived: the pole number of the rotor end needs to be different from that of the PMSM magnets and an appropriate slot/pole combination should be selected to guarantee both the torque production by the PMSM rotor and the saliency modulation signal strength provided by the rotor end. The fractional slot nonoverlapping concentrated stator winding configuration is analyzed and its capability to fit the proposed design is emphasized. Compared to the distributed fractional slot winding and distributed integer slot winding, the selected slot/pole combination is proved to be more suitable assisted by FEA simulation, as the concentrated winding provides the best saliency modulation signal strength: it is then selected for the prototype machine. The two operation modes in terms of high frequency rotating voltage injection for low speed operation and fundamental voltage excitation for medium and high speed operation are also explained.

With such a technique, the feasibility region of self-sensing control is extended. For high frequency injection scheme at low speed operation, it has been demonstrated that the position estimation performance is enhanced, thanks to its electrical asynchronous property. For medium and high speed range, rotor saliency modulation is realized with fundamental voltage as the carrier signal, which can be used to substitute the conventional model based methods. The technique is applicable to other pole/slot combinations of PMSM as long as the pole number of rotor end is different from that of PMSM rotor and is selected at one of the stator winding MMF harmonics. In this chapter, decision was made on using the non-overlapping concentrated winding both due to its strong excitation for the rotor end pole number, permitting to obtain a relatively strong overall saliency modulation signal, and because of its specific feature that the phase sequences for 14 pole

rotor and 10 pole rotor are opposite to each other. This is of great benefit for fundamental voltage excitation as the fundamental excited flux can be used to modulate the rotor end saliency.

The primary design process is mainly based on theoretical analysis; main design selections of the prototype are made. However, for better self-sensing capability, optimization of the rotor end structure is required, which is addressed in next chapter.

CHAPTER 6 OPTIMIZATION OF ROTOR END

Since the machine pole/slot combination has been selected at preliminary design stage, the optimization process of the rotor end geometry structure is developed for optimal self-sensing position signal quality in terms of larger current signal amplitude I_s and less harmonics. According to permitted modifications of the baseline motor, the dimensions of stator and PMSM rotor section are fixed; analysis is then focused on the rotor end geometry structure. The high frequency rotating voltage injection self-sensing mode is adopted for optimization with fixed carrier signal injected into the FEM machine model. Selection of the rotor end structure with its influence on the position signal is made by seeking optional compromises between the signal strength and the total harmonic distortions (THD).

6.1- Set-up of the GA optimization process

Considering the end section of the machine with the stator and the rotor end, principle of position estimation is similar to the mechanism of a variable reluctance (VR) resolver, the exciting winding is constituted by the three-phase stator windings and the end rotor is designed to provide air-gap reluctance variations. The advantage is that the excitation is realized by the main stator windings driven by the power converter, while the position signal demodulation process is applied within the main digital signal processing (DSP) unit: additional expensive electronic devices e.g. resolver to digital converters (RDCs), are not needed in comparison with a normal resolver. As the PMSM machine is selected and the winding configurations together with the stator core are fixed, the optimization is focused on rotor geometry structure for better position signal quality. Different from that of a conventional VR resolver with a smooth rotor contour, the considered structure of salient rotor end is rougher as shown in Fig 6.1. This is due to the tooth-wound concentrated types of stator winding: this way the side leakage between the flux paths with maximum and minimum reluctance is avoided.



Fig 6.1 Cross section of the saliency modulation rotor end per pole pitch.

The dimensions of the saliency modulation rotor end defining the rotor pole shape are provided in Fig.6.1, Among them, *Lgmin* is the minimum air gap length above the tooth top; w1 is the width of the tooth pole; h1 together with h2 defines the depth of the slot; K is the factor defining the air-gap length to ensure a sinusoidal trend of radial permeance above the tooth pole along the top of the tooth, according to the usual rotor contour design of VR resolvers [93]. For each tooth, By introducing the air-gap function factor K, the air-gap function becomes:

$$\delta(\theta) = \frac{KLgmin}{1 + (K-1)cos(\theta)}$$

$$1 \le K$$
(6.1)

The rotor tooth top contour variates by varying the value of K with a fixed minimum airgap length *Lgmin* as shown in Fig 6.2 (a). With larger value of K, the difference of the maximum and minimum inductance is lager, hence the position signal strength is stronger as in Fig 6.2 (b). K is limited to be equal to or greater than one, at which the tooth top is a circular arc. However, the factor K is defined for a VR resolver with distributed winding configuration: for the rotor end design under consideration, as the winding is



Fig 6.2. Definition of the air-gap function factor K [92]. (a) rotor contour (b) resolver signal strength

concentrated, only the part of the tooth top contour is defined by K as in Fig 6.1 and hence the influence of value K is relatively small.

As shown in Fig 6.3, the modeling and examination of the self-sensing performance of the rotor end is realized by a joint-simulation process of Matlab and the FEA software. Dimensions defined above are used as the input for the modeling process, defining the



Fig 6.3. Matlab and FEA joint-simulation process flow chart.

design variables of rotor end geometry structure; the output is the result of the current in terms of position signal strength and THD under high frequency voltage injection mode. Matlab scripting is used for analytical calculation and commanding of the FEA software Magnet. The modeling and analysis procedure is as follows:

- 1. The geometry dimensions for rotor end contour drawing are first calculated based on the design variables in Matlab.
- 2. The rotor end is constructed in FEA software according to previous calculations.
- 3. The Machine model is constructed for the new rotor end together with the fixed stator and external circuit which is set in advance including the high frequency rotating voltage injection. Simulation of the model is run in voltage driven mode including only high frequency voltage component, as the machine is operated at no load and fixed speed.
- 4. After the simulation ends, the three phase machine stator currents are extracted and transformed into stationary α - β reference frame, FFT analysis is used to derive the frequency spectrum of the current and the amplitudes of the current under each frequency are used to derive the position signal strength and THD.

Shown in Fig 6.4, the schematic of the joint simulation between Matlab and FEA software can be summarized as the rotating high frequency injection voltage vector is fed into the same stator winding, while in each cycle of optimization, a new rotor end is generated and used for simulation. The stator current is processed by SRFF: all current components are shifted and the saliency modulation current becomes the fundamental component. Only the noise components up to the injection frequency are considered in terms of THD calculation.

In Fig 6.5, the overall optimization process is shown; the 5 dimensions previously defined are used as the input design variables and objective functions are defined with two targets identifying the criteria to drive the genetic optimization towards maximizing the position signal strength and minimizing the THD. Design constrains are also set for mechanical rationality of the rotor structure and the minimum air-gap is limited to 0.4mm. A particular rotor structure identified by a specific set of design variables is analyzed in each cycle of optimization. The design is realized by joint simulation of Matlab and the FEM software as in Fig 6.3 so during each cycle, the motor model with the selected rotor end is created and examined by injecting high frequency voltage and calculating, the amplitude of the position signal and the THD can be derived. The multi-objective



Fig 6.4. Schematic of Matlab and FEA co-simulation

optimization (modeFrontier[©]) evaluates the simulation results based on objective functions and generates a new set of design variables according to genetic algorithms used to identify a new promising design in next cycle. Actually, the new design in each cycle is based on the analysis of the whole optimization history and all the data of analyzed designs are stored.



Fig 6.5. Design optimization process flowchart.

6.2- GA optimization Result

A large quantity of data were stored during the optimization process, more than 2000 designs of rotor end with different geometry structures were generated according to genetic algorithms. As shown in Fig 6.6, the criteria of the position signal quality in terms of position signal strength and the THD are set as the y-axis and x-axis of the bubble plot. The optimal limit (Pareto front) marked by the solid line shows the region of designs where the selection of rotor structure should be made.

As shown in Fig 6.6, the bubble color is selected as air-gap K to show its effect on position signal quality. It can be observed from the diagram that for best rotor end designs (design points near the Pareto front). Different from that of the VR resolvers, as the rotor pole is designed with slots and teeth, the influence of the K factor is smaller as the rotor structure saliency is more related to the rotor slot depth.



Fig 6.6. Bubble Diagram of the optimization result: Air-gap Factor K

As shown in Fig 6.7, the pole tooth height h1 is relative large (greater than 3.6mm) for the design points near Pareto front. This reflects that tall rotor teeth or deep slots are required to form up the rotor structure saliency, hence the reluctance variation of the flux path along the air gap is larger. Since the stator winding is selected as concentrated tooth wound, the fluxes correlated to d and q axis are not widely distributed; a steep rotor contour is needed to avoid the flux leakage between the two axes.

In Fig 6.8, the h2 dimension defining the pole slot depth is plotted and the value is larger than 2.5mm for best designs. The difference is that compared to h1, variations of h2 are more random. This is because this part of the slot is away from the air gap and its influence is relatively smaller. Hence the design of the slot bottom is less influential and its shape is modified for the final prototype machine considering mechanical manufacturing simplicity.



Fig 6.7. Bubble Diagram of the optimization result: Pole tooth height h1

The above three design variables roughly define the rotor pole shape for better position signal quality: each rotor pole is configured as a tooth, the tooth top is nearly a circular arc and a relatively deep slot is required. While in Fig 6.9, two decisive dimensions are selected in the synthesis diagram. Along the Pareto front, the position signal strength and the THD of the signal are highly depending on the minimum air gap length *Lgmin* and the width of the pole tooth wl represented by the color and size of the bubble respectively. The signal strength is higher with smaller minimum air gap and narrower pole tooth, but the signal THD is larger in this case; for rotor ends with larger air gap and wider tooth, position signal THD is smaller while its strength is also reduced. The optimization results in Fig 6.9 also illustrate that the two optimization objective functions conflict with each other in the sense that stronger signal strength leads to higher THD. This is because when the variation of the air-gap reluctance is large with small air gap and narrow rotor end tooth, the air gap permeance distribution is distorted from pure sinusoidal distribution as



Fig 6.8. Bubble Diagram of the optimization result: Pole slot depth h2



Fig 6.9. Bubble Diagram of the optimization result: Minimum Air-gap *Lgmin* and pole tooth width *w1*

the variation of the effective air-gap length is too steep. However, the reluctance difference cannot be made very large for a sinusoidal permeability distribution as variation of the effective air-gap is smooth.

6.3- Selection of Rotor end Structure

Rotor A and Rotor B as in Fig 6.10 are selected for comparison, their geometry dimensions are according to the design points as shown in Fig 6.9 for two representative cases. The structure of the rotor end for case A is with wider pole tooth and larger air-gap, hence the air-gap permeance distribution is smoother and thus there are fewer harmonics in the saliency modulation signal while the signal strength is relatively small. For case B, the rotor end is with narrower pole tooth and smaller air-gap, the saliency modulation signal strength is then larger but a lot of harmonics are introduced as the air-gap permeance distribution is too steep and the saliency inductance modulation is distorted from pure sinusoidal modulation.



Fig 6.10. Selected rotor structure designs from optimization results. (a) Rotor A (b) Rotor B $\,$



Fig 6.11.Frequency Spectrum of stator current with high frequency voltage injection at no-load using FEA simulation. (a) Rotor A (b) Rotor B

The frequency spectrum of the stator current shifted forward by the injection frequency (1k Hz) at no-load operation, rated current and peak current operation which is 3 times of the rated current are compared between rotor A and rotor B as in Fig 6.11, Fig 6.12 and



Fig 6.12.Frequency Spectrum of stator current with high frequency voltage injection at rated current using FEA simulation. (a) Rotor A (b) Rotor B

Fig 6.13 respectively. The position signal of rotor A is smaller than rotor B, however for rotor B, some harmonics exist even at no-load on accounting to the distortion of the airgap permeability distribution. Under rated and peak current operation, it is clear that large amount of noise harmonic components are introduced for rotor B. The harmonics which



Fig 6.13.Frequency Spectrum of stator current with high frequency voltage injection at peak current using FEA simulation. (a) Rotor A (b) Rotor B

are close to the position signal frequency are of great influence on position estimation, these noises are very difficult to be removed. For the case of rotor A, the modulation signal is much cleaner as the air gap permeability varies more smoothly: despite lower signal strength, this solution appears more suitable in terms of filtering effort.



Fig 6.14. Machine rotor with Saliency Modulation Rotor End (SMRE).

As a matter of fact, the distortion noises are a big problem in feedback control system and demodulation process. Harmonic components with frequencies close to the tracking signal are rarely eliminated completely. This will lead to system instability and increases the complexity of the self-sensing algorithm. Hence for smaller THD and reasonable signal strength, the rotor end structure of prototype machine shown in Fig 6.14 is designed based on Rotor A with the rotor slot bottom optimized to round shape.

6.4-3D FEA Simulation

As shown in Fig. 6.15, the prototype machine is modeled in FEA software with the rotor end at the end of the PMSM rotor for 3D analysis, the copper coil extended out of the stator iron core is connected with an external circuit to form up the stator winding, an air box is used to contain the machine model. It can be noticed that the end winding part of the machine is not included as its performance is not the scope of this project and the simulation can be achieved with the 2D winding. However, for the rotor of the prototype machine, as the PMSM part and the rotor end part cannot be configured with 2D simulation, the 3D simulation is used to analyze the overall performance of the rotor.



Fig 6.15. 3D FEA Model of the Prototype machine.



Fig 6.16. Schematic of the prototype machine.

One concern of the rotor flux distribution of the rotor with PM magnets is related to the axial flux distribution as shown in Fig 6.16. The paths of potential flux conduction from one magnet pole to another are through the rotor end. Since the machine is radical flux types, only the main flux penetrating into the stator through the main air-gap is related to torque production. The axial flux tubes as shown in Fig 6.16 are considered as flux leakage which leads to machine power loss, and it is a waste of the magnet usage. Hence an axial air-gap between the rotor end and the magnets is required to reduce the flux "short- circuit": its thickness should be greater than that of the main machine air-gap length which is 1mm. In the manufacturing process, a non-magnetic plastic ring with the thickness of 1.5 mm is put between the rotor end and the PMSM rotor. As shown in Fig 6.17, the axial flux distribution of the whole rotor when machine is operated at no-load condition, the fluxes short circuited from rotor end are reduced with the non-magnetic rotor ring which operates as an equivalent air-gap.



Fig 6.17. Axial flux distribution of the rotor at no-load operation point. (a) Without the plastic ring (b) With plastic ring

Another concern is related to the flux distribution in the rotor end as the fundamental torque current is increased for higher torque production, for examples during starting and acceleration. Also the stator excited flux reflected in the rotor end is asynchronous; this will give rises to iron losses in the rotor end. As shown in Fig 6.18, the flux density of the rotor and stator iron core is plotted when machine is operated at rated current, two times of the rated current and three times of the rated current respectively. The simulation points are selected when the rotor rotated to the time instant when maximum flux density is



Fig 6.18. Flux Density of the rotor and the stator of the rotor end section. (a) rated fundamental current operation (b) two times of the rated current operation (c) three times of the rated current operation.
achieved, i.e. when the rotor teeth are exactly aligned to the flux producing stator teeth. This is for the aim to analyze the maximum possible flux density of the rotor end machine section to check the influence of the stator flux caused by increased fundamental torque current. It can be shown from Fig 6.18 that the maximum flux density at peak current operation is 0.8T in the rotor and 0.85T in the stator iron, this is far lower than saturation flux density of the iron material M235-35A which is around 1.25 T. It can be stated that the rotor end section of the machine will not be saturated even at peak current operation. This derives from the optimization process produced a relatively large main airgap; also the use of the flux blocking plastic ring between the rotor end and the PMSM rotor avoids the leakage of the magnet flux into the rotor end section is only generated by stator winding excitation, so the resulting flux-density is far smaller without the magnet flux. Therefore, it can be proved that the position signal caused by saliency modulation of the rotor end is not weakened when the machine is operated with increased current.

In respect to the iron losses of the rotor end, FEM simulation results show that under rated speed (2400rpm) /current (8A rms) operation, the iron loss in the rotor end is only 0.135W. This also derives from the fact that the flux density in the machine end section is much lower than in the active region, as there is no PM flux and the air-gap between the rotor end and the stator is large, hence the increased rotor loss in the SMRE is very small.

6.5- Summary

In this chapter, the work on the optimization of the prototype machine is described. With the machine pole/slot combination selected by the preliminary design process, the task developed in this chapter is to optimize the machine for the aim of better self-sensing performance. The rotor end geometry structure is firstly defined with 5 dimension parameters; these parameters are then used as the free design variables in the GA optimization process. A joint simulation using FEA and Matlab is used for the optimization process: at each cycle, a rotor end with new structure is modeled and evaluated, and simulation results of stator currents are processed to derive the position signal strength and THD. The dedicated software ModeFrontier© is used to apply a GA

optimization: the evaluation of the rotor end structure design is driven by the goal function for better position signal quality in terms of high signal strength and low THD. A rotor end structure with relative large air-gap and smooth air-gap permeance distribution is finally selected, providing a cleaner position signal for less filtering effort in the control system; lower saturation, so that the saliency is not affected by over load; smaller flux density, so the iron loss of the rotor end can be minimized.

CHAPTER 7 EXPERIMENTAL VALIDATION

An experimental test-rig was built for several purposes to implement the self-sensing algorithms as discussed in Chapter 3; to investigate the self-sensing performances of conventional machines as in Chapter 4; to confirm the enhanced self-sensing capability of the prototype machine with the novel design of the saliency modulation self-sensing rotor end as discussed in Chapter 5; to validate and Chapter 6 is conclusions. Experimental results are presented conforming the high frequency rotating voltage injection self-sensing scheme for machine's zero and low speed operation and the particular self-sensing scheme termed as fundamental voltage modulation mode for high speed operation of the prototype machine.

7.1-Structure of the test-rig

Fig 7.1 shows the overall structure of the experimental test-rig. As shown in Fig 7.2, a commercial development platform: Myway Inverter Platform (Model: MWINV-9R144) is used as the controller of PM machines and to implement both the basic machine vector control and the self-sensing algorithms. The main power circuit is composed of an uncontrolled rectifier, the DC link and the inverter. The inverter is controlled by the Gate drivers in which the calculated switching signals are transformed into IGBT gate signals. The digital control platform features a F28335A DSP (Digital Signal Processor) as the core in the Myway expansion control board: PE-PRO is used for the integrated processing of the digital data. The inverter interface board is used for the pre-processing (analog to digital conversion) of the measured DC-link voltage and phase current signals, and also generates the IGBT switching signals in accordance to the calculated reference voltages from DSP in each switching period. The encoder installed on the PM machine is powered and decoded by an encoder interface board that converts the position signal of rotor angle into ABZ signals which can be directly processed by DSP. A real time CPU scope unit is used for the communication between PC and the DSP platform, permitting the algorithms coded in C++ to be loaded into the DSP chip. The data processed in the DSP can be



Fig.7.1 Overall structure of the experimental test-rig.

uploaded to PC to monitor the on-line parameters of the drive system. In the host control software, any global variables can be modified during the drive operation, a virtual oscilloscope is also included in the software to observe currents, voltages and rotor position provided with real-time parameter updates.

The PM machines under test are coupled with a DC machine trough a torque transducer as shown in Fig 7.3. The torque and the speed of the machine can be monitored on the torque meter screen. The DC machine is used to provide load for the test machines and can be also used to drive the test machines for back EMF measurements or to operate the



Fig.7.2 Myway Inverter Platform (Model: MWINV-9R144)



Fig.7.3 PM machine under test and the load DC Machine.

test machines in generating mode. A four-quadrant regenerative DC drive controller is used to control the load machine, permitting to operate it either in motoring or generating mode, using either speed control or torque control. Thus, the experimental tests can be carried out with various operation modes to simulate different machine working conditions; the whole system is also environmentally friendly when the machine is operating as motor, thanks to the regenerative function that makes the energy consumption during the tests very small.

7.2-Main Components of the Test-rig

7.2.1 Measurement and Protection of the Power circuit

In the Myway inverter platform, two current sensors for phase currents measurement and one voltage sensor and one current sensor for the DC link voltage and current measurement are installed as shown in Fig 7.4. The third phase current sensor is not needed since the third phase current can be calculated with other two phase currents for any balanced three phase load. This is a general approach for more economic drive systems. Both the voltage and current sensors are Hall-effect transducers with precise measurement and electrical insulation ability. The conventional input voltage range of analog to digital conversion (ADC) is \pm 5V and for this product: the ADC measurement resistances are set so that the current measurement range is 0A-65A for DC current and - 31.25A-31.25A for AC current, while the DC voltage measurement range is 0V-1000V.

For the protection of the drive system, in the inverter interface board, the hardware protect function can be set with variable resistance that the protection value can be set. When one of the protection conditions is detected, i.e any measured current/voltage exits the pre-permitted range, the protective function is activated and the gate signals are blocked (all switches turned-off). In addition to this, to prevent huge inrush current when charging the DC capacitors, an inrush current prevention circuit is installed. The contactor of the inrush current prevention circuit is kept off at system power on, so the current flows through the resistance and the capacitors are gradually charged; after a suited time,



Fig.7.4 Block diagram of the Myway inverter platform [94]

the contactor is closed to permit starting normal operation. A fuse protection on negative DC bar and a thermal monitoring mechanism of the heat sink tripping at 85° Care also included.

7.2.2 PM machines for testing

During the research work, the self-sensing performances of different types of machines were tested and compared. Three PM machines were analyzed: An IPM machine, a SMPM machine and the prototype machine with implementing the proposed solution of the saliency modulation self-sensing rotor. Their parameters are listed in Table 7.1: the IPM and the SMPM are the machines analyzed in Chapter 4, the rotor of the prototype machine in Fig .7.5 is manufactured according to the design provided in Chapter 5 and Chapter 6 including an optimized saliency modulation self-sensing rotor at the end of the SPM rotor with in accordance to the 3D model shown in Fig 6.14. These machines were installed on the test-rig and coupled to the same DC load machine and converter platform, hence their self-sensing performances can be fairly compared.

Parameters	Value(units)	Parameters	Value(units)
Rated Power	2.3kW	Rated Power	1.5kW
Rated Voltage	220V	Rated Voltage	220V
Pole Pairs	4	Pole Pairs	5
Rated Speed	2000r/min	Rated Speed	2000r/min
Rated Torque	10.8Nm	Rated Torque	6.9Nm
Connection	Y	Connection	Y
Stator Resistance(ph-ph)	0.8Ω	Stator Resistance(ph-ph)	0.63Ω
Inductance (Ld/Lq)	4.6mH/7.1mH	Inductance (Ld/Lq)	1.69mH/1.71mH
Ke	99.3Vrms/krpm	Ke	89.2Vrms/krpm

(a)

Parameters	Value(units)	
Rated Power	1kW	
Rated Voltage	220V	
Pole Pairs	7	
Rated Speed	2400r/min	
Rated Torque	3.9Nm	
Connection	Y	
Stator Resistance(ph-ph)	0.52Ω	
Inductance (Ld/Lq)	1.34mH/1.35mH	
Ke	72.5Vrms/krpm	

(b)

Table.7.1 Parameters of the PM machines under tests. (a) IPM (b) SMPM (c) prototype machine with the saliency modulation self-sensing rotor



Fig.7.5 Rotor of the prototype machine.

As the pole numbers, back EMF constants and the stator electrical constants for the three machines are different, an initial procedure for machine vector control set-up is required for each of them before examining their self-sensing performances. The task is mainly related to tuning the PI control loop and set-up of the encoder as discussed next.

7.2.3 Set-up of the Encoder

The output of the incremental encoder used for PM machine rotor position measurement is in digital pulse form. The signals of A and B are used to permit determining the position variation and rotation direction of the rotor by counting the pulse number and their phase relationship. The Z signal provides 1 pulse per mechanical rotation at the reference position to permit resetting the counter so that the absolute rotor position can be detected. For the research project, the measured rotor position is used for general machine vector control setup and can be also taken as the reference to compare with the rotor position estimated with self-sensing algorithms. Offset of the zero position indicated by



Fig.7.6 Z signal alignment with stator Phase A

the z pulse can be calibrated in the control algorithms: the position of the encoder can be adjusted such that the z pulse is aligned with the Phase A of the stator as shown in Fig 7.6. Thus, by rotating the rotor enough at start up to achieves the encoder reset, the encoder measured position can be used as the absolute rotor position during normal operation. It can be also noticed that as the machine neutral point is not accessible, the line voltages *Vab* and *Vac* are measured to reconstruct the phase back emf *VaN*.

7.3- Self-sensing Operation of Saliency Based Rotor Position Tracking

7.3.1 Rotor position estimation

As discussed in Chapter 3, when a superimposed high frequency voltage is injected to the machine with saliency, the resulting current is modulated and the rotor position angle can be derived within the demodulation process of the high frequency current. The current demodulation results are used in this section to verify the self-sensing algorithms



Fig.7.7. Experimental Result-position estimation current demodulation process

for position estimation, and the DC machine is controlled in torque command mode to load up the machine under testing.

The result of high frequency current demodulation process of the IPM machine at rated operation point is shown in Fig.7.7. According to (3.9), the shifted current in $\alpha - \beta$ frame is derived by synchronizing the measured current with the injection voltage vector angle $-\theta_h$. Rotor position signal modulation component in negative sequence is transformed into fundamental average component of the current i'_{α} and i'_{β} . In the meanwhile, the angular frequency of the positive sequence component is shifted to $2\omega_h$, which can be cancelled easily. The estimated angle θ_r is calculated with the filtered current i''_{α} and i''_{β} according to (3.11). Also, since the distribution of inductance in the rotor of PMSM machines is symmetrical, the initial magnet polarity is detected before starting the motor. Shown in Fig.7.8, the self-sensing estimated angle is recorded and compared with the rotor angle measured from an encoder attached on the shaft. It can be concluded that the



Fig.7.8. Experimental Result-Self-sensing estimated position at Motor starting

high frequency voltage injection self-sensing estimation of the rotor position is applicable at extremely low speed, even at zero speed when the rotor is locked before starting.

7.3.2 Vector control of the Self-sensing Drive

As the position signal is not directly derived from the encoder for the self-sensing drive system, some considerations should be taken during the controller design as the system parameters are changed. In this project, the self-sensing drive system contains two control loops, the inner current/torque control loop and the outer speed control loop. Both the loops can be approximately modeled as the general feedback control system shown in Fig.7.9 in which a PI controller is used and the 1st order process is related to the machine electrical model or mechanical model respectively for the inner and the outer loop. It should be also noticed that as the DSP is processing in discretized pattern, the process output is sampled for the feedback signal and the sampling frequency should be taken into account during the controller design process.

Neglecting at first the discrete natures, the closed loop transfer function of the system can be summarized as:

$$G_{cl} = \frac{G(s)}{1+G(s)} = \frac{(k_p s + k_i)k}{s^2 + (p + kk_p)s + kk_i} = \frac{(k_p s + k_i)k}{s^2 + 2\xi_d \omega_n s + \omega_n^2}$$
(7.1)



Fig.7.9. Vector Control Loop

where k_p and k_i are the proportional and integral gains of the PI controller, k and p are related to the process parameters which are previously defined. Hence the task of controller design is to derive the PI gains from the characteristics equation with selected damping factor ξ_d and bandwidth ω_n .

The inner loop of the vector control system is the current loop: the input and output of the process are the voltage and current of the machine respectively. There are two separated current control loops, the d-axis current loop controlling the flux and the q-axis current loop controlling the torque. As for the IPM machine, the electrical parameters are different for these two axes and hence the controllers are turned separately.

The electrical machine dynamic equation in the rotating synchronous reference frame is expressed as:

$$V_{d} = i_{d}R_{s} + L_{d}\frac{di_{d}}{dt} - \omega_{e}i_{q}L_{q}$$

$$V_{q} = i_{q}R_{s} + L_{q}\frac{di_{q}}{dt} + \omega_{e}i_{d}L_{d} + \omega_{e}\lambda_{M}$$
(7.2)

where the last terms are the voltage components related to the rotating flux-linkage. In vector control configuration, these terms are feedforward compensated with the machine mathematical model and the rotor speed. Hence the machine voltage equations contain only the voltage components related to stator resistance and inductance, the transfer functions for the d/q axes are then derived from the stator voltages to the currents:

$$G_{id}(s) = \frac{V_d(s)}{i_d(s)} = \frac{1}{L_d s + R_s}, G_{iq}(s) = \frac{V_q(s)}{i_q(s)} = \frac{1}{L_q s + R_s}$$
(7.3)

Substituted with the PM machine parameters, the process for the d/q axes are expressed as in the alternated form:

$$G_{id}(s) = \frac{V_d(s)}{i_d(s)} = \frac{217.4}{s+173.9}, G_{iq}(s) = \frac{V_q(s)}{i_q(s)} = \frac{140.8}{s+112.7}$$

For the self-sensing drive system, the cut-off frequency or the bandwidth of the closed current loop is selected to be similar to the conventional drive as the estimated rotor position is updating at every sampling point. A reasonable bandwidth is desired to be slower than 10-20 times of the sampling frequency and ω_n is selected so that the current loop bandwidth is defined as 400 Hz, the damping factor ξ is defined as 0.707 for an ideal damping ratio. By taking the design values into (7.1) the gains of the d/q current loop controller are derived:

$$G_{cd}(s) = \frac{19.6s + 4.54 \times 10^4}{s}, \ G_{cq}(s) = \frac{30.7s + 7 \times 10^4}{s}$$

Different from the sensored vector control, one should also be aware that there is a superposed high frequency current resulted by the high voltage frequency injection, hence the frequency of the injected signal should be larger than the current control loop bandwidth to avoid system instability by making sure that the control loop mechanism will not try to track the high frequency current response. For DSP programing, the controller is discretized in to digital form with bilinear method, provided with the sampling frequency of 10kHz.



Fig.7.10. Experimental results- reversed currents response.

$$G_{cd}(z) = \frac{21.9 \, z - 17.36}{z - 1}$$
, $G_{cq}(z) = \frac{34.24 \, z - 27.24}{z - 1}$

The actual response of d/q currents to reversed current commands are shown in Fig 7.10, showing that the measured stator currents are following the reference commands under a step change. The settling time of the current loop is around 1.3ms which is accordance to the theoretical value derived as $4\xi_d/\omega_n$.

In terms of the outer speed loop, the process is constituted by mechanical machine. For isotropic machines the torque equation is expressed as:

$$T_e = K_t i_q = T_L + J \frac{d\omega_r}{dt} + B\omega_r$$
(7.4)

Where K_t is the torque constant, T_L is the load torque *J* and *B* are the inertia and the friction coefficient of the test-rig mechanical system including the shaft and rotor of both the PMSM machine and the load DC machine. The plant transfer function is expressed as:



Fig.7.11. Reverse torque current test.

$$G_{sp}(s) = \frac{\omega_r(s)}{i_q(s)} = \frac{K_t}{Js+B}$$
(7.5)

A reverse torque current test is applied to estimate the mechanical plant parameters as shown in Fig.7.11. The torque constant K_t is derived by measuring the produced torque with a specific torque current (i_q only). Then the inertia and the friction coefficient are calculated from the speed curve under reverse torque current command. The mechanical process for the machine is then derived:

$$G_{sp}(s) = \frac{207.8}{s + 0.059}$$

The bandwidth of the speed loop is selected slower 5-10 times of the position estimation process, the filter bandwidth is 40Hz and the bandwidth is selected as 5Hz, hence the gains are calculated with 100 Hz sampling frequency:

$$G_{csp}(s) = \frac{21.37 \, s + 4.749}{s} \qquad \qquad G_{csp}(z) = \frac{21.37 \, z - 21.35}{z - 1}$$

7.4- Enhanced Self-sensing Feasibility of the prototype machine

7.4.1 Low speed operation with HF voltage injection scheme

The self-sensing position estimation at low speed operation of the prototype machine is similar to that for the conventional machines: the rotor angle is estimated by tracking the stator current response with injected high frequency rotating voltage. Superimposed on fundamental voltage reference derived from the motor control loop, the high frequency voltage signal is injected with fixed amplitude (30V) and frequency (1k Hz) to ensure sufficient position signal is detectable. To analyze the self-sensing performance under loaded situation, machines were operated at constant speed driven by the DC machine: current (torque) reference step changes were then applied to the prototype machine to

examine the position tracking property as shown in Fig.7.12. Demodulation algorithm is implemented in the DSP according to the rotating voltage injection self-sensing mode as discussed in Chapter 5.

The self-sensing performance under machine loaded situation is examined, for each experiment, the machine was driven by the DC machine at constant speed and the current was increased to examine the feasibility of position tracking with increased torque. The noises are rarely influential due the fact that the injection frequency is far higher than the current control bandwidth. The self-sensing estimated rotor position is compared with the encoder measured value and the estimation error is evaluated. One should be aware that the estimation error of rotor position does not have an impact on the current response, while in the current control loop, the accuracy of the current transformation from d-q frame to stationary reference frame is highly related to accurate position estimation, thus a compensation of error keeping into account the current level is required in the self-sensing algorithm to ensure that maximum torque production is achieved for each current value.

For each load step, the stator currents are recorded and shifted forward by the injection frequency in stationary reference frame and processed by DFT to analyze the experimental results in frequency domain. One frequency spectrum of the stator currents is as in Fig.7.12. According to (5.16), all the current components are shifted forward by the carrier frequency (1k Hz). Hence the positive current component i_p is located at 2k Hz. As the motor is rotating at the speed of 60 RPM, the 7Hz fundamental current component for the 14 pole PMSM rotor i_f is shifted to 1007Hz. Saliency modulation current component resulted by the 10 pole self-sensing rotor end i_s contained in negative sequence current is then shifted to the low frequency range for extraction with the frequency of 10Hz.

As shown in Fig.7.13, the position signal strength of self-sensing rotor saliency modulation current under high frequency voltage injection mode is examined. Three experiments were carried out where the prototype machine was operated at 30RPM,



Fig.7.12. Frequency Spectrum of measured stator current with high frequency voltage injection under rated fundamental current when motor speed is 60 RPM.



Fig.7.13. High frequency voltage injection self-sensing rotor saliency modulation position signal strength with increased load under different speeds.

60RPM and 90RPM. For each operation speed, the position signal strength is derived from the shifted stator current DFT analysis and the current amplitudes are reordered for each current (torque) step with the torque increased up to 300% of the rated value. From the results, it can be concluded that the machine operation speed is not an impact factor to the signal strength of saliency modulation current, while all the three curves highlight that the signal strength is increases with torque. The position estimation is based on saliency modulation of the 10 pole self-sensing rotor end, its saliency ratio is unaffected as the reflected fundamental flux is rotating asynchronously. On this basis, the overall machine inductance is decreased when the fundamental current is larger due to iron saturation, thus the inductance saliency caused by the self-sensing rotor end is more apparent. The curve is similar to the machine saturation curve that the increment of signal strength is less significant above 150% load, as the inductance variation is much smaller when machine is saturated. This is a major benefit of the proposed solution: the machine saturation under large torque current makes the position estimation more reliable rather than affecting the self-sensing performance, as in common solutions.

As shown in Fig.7.14, the phase shift of the position signal is derived by comparing the phase of the self-sensing rotor modulation component i_s with the reference signal which is set as zero at no-load. It can be also derived by calculate the difference between the encoder angle and the estimated angle. The phase shift introduced by cross-saturation effect increases with 14 electrical degrees, while the variations of the cross-saturation phase shift is much smaller when the machine operates above the rated torque since the variation range is decreased within 3 electrical degrees.

In Fig.7.15, the self-sensing capability of the prototype machine is compared with the IPM and the Surface Mounted PM machine vs. increased load (in terms of position signal strength). As the values of machine inductances are different, the absolute values of the saliency modulation current amplitude are not appropriate for comparison, since they are dependent on machine overall inductances in which the rotor saliency is directly reflected. Hence the same high frequency voltage signal (30V, 1k Hz) is injected during experiment



Fig.7.14. High frequency voltage injection self-sensing rotor saliency modulation position signal phase shift with increased load under different speeds.



Fig.7.15.High frequency injection self-sensing position signal strength comparison of the prototype machine, IPM and Surface Mounted PM Machine.

and the signal strength ratio (SSR) of position signal amplitude i_s over the positive sequence current i_p is introduced in (7.6) for a fare comparison.

$$SSR = \frac{i_s}{i_p}$$
(7.6)

As shown in Fig.7.15, with increased load, the position signal strength ratios of IPM and SMPM are significantly reduced, since their saliency ratios are heavily affected by machine saturation. Under light load, IPM behaves as a good candidate for self-sensing operation; however, with increased torque, the saliency tends to disappear and the signal strength ratio drops smaller than that of the prototype machine at about 160% of the rated current, this fact limits its self-sensing performance for heavy load operation. For the case of SMPM, the saliency is formed by magnet flux saturation: the saliency ratio is relatively small, and the position signal becomes almost undetectable above two times the rated current when the saliency is very small. On the opposites, for the prototype machine the position signal strength ratio increases slightly with load. Although the increase is not significant, the position signal strength ratio is good for the whole torque range. Different from the absolute current amplitude, the amplitude of positive current i_p also increases as the overall inductance decreased, hence the increase of signal strength ratio is not as significant as the absolute current amplitude in Fig.7.13.

The phase shift of the position signal is also compared for the three machines in Fig.7.16. The prototype motor is less affected by machine cross saturation compared to IPM and SMPM. In addition, the phase variation is very small when machine operates above the rated torque, this feature significantly simplifies the compensation of cross-saturation shift on position estimation algorithm. If the stator current is limited to be larger than its rated value, the machine can be operated in self-sensing control without on-line compensation, only by setting initial constant phase shift compensation.



Fig.7.16.High frequency injection self-sensing position signal phase shift comparison of the prototype machine, IPM and Surface Mounted PM Machine.

7.4.2 High speed operation with fundamental voltage modulation scheme

When the machine operates at high speed so that the amplitude of fundamental voltage vector is sufficient for saliency modulation, the self-sensing position tracking can be realized without superimposed signal injection based on (5.19) as discussed in Chapter 5. The fundamental voltage vector can be seen as the carrier signal, thanks to the electrical asynchronous modulation behavior of the 10 pole self-sensing rotor end. The experimental setup to examine the fundamental voltage modulation self-sensing performance is similar to that of high frequency scheme. The current tracks the referent current closely without much noise, different from the case of high frequency injection. Hence the saliency tracking scheme realized by fundamental voltage modulation is superior in terms of machine vibration, torque ripple and audible noises.

During the experiment, for each load step, DFT analysis is applied to the stator currents in stationary reference frame which is shifted forward by the fundamental frequency, it is the same method used for high frequency injection, yet the carrier signal used in this case is the fundamental voltage. A frequency spectrum under rated speed operation (2400RPM) is reported as example in Fig.7.17. According to (5.19), all the current components are shifted forward by the fundamental frequency (280 Hz). The fundamental current component i_f to drive the 14 pole PMSM rotor is shifted to 560 Hz. The negative fundamental frequency of i_{sf} in (5.19) is then shifted to 0 and the 10 pole self-sensing rotor saliency modulation current component is located at 400 Hz. By adopting the SRFF scheme, the frequency of i_{sf} is always smaller than the frequency of i_f . In position estimation process, i_f is considered as the main noise and the position modulation signal i_{sf} is extracted by low pass filtering.



Fig.7.17. Frequency Spectrum of measured stator current with fundamental voltage modulation when motor speed is 2400 RPM.



Fig.7.18. Position signal Strength with increasing speed under different load by adopting fundamental voltage excitation.

As shown in Fig.7.18, the signal strength is calculated vs increasing speed under different load level: the absolute value of the saliency modulation current signal strength is increasing with speed, since for larger speed, the amplitude of the carrier voltage (fundamental voltage) is larger. If the self-sensing feasible region is defined with the position current signal strength greater than 0.03 A, the fundamental voltage modulation scheme can be adopted when machine speed is larger than 300RPM for rated torque operation. It can be noticed from the graph that the feasible region is wider if the applied load is larger, while if the drive system is installed with high precision current sensors, the feasible region can be further extended.

In Fig. 7.19 (a), the position signal strength is plotted vs. increasing torque, the signal is linearly increasing with torque and the increase is much more significant than the case of



Fig.7.19. Fundamental voltage saliency modulation self-sensing performance with increased load under different speed.(a) Position Signal Strength (b) Position Signal Phase Shift.

high frequency carrier signal injection. This is due to the fact that other than the dominance of rotor end saliency under heavy load, the carrier signal (fundamental voltage) amplitude is also increased under larger torque. Also, for the position signal phase shift angle as in Fig.7.19 (b), the saliency shift trends to become more constant at larger load, hence the difficulty of position estimation error compensation is reduced. In addition, it can be noticed that the smaller phase shift variation at higher speed emphases the superiority of the carrier signals with higher frequency as resistance effect is smaller.

7.5-Summary

In this chapter, the experimental results of self-sensing performances in accordance to the discussions in the previous chapters are presented. The experimental test- rig used for the project is described. A commercial product: Myway Inverter Platform is used as the drive system for the machines under testing: the drive platform is provided with high functionality and hardware integrity, both the vector control and self-sensing algorithms can be easily programed and implemented. On the load side, a DC machine coupled with commercial regenerative drive is used permitting to operate it in 4 quadrants, either in speed control mode or torque control mode. Hence the machines under testing can be operated both in full speed closed loop control and in current (torque) control to examine their self-sensing performances under different load level.

The self-sensing operation of saliency based rotor position tracking method is verified. The results of demodulation process for position estimation are presented and the high frequency rotating voltage injection self-sensing estimation is proved to be accurate at zero and low speed with the implementation errors discussed in Chapter 3 well compensated. Also, the injection frequency should be high enough to avoid any interaction with fundamental current operation.

The enhanced self-sensing feasibility of the prototype machine with the design of saliency modulation rotor end described in Chapter 5 and Chapter 6 is proved with experimental results. The two tracking mechanisms for the saliency modulation of the novel design are

verified. For machine operation at low speed, the rotor end saliency is extracted with conventional high frequency voltage injection method. The position signal strength and its phase shift are examined under different torque/speed tests. Comparing with the IPM machine and the SMPM machine, the proposed machine structure is proved to be superior for self-sensing control in terms of heavy load operation. Under machine saturation condition, its position signal strength is higher and the phase shift variation is smaller thanks to the asynchronous saliency modulation provided by the self-sensing rotor end. The specific saliency modulation tracking scheme at high speed operation termed as fundamental voltage modulation method is also examined. Similar to the case of high frequency excitation, the self-sensing capability is good under heavy load as the two methods are essentially the same. As a matter of fact, the high frequency injection method is a supplement of the fundamental modulation method at low speed when the modulation voltage is not enough for excitation. In general, the prototype machine is provided with better saliency characteristics compared to conventional PMSM machines, and the feasibility of self-sensing operation is increased.

CHAPTER 8 CONCLUSION AND DISCUSSION

8.1- Conclusions

The work in this thesis is focused on improving the capability of PM machine under saliency based self-sensing control. As the machine saliency is seriously affected by saturation effects, the disadvantages of saliency based position tracking mechanism are obvious for conventional PMSMs: in fact, the self-sensing capability is significantly degraded under heavy load operation. Efforts have been taken in previous researches either from the control algorithm perspective, by compensating the cross saturation phase shift and restricting the current operation point in a specific region; or from the machine design perspective, with self-sensing oriented optimization and modification of rotor structure with intentionally introduced saliency features. Nevertheless, the problems are not entirely resolved with these solutions, so that either some compromise is made to sacrifice the machine general performances or the machine is not utilized to its full potential. In this thesis, a novel motor structure is proposed in which the self-sensing position tracking is achieved with saliency modulation signal of the SMRE. Provided with these specific characteristic, the self-sensing operation feasibility is improved as the SMRE saliency is less affected by machine saturation due to increased fundamental flux. Moreover, the fundamental voltage can be utilized as the carrier signal for saliency modulation at higher machine speed.

The self-sensing control technique used in this project is based on high frequency rotating voltage injection. This strategy can be easily implemented and its hardware requirements for signal acquisition and processing are relatively low. The theory of rotor position detection is described along with several issues related to its implementation. Hence, the sources of estimated position angle deviations are related only to the PM Machine.

For conventional PMSM machines, self-sensing performances degrades with increased torque. In this thesis, an IPM machine and a SMPM machine are considered for demonstration with FEM analysis. The incremental inductances are measured according to the variation of flux-linkage caused by small incremental current. Together with the

CONCLUSION

experimental tests in Chapter 7, their self-sensing capabilities with increased torque current are analyzed. The IPM machines with strong saliency ratios are suitable for saliency tracking at normal load level, however with increased load, the q-axis inductance decreases quickly making the position signal hard to detect; for the case of SMPM machines, although the inductance variations are less significant, the saliency is barely detectable along the whole operation range. In terms of cross-saturation phase shift, both types of machine are then not ideal since the shift angle is large and the trend is nonlinear.

The design aspects of a PMSM with the novel configuration of SMRE are provided in, a further optimization process is also developed for optimal position signal quality. With such a technique, the feasibility region of self-sensing control is extended. For high frequency injection at low speed operation, it has been demonstrated that the position estimation performance based on saliency modulation of the rotor end is not affected by machine saturation, due to its asynchronous operation. In fact, as demonstrated also by experimental tests, the performance is even better at high current level, the position signal strength is higher and cross-saturation phase shift becomes more constant. For medium and high speed range, saliency modulation is realized without HF injection, which can be used to substitute the conventional model based methods.

8.2-Future work

Although the self-sensing feasibility of PMSM machines can be significantly improved with the novel configuration of the saliency modulation rotor end, there are some aspects in the technique that would benefit from future development. Firstly, a more comprehensive optimization process is desired that properly weighs the machine general performances and the self-sensing capabilities. The optimal balance highly depends on the specific requirements for different applications: for low cost self-sensing drives, where the quality of current measurement and signal processing are limited, a strong position signal is desired, and this can be achieved either by increasing the SMRE section length or designing its profile to be more salient with a thinner air-gap. In case of high standard drives where lower torque ripple is expected, the harmonic component caused by

CONCLUSION

SMRE should be reduced with more "smooth" design of SMRE, yet conversely the position signal becomes smaller and more measuring/processing efforts are required. Additional loss caused by SMRE should be also considered: this may not be a serious problem as the total flux in the end section is small without PM excitations, however, for high speed drives, considerations should be taken in design process to reduce harmonic components.

BIBLIOGRAPHY

- [1] B. Felix, Method For Controlling Asynchronous Machines, 1974.
- [2] D. W. Novotny and T. A. Lipo, *Vector Control and Dynamics of AC Drives*. Oxford University Press, 1996.
- [3] *Field Oriented Control of 3-Phase AC-Motors*. Texas Instruments Europe, 1998.
- [4] G. Asher, Advanced AC drives: Msc Modules H64AMD course notes.
- [5] C. Gerada, *Advanced Electrical Machines*: Msc Modules H64AEM course notes.
- [6] K. Sungmin, Y. Young-Doo, S. Seung-Ki, and K. Ide, "Maximum Torque per Ampere (MTPA) Control of an IPM Machine Based on Signal Injection Considering Inductance Saturation," *Power Electronics, IEEE Transactions on,* vol. 28, pp. 488-497, 2013.
- [7] M. Ned, Advanced electric drives: analysis, control, and modeling using MATLAB/Simulink. Hoboken, New Jersey: John Wiley & Sons, 2014.
- [8] J. F. Gieras and M. Wing, *Permanent Magnet Motor Technology: Design and Applications*, 3rd Edition ed. New York Basel: Marcel Dekker, 2010.
- [9] Y. Hua, "Sensorless Control of Surface Mounted Permanant Magnet Machine Using Fundamental PWM Excitation," PHD, University of Nottingham, 2009.
- [10] W. Gaolin, Y. Rongfeng, W. Yangwei, Y. Yong, and X. Dianguo, "Initial rotor position estimation for sensorless interior PMSM with signal injection," in *Power Electronics Conference (IPEC), 2010 International*, 2010, pp. 2748-2752.
- [11] N. Bianchi and S. Bolognani, "Influence of rotor geometry of an interior PM motor on sensorless control feasibility," in *Industry Applications Conference, 2005. Fourtieth IAS Annual Meeting. Conference Record of the 2005,* 2005, pp. 2553-2560 Vol. 4.
- [12] P. Sergeant, F. De Belie, and J. Melkebeek, "Effect of Rotor Geometry and Magnetic Saturation in Sensorless Control of PM Synchronous Machines," *Magnetics, IEEE Transactions on*, vol. 45, pp. 1756-1759, 2009.
- [13] N. Bianchi, E. Fornasiero, and S. Bolognani, "Effect of Stator and Rotor Saturation on Sensorless Rotor Position Detection," *leee Transactions on Industry Applications*, vol. 49, pp. 1333-1342, May-Jun 2013.
- [14] F. Blaschke, J. van der Burgt, and A. Vandenput, "Sensorless direct field orientation at zero flux frequency," in *Industry Applications Conference, 1996. Thirty-First IAS Annual Meeting, IAS '96., Conference Record of the 1996 IEEE*, 1996, pp. 189-196 vol.1.
- [15] P. L. Jansen and R. D. Lorenz, "Transducerless position and velocity estimation in induction and salient AC machines," *Industry Applications, IEEE Transactions on*, vol. 31, pp. 240-247, 1995.
- [16] J. Yu-Seok, R. D. Lorenz, T. M. Jahns, and S. Seung-Ki, "Initial rotor position estimation of an interior permanent-magnet synchronous machine using carrier-frequency injection methods," *Industry Applications, IEEE Transactions on*, vol. 41, pp. 38-45, 2005.
- [17] J. Cilia, G. M. Asher, K. J. Bradley, and M. Sumner, "Sensorless position detection for vectorcontrolled induction motor drives using an asymmetric outer-section cage," *Industry Applications, IEEE Transactions on*, vol. 33, pp. 1162-1169, 1997.
- [18] N. Teske, G. M. Asher, M. Sumner, and K. J. Bradley, "Analysis and suppression of high-frequency inverter modulation in sensorless position-controlled induction machine drives," *Industry Applications, IEEE Transactions on*, vol. 39, pp. 10-18, 2003.
- [19] M. W. Degner and R. D. Lorenz, "Using multiple saliencies for the estimation of flux, position, and velocity in AC machines," *Industry Applications, IEEE Transactions on*, vol. 34, pp. 1097-1104, 1998.
- [20] M. W. Degner and R. D. Lorenz, "Position estimation in induction machines utilizing rotor bar slot harmonics and carrier-frequency signal injection," *Industry Applications, IEEE Transactions on*, vol. 36, pp. 736-742, 2000.
- [21] N. Teske, G. M. Asher, M. Sumner, and K. J. Bradley, "Suppression of saturation saliency effects

- for the sensorless position control of induction motor drives under loaded conditions," *Industrial Electronics, IEEE Transactions on,* vol. 47, pp. 1142-1150, 2000.
- [22] J. Ji-Hoon, S. Seung-Ki, H. Jung-Ik, K. Ide, and M. Sawamura, "Sensorless drive of surface-mounted permanent-magnet motor by high-frequency signal injection based on magnetic saliency," *Industry Applications, IEEE Transactions on,* vol. 39, pp. 1031-1039, 2003.
- [23] J. Holtz, "Acquisition of Position Error and Magnet Polarity for Sensorless Control of PM Synchronous Machines," *Industry Applications, IEEE Transactions on,* vol. 44, pp. 1172-1180, 2008.
- [24] M. J. Corley and R. D. Lorenz, "Rotor position and velocity estimation for a salient-pole permanent magnet synchronous machine at standstill and high speeds," *Industry Applications, IEEE Transactions on,* vol. 34, pp. 784-789, 1998.
- [25] T. Aihara, A. Toba, T. Yanase, A. Mashimo, and K. Endo, "Sensorless torque control of salient-pole synchronous motor at zero-speed operation," *Power Electronics, IEEE Transactions on*, vol. 14, pp. 202-208, 1999.
- [26] M. Linke, R. Kennel, and J. Holtz, "Sensorless speed and position control of synchronous machines using alternating carrier injection," in *Electric Machines and Drives Conference, 2003. IEMDC'03. IEEE International*, 2003, pp. 1211-1217 vol.2.
- [27] H. Jung-Ik and S. Seung-Ki, "Sensorless field-orientation control of an induction machine by high-frequency signal injection," *Industry Applications, IEEE Transactions on,* vol. 35, pp. 45-51, 1999.
- [28] P. Garcia, F. Briz, M. W. Degner, and D. Diaz-Reigosa, "Accuracy, Bandwidth, and Stability Limits of Carrier-Signal-Injection-Based Sensorless Control Methods," *Industry Applications, IEEE Transactions on*, vol. 43, pp. 990-1000, 2007.
- [29] H. Jung-Ik and S. Seung-Ki, "Physical understanding of high frequency injection method to sensorless drives of an induction machine," in *Industry Applications Conference, 2000. Conference Record of the 2000 IEEE*, 2000, pp. 1802-1808 vol.3.
- [30] J. M. Guerrero, M. Leetmaa, F. Briz, A. Zamarron, and R. D. Lorenz, "Inverter nonlinearity effects in high-frequency signal-injection-based sensorless control methods," *Industry Applications, IEEE Transactions on*, vol. 41, pp. 618-626, 2005.
- [31] D. Raca, P. Garcia, D. Reigosa, F. Briz, and R. Lorenz, "A comparative analysis of pulsating vs. rotating vector carrier signal injection-based sensorless control," in *Applied Power Electronics Conference and Exposition, 2008. APEC 2008. Twenty-Third Annual IEEE*, 2008, pp. 879-885.
- [32] K. Hyunbae and R. D. Lorenz, "Carrier signal injection based sensorless control methods for IPM synchronous machine drives," in *Industry Applications Conference, 2004. 39th IAS Annual Meeting. Conference Record of the 2004 IEEE,* 2004, pp. 977-984 vol.2.
- [33] M. Schroedl, "Sensorless control of AC machines at low speed and standstill based on the "INFORM" method," in *Industry Applications Conference, 1996. Thirty-First IAS Annual Meeting, IAS '96., Conference Record of the 1996 IEEE*, 1996, pp. 270-277 vol.1.
- [34] Q. Gao, G. M. Asher, M. Sumner, and P. Makys, "Position Estimation of AC Machines Over a Wide Frequency Range Based on Space Vector PWM Excitation," *Industry Applications, IEEE Transactions on*, vol. 43, pp. 1001-1011, 2007.
- [35] C. Caruana, G. M. Asher, and J. C. Clare, "Sensorless flux position estimation at low and zero frequency by measuring zero-sequence current in delta-connected cage induction machines," *Industry Applications, IEEE Transactions on,* vol. 41, pp. 609-617, 2005.
- [36] E. Robeischl and M. Schroedl, "Optimized INFORM measurement sequence for sensorless PM synchronous motor drives with respect to minimum current distortion," *Industry Applications, IEEE Transactions on,* vol. 40, pp. 591-598, 2004.
- [37] G. Dajaku and D. Gerling, "The correct analytical expression for the phase inductance of salient pole machines," in *Electric Machines & Drives Conference, 2007. IEMDC '07. IEEE International*, 2007, pp. 992-996.
- [38] J. Holtz and P. Hangwen, "Acquisition of rotor anisotropy signals in sensorless position control systems," *Industry Applications, IEEE Transactions on,* vol. 40, pp. 1379-1387, 2004.
- [39] T. M. Wolbank and J. Machl, "A modified PWM scheme in order to obtain spatial information of AC machines without mechanical sensor," in *Applied Power Electronics Conference and Exposition*, 2002. APEC 2002. Seventeenth Annual IEEE, 2002, pp. 310-315 vol.1.

- [40] M. A. Vogelsberger, M. Riepler, S. Grubic, T. G. Habetler, and T. M. Wolbank, "Integration of transient and fundamental wave excitation for zero speed sensorless control of AC machines," in *Electrical Machines, 2008. ICEM 2008. 18th International Conference on*, 2008, pp. 1-6.
- [41] T. M. Wolbank, H. Giuliani, R. Woehrnschimmel, and J. L. Machl, "Sensorless control of induction machines by combining fundamental wave models with transient excitation technique," in *Electric Machines and Drives, 2005 IEEE International Conference on*, 2005, pp. 1379-1384.
- [42] K. S. Saleh, "Sensorless Control of High Power Induction Motors Using Multilevel Converters," PhD, University of Nottingham, 2010.
- [43] J. Holtz and J. Juliet, "Sensorless acquisition of the rotor position angle of induction motors with arbitrary stator windings," *Industry Applications, IEEE Transactions on,* vol. 41, pp. 1675-1682, 2005.
- [44] P. Guglielmi, M. Pastorelli, and A. Vagati, "Impact of cross-saturation in sensorless control of transverse-laminated synchronous reluctance motors," *IEEE Transactions on Industrial Electronics*, vol. 53, pp. 429-439, 2006.
- [45] N. Bianchi, S. Bolognani, J. H. Jang, and S. K. Sul, "Comparison of PM Motor Structures and Sensorless Control Techniques for Zero-Speed Rotor Position Detection," *IEEE Transactions on Power Electronics*, vol. 22, pp. 2466-2475, 2007.
- [46] Z. Q. Zhu and L. M. Gong, "Investigation of Effectiveness of Sensorless Operation in Carrier-Signal-Injection-Based Sensorless-Control Methods," *IEEE Transactions on Industrial Electronics*, vol. 58, pp. 3431-3439, 2011.
- [47] P. Guglielmi, M. Pastorelli, and A. Vagati, "Cross-Saturation Effects in IPM Motors and Related Impact on Sensorless Control," *IEEE Transactions on Industry Applications*, vol. 42, pp. 1516-1522, 2006.
- [48] P. Sergeant, F. D. Belie, and J. Melkebeek, "Effect of Rotor Geometry and Magnetic Saturation in Sensorless Control of PM Synchronous Machines," *IEEE Transactions on Magnetics*, vol. 45, pp. 1756-1759, 2009.
- [49] N. B. M. I. a. S. B. Bianchi and Bolognani, "Sensorless-Oriented Design of PM Motors," *IEEE Transactions on Industry Applications,* vol. 45, pp. 1249-1257, 2009.
- [50] P. Sergeant, F. D. Belie, and J. Melkebeek, "Rotor Geometry Design of Interior PMSMs With and Without Flux Barriers for More Accurate Sensorless Control," *IEEE Transactions on Industrial Electronics*, vol. 59, pp. 2457-2465, 2012.
- [51] R. Wrobel, A. S. Budden, D. Salt, D. Holliday, P. H. Mellor, A. Dinu, et al., "Rotor Design for Sensorless Position Estimation in Permanent-Magnet Machines," *IEEE Transactions on Industrial Electronics*, vol. 58, pp. 3815-3824, 2011.
- [52] P. Guglielmi, M. Pastorelli, G. Pellegrino, and A. Vagati, "Position-sensorless control of permanent-magnet-assisted synchronous reluctance motor," *IEEE Transactions on Industry Applications*, vol. 40, pp. 615-622, 2004.
- [53] R. Raute, C. Caruana, C. S. Staines, J. Cilia, M. Sumner, and G. M. Asher, "Analysis and Compensation of Inverter Nonlinearity Effect on a Sensorless PMSM Drive at Very Low and Zero Speed Operation," *IEEE Transactions on Industrial Electronics*, vol. 57, pp. 4065-4074, 2010.
- [54] N. Bianchi, S. Bolognani, J. Ji-Hoon, and S. Seung-Ki, "Advantages of Inset PM Machines for Zero-Speed Sensorless Position Detection," *Industry Applications, IEEE Transactions on*, vol. 44, pp. 1190-1198, 2008.
- [55] N. Bianchi and S. Bolognani, "Sensorless-Oriented Design of PM Motors," *Industry Applications, IEEE Transactions on,* vol. 45, pp. 1249-1257, 2009.
- [56] A. S. Budden, R. Wrobel, D. Holliday, P. H. Mellor, A. Dinu, P. Sangha, et al., "Impact of Rotor Design on Sensorless Position Estimation," in *IEEE Industrial Electronics, IECON 2006 - 32nd Annual Conference on*, 2006, pp. 787-792.
- [57] R. Wrobel, A. S. Budden, D. Salt, D. Holliday, P. H. Mellor, A. Dinu, et al., "Rotor Design for Sensorless Position Estimation in Permanent-Magnet Machines," *Industrial Electronics, IEEE Transactions on*, vol. 58, pp. 3815-3824, 2011.
- [58] Y. Kano, T. Kosaka, N. Matsui, and T. Nakanishi, "Sensorless-oriented design of IPM motors for general industrial applications," in *Electrical Machines, 2008. ICEM 2008. 18th International*



	48, pp. 2157-2164, 2012.
[77]	Y. Chen-Yen, J. Tamura, D. D. Reigosa, and R. D. Lorenz, "Position Self-Sensing Evaluation of a FI- IPMSM Based on High-Frequency Signal Injection Methods," <i>Industry Applications, IEEE</i>
[78]	<i>Transactions on,</i> vol. 49, pp. 880-888, 2013. T. A. Nondahl, C. Ray, P. B. Schmidt, and M. L. Gasperi, "A permanent-magnet rotor containing an electrical winding to improve detection of rotor angular position," <i>Industry Applications, IEEE Transactions on</i> , vol. 35, pp. 819-824, 1999.
[79]	N. Bianchi, S. Bolognani, and A. Faggion, "A ringed-pole SPM motor for sensorless drives - electromagnetic analysis, prototyping and tests," in <i>Industrial Electronics (ISIE), 2010 IEEE</i> International Symposium on, 2010, pp. 1193-1198.
[80]	A. Faggion, N. Bianchi, and S. Bolognani, "Ringed-Pole Permanent-Magnet Synchronous Motor for Position Sensorless Drives," <i>Industry Applications, IEEE Transactions on,</i> vol. 47, pp. 1759-1766, 2011.
[81]	 A. Faggion, E. Fornasiero, N. Bianchi, and S. Bolognani, "Sensorless Capability of Fractional-Slot Surface-Mounted PM Motors," <i>Industry Applications, IEEE Transactions on</i>, vol. 49, pp. 1325-1332, 2013.
[82]	J. Graus, A. Rambetius, and I. Hahn, "Comparison of the resistance- and inductance-based saliency of a PMSM due to a short-circuited rotor winding," in <i>Power Electronics Conference</i> (<i>IPEC-Hiroshima 2014 - ECCE-ASIA</i>), 2014 International, 2014, pp. 270-277.
[83]	J. Graus and I. Hahn, "Modelling and optimization of a short-circuited rotor winding of a PMSM for saliency tracking," in <i>Sensorless Control for Electrical Drives (SLED), 2014 IEEE 5th International Symposium on</i> , 2014, pp. 1-8.
[84]	L. Alberti, N. Bianchi, M. Morandin, and S. Bolognani, "Analysis and Tests of the Sensorless Rotor Position Detection of Ringed-Pole Permanent-Magnet Motor," <i>Industry Applications, IEEE Transactions on</i> , vol. 50, pp. 3278-3284, 2014.
[85]	L. Alberti, N. Bianchi, M. Morandin, and J. J. C. Gyselinck, "Finite-Element Analysis of Electrical Machines for Sensorless Drives With High-Frequency Signal Injection," <i>Industry Applications, IEEE Transactions on</i> , vol. 50, pp. 1871-1879, 2014.
[86]	 M. Morandin, A. Faggion, and S. Bolognani, "Integrated Starter–Alternator With Sensorless Ringed-Pole PM Synchronous Motor Drive," <i>Industry Applications, IEEE Transactions</i> on, vol. 51, pp. 1485-1493, 2015.
[87]	M. Morandin, S. Bolognani, and A. Faggion, "Outer-rotor ringed-pole SPM starter-alternator suited for sensorless drives," in <i>Sensorless Control for Electrical Drives (SLED), 2011 Symposium on</i> , 2011, pp. 96-101.
[88]	J. Graus and I. Hahn, "Sensorless capability of an interior permanent magnet synchronous machine with a short-circuited rotor winding," in <i>Industrial Electronics Society, IECON 2014 - 40th</i> Annual Conference of the IEEE, 2014, pp. 831-837.
[89]	D. Mingardi, E. Fornasiero, N. Bianchi, S. Bolognani, and A. Faggion, "Ring Losses Evaluation in Ringed-Pole PM Motors," Industry Applications, IEEE Transactions on, vol. 51, pp. 3686-3695, 2015.
[90]	S. Buso and P. Mattavelli, "Digital Control in Power Electronics," in International Power Electronics Congress, IEEE, 2006, pp. A-190-A-190.
[91]	Y. Li, Z. Q. Zhu, D. Howe, C. M. Bingham and D. A. Stone, "Improved Rotor-Position Estimation by Signal Injection in Brushless AC Motors, Accounting for Cross-Coupling Magnetic Saturation," in <i>IEEE Transactions on Industry Applications</i> , vol. 45, no. 5, pp. 1843-1850, Septoct. 2009.
[92]	Z. Q. Zhu, "Fractional slot permanent magnet brushless machines and drives for electric and hybrid propulsion systems," Compel International Journal of Computations & Mathematics in Electrical, vol. 30, pp. 9-31(23), 2011.
[93]	X. Ge and Z. Q. Zhu, "A Novel Design of Rotor Contour for Variable Reluctance Resolver by Injecting Auxiliary Air-Gap Permeance Harmonics," IEEE Transactions on Energy Conversion, vol. 31, 10.1109/TEC.2015.2470546, pp. 345-353, 2016.
[94]	Myway platform datasheet: Inverter Unit MWINV-9R144 Hardware User's Manual
	163

APPENDIX

$$\begin{split} &V_{a} = pi_{d}L_{d}\cos\theta - pi_{q}L_{q}\sin\theta \\ &= pi_{d}L_{d}\cos\theta(\cos^{2}\theta + \sin^{2}\theta) - pi_{q}L_{q}\sin\theta(\cos^{2}\theta + \sin^{2}\theta) \\ &= pi_{d}[L_{q}(\sin^{2}\theta\cos\theta - \sin^{2}\theta\cos\theta) + L_{d}(\cos^{3}\theta + \sin^{2}\theta)] \\ &\quad + pi_{q}[L_{q}(-\sin^{3}\theta - \sin\theta\cos^{2}\theta) \\ &\quad + L_{d}(\sin\theta\cos^{2} - \sin\theta\cos^{2})] \\ &= p[i_{d}\cos\theta - i_{q}\sin\theta][L_{q}\sin^{2}\theta + L_{d}\cos^{2}\theta] \\ &\quad - p[i_{d}\sin\theta + i_{q}\cos\theta]\left[\left(\frac{L_{q} - L_{d}}{2}\right)2\cos\theta\sin\theta\right] \\ &= p[i_{d}\cos\theta - i_{q}\sin\theta]\left[\frac{L_{q}}{2}(2 - 2\cos^{2}\theta) + \frac{L_{d}}{2}(2\cos^{2}\theta - 1 + 1)\right] \\ &\quad - p[i_{d}\sin\theta + i_{q}\cos\theta]\left[\left(\frac{L_{q} - L_{d}}{2}\right)\sin2\theta\right] \\ &= p[i_{d}\cos\theta - i_{q}\sin\theta]\left[\frac{L_{q}}{2}(1 - \cos2\theta) + \frac{L_{d}}{2}(1 + \cos2\theta)\right] \\ &\quad - p[i_{d}\sin\theta + i_{q}\cos\theta]\left[\left(\frac{L_{q} - L_{d}}{2}\right)\sin2\theta\right] \\ &= p[i_{d}\cos\theta - i_{q}\sin\theta]\left[\left(\frac{L_{q} + L_{d}}{2}\right) - \left(\frac{L_{q} - L_{d}}{2}\right)\cos2\theta\right] \\ &\quad - p[i_{d}\sin\theta + i_{q}\cos\theta]\left[\left(\frac{L_{q} - L_{d}}{2}\right)\cos2\theta\right] \\ &\quad - p[i_{d}\sin\theta + i_{q}\cos\theta]\left[\left(\frac{L_{q} - L_{d}}{2}\right)\cos2\theta\right] \\ &\quad - p[i_{d}\sin\theta + i_{q}\cos\theta]\left[\left(\frac{L_{q} - L_{d}}{2}\right)\sin2\theta\right] \end{split}$$

Since

$$i_{\alpha} = i_{d}\cos\theta - i_{q}\sin\theta \qquad i_{\beta} = i_{d}\sin\theta + i_{q}\cos\theta$$

$$\Sigma L_{s} = (L_{q} + L_{d})/2 \qquad \Delta L_{s} = (L_{q} - L_{d})/2$$

$$\therefore V_{\alpha} = pi_{\alpha}[\Sigma L_{s} - \Delta L_{s}\cos 2\theta] - pi_{\beta}\Delta L_{s}\sin 2\theta$$