SENSORLESS CONTROL OF AC MACHINES FOR INTEGRATED STARTER GENERATOR APPLICATION

DISSERTATION

Presented in Partial Fulfillment of the Requirements for
the Degree Doctor of Philosophy in the
Graduate School of The Ohio State University

By

Jiangang Hu, M.S.

* * * * *

The Ohio State University

2007

Dissertation Committee:
Professor Longya Xu, Adviser
Professor Donald Kasten
Professor Stephen Sebo

Approved by

Adviser
Graduate Program in Electrical and Computer Engineering
Using a single AC machine in an Integrated Starter Generator (ISG) system, becomes popular for automotive, aircraft and other industrial applications. This dissertation focuses on sensorless control algorithms of AC machines over a wide speed range for ISG applications.

High frequency current responses of a permanent magnetic (PM) synchronous machine with and without considering the eddy current effects are considerably different. This dissertation identifies and investigates the eddy current effects on rotor position estimation and contributes to an improved high frequency injection (HFI) based sensorless control of a PMSM at zero and low speeds. In addition, an improved magnetic pole identification method based on space vector PWM for arbitrary initial rotor position estimation and sensorless control of PM synchronous machines is presented. Through applying vector controlled pilot voltages by SVPWM, the N-S pole can be identified at any initial rotor positions without rotor alignment actions. Because of the limitation of HFI in the high speed range, rotor position estimation based on Back-EMF is also proposed in this dissertation for PMSM operating in high speed ranges.

A voltage model based rotor position observer for induction machine ISG is presented and used for test validation. The method works well at very low speed regions where full torque is required. Test results with full starting torque over a wide speed range are presented based on the proposed rotor position observer. A new sensorless control method
in deep flux weakening region is presented. AC phase voltages reach their limits in the generating mode because of limited DC bus voltage in the ISG system. Only slip can be controlled through DC bus voltage PI controller. Accordingly, the controller maintains the constant DC bus voltage through adjusting the slip speed. The proposed methods can ensure the effective estimation of rotor position in the starting mode and generating mode.

The validity of the proposed methods has been proven through computer simulation and experimental results.
Dedicated to my dear wife, Jingbo, my daughter, Amy, and beloved parents.
ACKNOWLEDGMENTS

First of all, I would like to express my gratitude to my adviser, Professor Longya Xu, for providing an opportunity to perform research in Power Electronics and Electric Machine (PEEM) Group. Without his guidance and encouragement, this dissertation would not have been possible. He has helped me earn a lot of knowledge, and has given me confidence in my field of study.

I would like to thank Professor Donald Kasten and Professor Stephen Sebo for being my dissertation committee members. They helped me review and revise my dissertation. Also they taught me technical writing skills.

I would like to thank my colleagues and friends for their help and the fun that they brought to me. And I really enjoy the time I spent with them: Dr. Shengming Li, Dr. Jingchuan Li, Dr. Song Chi, Dr. Wenzhe Lu, Dr. Min Dai, Dr. Xin Liu, Ms. Yuan Zhang, Mr. Bo Guan. I collaborated with most of them in different projects and the exciting discussions with them always broadened my knowledge in the field. Also I would like to thank other professors and staff members in Department of Electrical and computer Engineering. Their support and friendship mean a lot to me in my life at The Ohio State University.

Finally, I want to thank my dear wife, Jingbo Liu, my parents for their love and dedication that have supported me to go so far, and their great efforts to build up my confidence and love of life. I am truly grateful for having such a happy and supportive family.
VITA

January 24, 1974 ................................. Born - Laizhou, Shandong, China

June, 1997 ................................. B.S., Electrical Engineering,
Huazhong University of Science and Technology, Wuhan, Hubei, China

June, 2000 ................................. M.S., Electrical Engineering,
Huazhong University of Science and Technology, Wuhan, Hubei, China

September, 2002 - June, 2007 ........................ Graduate Research Associate,
The Ohio State University, Columbus, OH

June, 2007 - November, 2007 ........................ Co-op,
Rockwell Automation, Mequon, WI

November, 2007 - Present ........................ Sr. Development Engineer,
Rockwell Automation, Mequon, WI

PUBLICATIONS


**FIELDS OF STUDY**

Major Field: Electrical Engineering

Studies in power electronics, electric machinery and control: Professor Longya Xu
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Chapter</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Introduction</td>
<td>1</td>
</tr>
<tr>
<td>1.1</td>
<td>Energy Conversion</td>
<td>1</td>
</tr>
<tr>
<td>1.2</td>
<td>Integrated Starter Generator</td>
<td>3</td>
</tr>
<tr>
<td>1.3</td>
<td>Sensorless Technology for AC Electric Machine</td>
<td>4</td>
</tr>
<tr>
<td>1.3.1</td>
<td>Review of Sensorless Control for Synchronous Machine</td>
<td>5</td>
</tr>
<tr>
<td>1.3.2</td>
<td>Review of Sensorless Control for Induction Machine</td>
<td>7</td>
</tr>
<tr>
<td>1.4</td>
<td>DSP Based Electric Drive System</td>
<td>11</td>
</tr>
<tr>
<td>1.5</td>
<td>Contributions of the Dissertation</td>
<td>12</td>
</tr>
<tr>
<td>1.6</td>
<td>Dissertation Organization</td>
<td>13</td>
</tr>
</tbody>
</table>

2. Dynamic Model of Electric Machines | 15 |

2.1 Reference Frame Theory | 15 |
2.1.1 Equations of Transformation: Change of Variables | 17 |
2.1.2 Stationary Circuit Variables Transformed to the Arbitrary Reference Frame | 18 |
2.1.3 Transformations in Electric Machines ........................................ 20
2.2 Dynamic Model of Synchronous Machines .................................... 23
2.3 Dynamic Model of Induction Machines ......................................... 31
  2.3.1 Dynamic Model in Synchronous Reference Frame ...................... 31
  2.3.2 Dynamic Model in Stationary Reference Frame ....................... 32
  2.3.3 Dynamic Model in Arbitrary Reference Frame .......................... 34
2.4 Comparisons between Permanent Magnetic Synchronous Machine and
  Induction Machine ........................................................................ 36

3. Sensorless Control for Permanent Magnet Synchronous Machine Based Integrated Starter Generator .................................................. 40
  3.1 High Frequency Injection Based Sensorless Control Technology for PM
      Synchronous Machine .................................................................. 40
  3.2 Modeling of PM Synchronous Machine Considering Eddy Current Effects 48
  3.3 Identifying Permanent Magnet Polarity ............................................ 52
  3.4 Simulation and Experimental Results of High Frequency Injection and
      Permanent Magnet Polarity Identification ...................................... 58
  3.5 Synchronous Machine Voltage Model Sensorless Technology ............. 65
  3.6 Experimental Results of Voltage Model Sensorless Algorithm .......... 69
  3.7 Full Speed Range Sensorless Control for PMSM ............................ 71

4. Sensorless Control for Induction Machine Based Integrated Starter Generator ........................................... 73
  4.1 IM ISG System Hardware Design ................................................. 73
  4.2 Transformation Between Y-Connection And Delta-Connection ............ 80
  4.3 Sensorless Control Algorithm for Starting Mode ............................. 86
  4.4 Sensorless Control Algorithm for Generating Mode ........................ 87
    4.4.1 Negative \( I_{q,\text{ref}} \) Command .............................................. 89
    4.4.2 \( I_{d,\text{ref}} \) Command Reduction in Generating Mode ................. 92
    4.4.3 Slip Control in Generating Mode ........................................... 97
  4.5 Induction Machine ISG Torque-Speed Characteristics ....................... 108
  4.6 Induction Machine Performance Testing Results ........................... 113

5. Conclusions and Future Work .......................................................... 117
  5.1 Conclusions ............................................................................... 117
  5.2 Future Work ............................................................................. 118

Appendices:
A. Digital Filters for the High Frequency Injection Method ........................................ 120  
B. List of Abbreviations .................................................................................................... 124  
C. List of Symbols ........................................................................................................... 127  
Bibliography ..................................................................................................................... 133
# LIST OF TABLES

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1 Comparison of drive systems</td>
<td>10</td>
</tr>
<tr>
<td>1.2 Feature comparison between TMS320F2812 and TMS320F2407</td>
<td>12</td>
</tr>
<tr>
<td>2.1 Mechanical comparison between IM and PMSM</td>
<td>38</td>
</tr>
<tr>
<td>2.2 Magnetic comparison between IM and PMSM</td>
<td>39</td>
</tr>
<tr>
<td>2.3 Electrical comparison between IM and PMSM</td>
<td>39</td>
</tr>
<tr>
<td>4.1 ISG system specifications</td>
<td>75</td>
</tr>
<tr>
<td>4.2 Parameters of the induction machine</td>
<td>79</td>
</tr>
</tbody>
</table>
# LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Project $abc$ to $\alpha\beta$ coordinates</td>
<td>21</td>
</tr>
<tr>
<td>2.2</td>
<td>Project $dq$ to $\alpha\beta$ coordinates</td>
<td>22</td>
</tr>
<tr>
<td>2.3</td>
<td>Basic representation of a salient pole synchronous machine with damper windings</td>
<td>24</td>
</tr>
<tr>
<td>2.4</td>
<td>Reference frames</td>
<td>27</td>
</tr>
<tr>
<td>2.5</td>
<td>Salient pole synchronous machine equivalent circuit in $dq$ reference frame</td>
<td>29</td>
</tr>
<tr>
<td>3.1</td>
<td>Rotor position estimation based on high frequency injection method</td>
<td>42</td>
</tr>
<tr>
<td>3.2</td>
<td>Simulation results of $i_\alpha$ and $i_\beta$</td>
<td>43</td>
</tr>
<tr>
<td>3.3</td>
<td>$i_{\alpha 1}$ and $i_{\beta 1}$</td>
<td>44</td>
</tr>
<tr>
<td>3.4</td>
<td>Details of $i_{\alpha 1}$ and $i_{\beta 1}$</td>
<td>44</td>
</tr>
<tr>
<td>3.5</td>
<td>$i_{\alpha 2}$ and $i_{\beta 2}$</td>
<td>45</td>
</tr>
<tr>
<td>3.6</td>
<td>Details of $i_{\alpha 2}$ and $i_{\beta 2}$</td>
<td>45</td>
</tr>
<tr>
<td>3.7</td>
<td>$i_{\alpha 3}$ and $i_{\beta 3}$</td>
<td>46</td>
</tr>
<tr>
<td>3.8</td>
<td>Details of $i_{\alpha 3}$ and $i_{\beta 3}$</td>
<td>46</td>
</tr>
<tr>
<td>3.9</td>
<td>$i_{\alpha_{dem}}$ and $i_{\beta_{dem}}$</td>
<td>47</td>
</tr>
<tr>
<td>3.10</td>
<td>$2\dot{\theta}_r$</td>
<td>47</td>
</tr>
<tr>
<td>Section</td>
<td>Title</td>
<td>Page</td>
</tr>
<tr>
<td>---------</td>
<td>----------------------------------------------------------------------</td>
<td>------</td>
</tr>
<tr>
<td>3.11</td>
<td>PM synchronous machine with shorted coils accounting for eddy currents</td>
<td>49</td>
</tr>
<tr>
<td>3.12</td>
<td>Rotor position estimation with eddy current effect compensation</td>
<td>50</td>
</tr>
<tr>
<td>3.13</td>
<td>Eddy current effect compensation</td>
<td>52</td>
</tr>
<tr>
<td>3.14</td>
<td>Magnetic pole identification based on saturation effects</td>
<td>53</td>
</tr>
<tr>
<td>3.15</td>
<td>System diagram for rotor position estimation</td>
<td>55</td>
</tr>
<tr>
<td>3.16</td>
<td>Space vectors</td>
<td>56</td>
</tr>
<tr>
<td>3.17</td>
<td>The pilot voltage pulses</td>
<td>57</td>
</tr>
<tr>
<td>3.18</td>
<td>Pilot voltage vectors in one PWM cycle for magnetic pole identification:</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$V_{NS}$ and $-V_{NS}$</td>
<td>57</td>
</tr>
<tr>
<td>3.19</td>
<td>Simulation results of $\alpha$ axis current, $2\hat{\theta}<em>{r_HFI}$ and $\hat{\theta}</em>{r_HFI}$ for the synchronous machine without shorted coils</td>
<td>58</td>
</tr>
<tr>
<td>3.20</td>
<td>Simulation results of $\alpha$ axis current, $2\hat{\theta}<em>{r_HFI}$ and $\hat{\theta}</em>{r_HFI}$ for the synchronous machine with shorted coils</td>
<td>60</td>
</tr>
<tr>
<td>3.21</td>
<td>Low speed ($\pm 1Hz$) waveforms without load (top trace is $\hat{\theta}<em>{r_HFI}$, middle is $2\hat{\theta}</em>{r_HFI}$ and bottom trace is the phase A current (10A/div))</td>
<td>61</td>
</tr>
<tr>
<td>3.22</td>
<td>Low speed (2.25Hz) with fundamental current waveforms (top trace is $2\hat{\theta}<em>{r_HFI}$, middle is $\hat{\theta}</em>{r_HFI}$ and bottom is the phase A current (10A/div))</td>
<td>62</td>
</tr>
<tr>
<td>3.23</td>
<td>Magnetic pole identification (Channel A: applied voltage pulses; Channel D: the current in synchronous reference frame)</td>
<td>64</td>
</tr>
<tr>
<td>3.24</td>
<td>$\hat{\theta}<em>{r_HFI}$, $\hat{\theta}</em>{r_HFI_cps}$ and $\theta_{r_encoder}$</td>
<td>64</td>
</tr>
<tr>
<td>3.25</td>
<td>Compensated $\hat{\theta}_{r_HFI}$ and NS pole identification</td>
<td>65</td>
</tr>
<tr>
<td>3.26</td>
<td>Waveforms of voltage model rotor position estimation (Channel C: estimated rotor position at 80Hz; Channel D: phase current (20A/div))</td>
<td>70</td>
</tr>
<tr>
<td>3.27</td>
<td>Full speed range sensorless control with crossover function blocks</td>
<td>72</td>
</tr>
<tr>
<td>Section</td>
<td>Title</td>
<td>Page</td>
</tr>
<tr>
<td>---------</td>
<td>----------------------------------------------------------------------</td>
<td>------</td>
</tr>
<tr>
<td>4.1</td>
<td>The ISG overall structure</td>
<td>74</td>
</tr>
<tr>
<td>4.2</td>
<td>3-D drawing of the inverter</td>
<td>76</td>
</tr>
<tr>
<td>4.3</td>
<td>Prototype of the 10kW inverter</td>
<td>77</td>
</tr>
<tr>
<td>4.4</td>
<td>Induction machine ISG torque-speed characteristics</td>
<td>78</td>
</tr>
<tr>
<td>4.5</td>
<td>Prototype of induction machine in the 10kW IM ISG system</td>
<td>79</td>
</tr>
<tr>
<td>4.6</td>
<td>The relationship between $V_{\alpha 0}$ and $V_{\beta 0}$ and $V_{\alpha}$ and $V_{\beta}$</td>
<td>82</td>
</tr>
<tr>
<td>4.7</td>
<td>The relationship between $i_{\alpha l-1}$ and $i_{\beta l-1}$ and $i_{\alpha n}$ and $i_{\beta n}$</td>
<td>83</td>
</tr>
<tr>
<td>4.8</td>
<td>Phase voltage calculation</td>
<td>83</td>
</tr>
<tr>
<td>4.9</td>
<td>The basic Delta-connection system diagram</td>
<td>85</td>
</tr>
<tr>
<td>4.10</td>
<td>Block diagram for rotor position sensorless algorithm</td>
<td>87</td>
</tr>
<tr>
<td>4.11</td>
<td>ISG characteristics</td>
<td>88</td>
</tr>
<tr>
<td>4.12</td>
<td>The conventional sensorless control of induction machine in flux weakening region</td>
<td>90</td>
</tr>
<tr>
<td>4.13</td>
<td>$T_e$, $T_{load}$ and rotor speed</td>
<td>92</td>
</tr>
<tr>
<td>4.14</td>
<td>$I_d$ and $I_q$</td>
<td>93</td>
</tr>
<tr>
<td>4.15</td>
<td>$V_d$ and $V_q$</td>
<td>93</td>
</tr>
<tr>
<td>4.16</td>
<td>$\theta_{est}$</td>
<td>94</td>
</tr>
<tr>
<td>4.17</td>
<td>Phase currents $I_a, I_b, I_c$</td>
<td>94</td>
</tr>
<tr>
<td>4.18</td>
<td>The experimental results of $I_{DC}$ (CH3) and $\theta_{est}$ (CH4) with $-I_{q,ref}$ command</td>
<td>95</td>
</tr>
<tr>
<td>4.19</td>
<td>The system block diagram with commanding $I_{d,ref}$</td>
<td>96</td>
</tr>
</tbody>
</table>
4.20 $T_e$, $T_{load}$ and rotor speed ....................................................... 98

4.21 $I_{d\_ref}$, $I_{d\_fdb}$, $I_{q\_ref}$ and $I_{q\_fdb}$ ......................................................... 98

4.22 $I_{d\_ref}$ and $I_{d\_fdb}$ ................................................................. 99

4.23 $V_d$ and $V_q$ ................................................................. 99

4.24 Rotor angle ................................................................. 100

4.25 Phase currents $I_a$, $I_b$, $I_c$ ....................................................... 100

4.26 Indirect field orientation control through slip control in the flux weakening region .................................................. 101

4.27 $T_e$ ................................................................. 102

4.28 DC bus current ................................................................. 102

4.29 The estimated rotor flux angle ....................................................... 103

4.30 The DC bus voltage reference and feedback .................................................. 103

4.31 Phase currents $I_a$, $I_b$, $I_c$ ....................................................... 104

4.32 Torque and speed waveforms in generating mode .................................................. 105

4.33 Load testing waveforms for $V_{DC}$ (CH2), $I_{DC}$ (CH3: 100A/div) and $I_{phase}$
(CH4: 400A/div) ................................................................. 106

4.34 Load change from full to zero: $V_{DC}$ (CH2), $I_{DC}$ (CH3: 100A/div) and $I_{phase}$
(CH4: 400A/div) ................................................................. 106

4.35 Load change from zero to full: $V_{DC}$ (CH2), $I_{DC}$ (CH3: 100A/div) and $I_{phase}$
(CH4: 400A/div) ................................................................. 107

4.36 Control diagram for ISG ................................................................. 109

4.37 Induction machine ISG torque-speed characteristics [56] .................................................. 110

4.38 Induction machine ISG torque-speed characteristics .................................................. 111

xv
4.39 Induction machine ISG starting performance with Y-connection . . . . . 114

4.40 Induction machine ISG starting performance with Delta-connection . . . 115

4.41 Induction machine ISG generating performance with Delta-connection . . . 116

A.1 Second order high pass filter diagram . . . . . . . . . . . . . . . . . . . 121

A.2 Sixth order low pass filter diagram . . . . . . . . . . . . . . . . . . . . 122
CHAPTER 1

INTRODUCTION

1.1 Energy Conversion

Energy has played a key role in human life since the Industrial Revolution that began in 1860s. It is only stored in several forms naturally, such as fossil fuels, solar, tidal, geothermal, nuclear forms, and so on. But above energy cannot be used directly. All these nature energy should be converted to thermal or mechanical energy that can be used directly by customers in the Industrial Revolution. Thanks to Michael Faraday, Thomas Edison, Nikola Tesla, George Westinghouse and other scientists, DC and AC electrical generator were invented in the Second Industrial Revolution that can convert thermal and mechanical energy to electric energy through all kinds of turbines (prime movers). The main types of turbines are listed as follows:

(1) Steam turbines
(2) Gas turbines
(3) Hydraulic turbines
(4) Wind turbines
(5) Internal Combustion Engines (ICE)
The benefit of the conversion is that electric energy is easy to transmit for long distance and flexible to distribute for customer’s needs. Nowadays, more than 30% of energy is converted to electric energy before usage. Water potential, kinetic energy and wind energy are also converted to electric energy with electrical generators through prime movers in which mechanical energy is stored [1].

Most electric energy is produced through constant speed synchronous generators with constant voltage and frequency. Then it is transmitted and distributed to various consumers through power grids. Recently, with the higher standard requirements of low emission and low pollution, variable speed electric generators become a hot popular topic in research and industrial areas. In this dissertation, both constant and variable speed generator systems that operate in stand-alone and with power grid capacities will be introduced briefly.

Hydraulic potential energy is converted to mechanical potential energy in hydraulic energy turbines. Then it drives electric generators to produce electric energy. In general, the hydraulic turbines speed is very low, normally below 100 rpm.

Wind energy is converted to mechanical energy in wind turbines. It is a typical variable speed generation system with a wide speed range. It means that sophisticated and high performance power electronics and motor control will be needed. With the quickly growing up of wind energy application, it is estimated that wind energy will provide more than 10% of electric energy by 2020 in the world.

The steam and gas turbines and ICEs are widely and practically used in the industry with all fossil fuels. Recently, high speed gas turbines are available in the range from 70,000 rpm to 80,000 rpm.
1.2 Integrated Starter Generator

The concept of an Integrated Starter Generator (ISG) system is introduced into automotive [2, 3, 4, 5, 6, 7], aircrafts [8, 9, 10, 11, 12, 13, 14, 15] and other industrial applications [16] where space and weight reduction is of critical importance. In general, the electric machine in an ISG system will operate as a motor to start the engine with a very high torque and then act as a generator at higher speeds supplying constant power to the DC battery bank and electrical load. The existing commercial ISG systems are based on DC machine. The disadvantage of the DC machine is its brushes. With large starting torque, the DC current is very high and the brush contact resistance will also generate high power loss. In addition, sparks can damage brushes and cause interference within the whole system. Also, the maintenance of the brushes is expensive. With the maturing of AC machine control technology, in order to overcome the above problem, more and more researchers are working on using AC machine ISG to replace existing DC machine ISG. One of the striking benefits of AC machine ISG is that AC machine ISG can work with brushless and then dramatically reduce DC side current. As a result, the battery current stress is substantially relieved and battery life much extended. In addition, a DC/AC inverter powered ISG system can save the high costs for the expensive brushes and commutators of DC machines [17].

The torque-speed characteristic for AC machines in an ISG application is very special as compared to other drive systems, which imposes many system challenges [8, 18]. In other words, AC machine in an ISG system has to satisfy very special torque-speed characteristics for starting and generating respectively [19]. For example, an induction machine ISG system presented in this dissertation should first run as a starter and supply high starting torque (above 40Nm) at lower speed range from 0rpm to 5000rpm. Then, the AC machine
switches to generating operation from about three times of the base speed with a constant power output. The generating mode could run up to a very high speed \((12000\text{rpm})\), six times of the base speed. In addition, the ISG system should deliver \(10kW\) power to the battery bank and load at higher speeds from \(7200\text{rpm}\) to \(12000\text{rpm}\) [20].

Higher speeds also mean volume and weight reductions in aircraft applications. High speed, small size, light weight and reliability are key issues for electric generators on aircrafts. Power rating of starter generator (SG) is from tens of kilowatts to \(1\text{MW}\) for aircrafts. For the maintenance expense and safety issue of brushes of DC machines, it is an industry trend to replace brush DC SG system with brushless AC SG system. Another issue of the application of ISG in aircraft is that a speed or position sensor is undesirable in a high speed drive system because it has the following problems:

1. Cost and reliability;
2. Accuracy limited because of environmental factors;
3. Maintenance requirements;
4. Shaft extension and mounting arrangements.

In order to overcome these problems, the sensorless technology is discussed and implemented in the dissertation with two different electric machines: permanent magnet synchronous machine (PMSM) and induction machine (IM).

### 1.3 Sensorless Technology for AC Electric Machine

Obviously, in case of hostile environment or very high speed applications, speed sensors are either not available or too expensive. Recently, a lot of research work has focused on the elimination of the mechanical speed sensor at the machine shaft without reducing the dynamic performance of the whole system. The sensorless control of AC drive systems
has the attractions of low cost and high reliability. Nowadays, it is attractive and feasible to implement sensorless field orientation control algorithm for the wide speed range ISG system with the high speed high performance Digital Signal Processors (DSP) or Micro Controllers (MCU). Sensorless speed estimation methods proposed in literature can be classified as:

1. Synthesis from state equations;
2. Model Reference Adaptive System (MRAS);
3. Extended Kalman Filter (EKF);
4. Slip calculation;
5. Slot harmonics;
6. High frequency signal injection;
7. Sliding mode observer (SMO).

### 1.3.1 Review of Sensorless Control for Synchronous Machine

As is well known, most sensorless methods for PMSM fail at low and zero speeds because the rotor position estimation fundamentally relies on the Back-EMF or speed dependent flux variables. The rotor saliency related information can be extracted from phase voltages and currents with high frequency injection (HFI). The method can be in forms of open-loop and close-loop. The position accuracy is dependent on given machine design, that is rotor saliency. Rotor position estimation by HFI to track rotor saliency has been successful in applications needing zero speed starting (zero speed torque) for synchronous machines. Although significant research progress on the sensorless control scheme of PM synchronous machines has been reported in the literatures [21, 22, 23, 24, 25, 26, 27, 28], few publications so far have addressed the issues related to the effects of eddy current on the
rotor position estimation for sensorless control. For sensorless control of PM synchronous machines at zero speed, both the initial $d$ axis position and the polarity of the magnetic pole are to be identified. Although the HFI method for initial rotor position estimation can identify the axis of the magnets but not the magnetic polarity. The conventional method for the North-South (N-S) pole is determined based on the magnetic saturation. The commonly used scheme is to inject pilot voltages and then detect the corresponding currents to determine the pole polarity [29, 30, 31, 32, 33, 34].

This dissertation thoroughly investigates the eddy current effects and contributes to the improved modeling and implementation of rotor position estimation and sensorless control based on HFI [25]. It is shown that the rotor estimation results of a PM synchronous machine with and without considering the eddy current effects are significantly different. The improved HFI control scheme is developed with compensation for the eddy current effects.

In order to obtain right North and South pole information, a simple yet effective method is proposed to identify the N-S pole based on space vector PWM suitable for arbitrary initial rotor position at zero speed [32]. The principle of the magnetic pole identification is discussed in details and the proposed method introduced. The validity of the proposed identification method is verified by experiment results.

The HFI method can only be used in zero and low speed range. When the PMSM runs to a higher speed range, the HFI is not a proper sensorless scheme any more. In this dissertation, a new voltage model based sensorless algorithm is developed in high speed ranges. In addition, a practical transition method between HFI and voltage model sensorless algorithms will be discussed briefly. The experimental results will be given to validate the effectiveness of the proposed sensorless scheme.
1.3.2 Review of Sensorless Control for Induction Machine

Generally, there are three typical control methods for the induction machine drive system: (1) Scalar control; (2) Direct torque control; (3) Vector control.

1. Scalar control

Before the vector control theory was introduced, the volts-per-hertz control method was widely used for industrial applications. It can provide simple control structure and good steady state performance. Control variables are voltage and its frequency. The flux is provided with constant V/f ratio. The system is low cost and simple because there are no feedback devices required. However, their dynamic performances are poor when the system subject to a load change. So the volts-per-hertz control method is not suitable for the high performance drive system that requires fast and precise torque response.

2. Direct Torque Control

The motor’s torque and flux can be controlled directly with the direct torque control (DTC) method that can calculate the motor torque directly and without using modulation and feedback. Therefore, the torque response is dramatically fast.

3. Field Orientation Control

The field orientation control (FOC) theory was proposed to meet high performance drive requirements. There are three different FOC methods: (1) rotor field orientation; (2) stator field orientation; (3) air gap field orientation. In addition, there are direct and indirect implementation methods for each field orientation method [35].

The aim of the field orientation control is to decouple the induction machine from non-linear, high order coupled system to simulated decoupled DC machine. The three phase system (voltages, currents and flux linkages) can be decoupled into the two phase system ($dq$ coordination). The two phase variables are orthogonal in the coordination. Like DC
machine, induction machine can be controlled by torque component (aligned with $q$ axis) and flux linkage component (aligned with $d$ axis) independently with FOC. In addition, the drive system has current feedback loop. So, both steady state and dynamic characteristics are improved in the induction machine drive system [36, 37, 38, 39].

In this dissertation, only direct and indirect rotor flux orientation are discussed. Because rotor flux cannot be measured directly with flux sensors, for the direct field orientation control (DFOC), the rotor flux is estimated or observed with several different observers using the induction machine’s terminal quantities such as currents, voltages and DC bus voltage. A major problem of DFOC is the implementation of the integration algorithm in the DSP control system. For the indirect field orientation control (IFOC) the rotor flux can be estimated with slip frequency and rotor speed. However, the slip frequency calculation depends on rotor parameters. When the rotor parameters are changed according to its temperature, the actual slip frequency should be changed accordingly. Because the rotor parameters can not be measured directly, many researchers presented their rotor parameters estimation methods to overcome this problem. The comparison of the three drive systems is listed in Table 1.1 [40, 41, 42, 43].

FOC is applied to ensure that the torque and flux are controlled independently and thus, high performance obtained. One method of FOC is the flux observer based method, which is usually used in sensorless FOC induction machine drives. Many types of close-loop flux observers exist in literature [44, 45, 46, 47, 44, 45]. Close-loop observers with the feedback loop use the estimated stator current error or the estimated stator voltage error to push the estimated rotor flux converges to the actual flux. Many close-loop observers are supposed to have better accuracy with more design complexity imposed [48, 49, 50].
In this dissertation, a new control method for the sensorless control of induction machine in flux weakening region (generating mode) is presented. In the generating mode, AC phase voltages reach their limits because of the limited DC bus voltage. The maximum AC phase voltage is $\frac{2\pi}{\pi}V_{DC}$. In generating mode, only slip frequency can be controlled through DC bus voltage PI controller. The system synchronous speed is the sum of the slip speed and the estimated rotor speed. In short, the controller maintains the DC bus voltage constant through adjusting the slip speed.
<table>
<thead>
<tr>
<th>Drive System</th>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>V/f</td>
<td>low cost</td>
<td>Flux uncontrollable</td>
</tr>
<tr>
<td></td>
<td>No feedback device required</td>
<td>Poor dynamic torque response</td>
</tr>
<tr>
<td></td>
<td>Simple structure</td>
<td>low accuracy</td>
</tr>
<tr>
<td>FOC</td>
<td>Good torque response</td>
<td>Feedback needed</td>
</tr>
<tr>
<td></td>
<td>Accurate speed control</td>
<td>Costly</td>
</tr>
<tr>
<td></td>
<td>Full torque at zero speed</td>
<td>Modulator needed</td>
</tr>
<tr>
<td></td>
<td>Performance approaching DC drive</td>
<td>lower torque ripple</td>
</tr>
</tbody>
</table>

| DTC          | Fast torque response | High torque ripple |
|              | Torque controlled directly | Flux control poor at low load and speed |
|              | Full torque at zero speed | High current distortion and harmonics |
|              | No modulator needed | Motor status unknown |
|              | low parameter sensitivity | |

Table 1.1: Comparison of drive systems
1.4 DSP Based Electric Drive System

Thanks for the full digital control for AC drives during the past twenty years, most of engineers have been released from analog world, and devoted into digital control algorithm development and performance improvement. With the development of the DSP, more powerful motor control DSP chips are introduced into the industry applications. In the Power Electronics and Electric Machine Group (PEEM) at The Ohio State University, Texas Instrument’s (TI) TMS320 series DSPs (F240,F2407/A,F2812) have been applied to many motor drive projects. TMS320F281X DSP has been released since 2002 with optimized cost and performance and especially suitable for high precision motor control. The TMS320F2812 DSP products, members of the TI C28X DSP generation, are highly integrated, high performance solutions for demanding control applications.

The highly integrated 32-bit flash based F2812 DSP is revolutionary in many ways:

(1) The 32-bit TMS320C28x DSP core is the industry’s most powerful fixed-point architecture capable of 150 MIPS of performance and a single-cycle 32X32-bit MAC.

(2) These new devices combine together with the C28x DSP core, a large amount of fast access flash memory, high precision analog peripherals, and many other control and communication modules making these devices the industry’s most integrated DSPs.

(3) The C28x DSP core is the most C/C++ efficient of the industry with a C-to-Assembly ratio of 1.1 and a unique "virtual floating-point" programming capability.

A detailed comparison between TMS320F2812 and TMS320F2407 is shown in Table 1.2. It is shown taht TMS320F2812 improves its performance dramatically. Advanced control algorithms can be easily implemented with this DSP. In addition, the higher C-to-Assembly ratio will help the developers focus more on control algorithm design rather than the hard-to-understand assembly language.
Table 1.2: Feature comparison between TMS320F2812 and TMS320F2407

<table>
<thead>
<tr>
<th></th>
<th>TMS320F2812</th>
<th>TMS320F2407A</th>
</tr>
</thead>
<tbody>
<tr>
<td>Instruction Cycle</td>
<td>6.67nS (150MHz)</td>
<td>25nS (40MHz)</td>
</tr>
<tr>
<td>On-chip Flash</td>
<td>128K</td>
<td>32K</td>
</tr>
<tr>
<td>SARAM</td>
<td>18K</td>
<td>2K</td>
</tr>
<tr>
<td>Digital I/O</td>
<td>56</td>
<td>41</td>
</tr>
<tr>
<td>ADC</td>
<td>12 bit</td>
<td>10 bit</td>
</tr>
<tr>
<td>ADC conversion time</td>
<td>200ns</td>
<td>500ns</td>
</tr>
</tbody>
</table>

1.5 Contributions of the Dissertation

In the dissertation, two types of ISG systems are discussed. The specific contributions for the synchronous machine based ISG system are:

(1) Proposed and developed a new high frequency injection based initial rotor position estimation method for PMSM at zero and low speed. It is a new idea that considers eddy current effect compensation for the high frequency injection method.

(2) Proposed and developed an improved magnetic pole identification algorithm for PMSM at zero speed. The algorithm can identify N-S pole with arbitrary rotor position without pre-alignments, which is different from other existed methods.

(3) Proposed and developed a new rotor position observer for synchronous machine sensorless control in high speed range. The observer can decouple the model of the salient
pole synchronous machine. The rotor position can be easily estimated with the proposed observer.

For an induction machine based ISG system, the specific contributions are listed as follows:

(1) Developed a rotor position observer for induction machine sensorless control in wide speed range. The observer works well from zero speed to high speed (0 rpm to 12000 rpm).

(2) Proposed and developed a field orientation control algorithm for starting mode in a Delta-connection induction machine ISG system. In the starting mode, the system can meet the special torque-speed characteristics required by an ISG system.

(3) Proposed and developed a new sensorless control structure for generating mode. Different from the drive system structure with speed sensor, the proposed slip control structure is suitable for sensorless control in high speed generator system in deep flux weakening region.

1.6  Dissertation Organization

There are five chapters in this dissertation:

In Chapter 1, the history of energy conversion and the concept of integrated starter generator are introduced briefly. Sensorless control technology of the synchronous machine and induction machine are reviewed. The comparison of the two drive systems are discussed. In addition, DSP based electric drive system is introduced. The objectives and organization of the dissertation are presented at the end of this chapter.

The fundamental reference frame theory is an important concept for electric machine drive system. It is introduced briefly in Chapter 2. The dynamic models of synchronous
machines and induction machines are presented in this chapter. The comparison between permanent magnetic synchronous machine and induction machine is discussed in details in Chapter 2.

The principles related to the sensorless control algorithm of PMSM are discussed in Chapter 3. It includes: (1) initial rotor position estimation with high frequency injection technology; (2) rotor polarity identification with magnetic saturation theory; (3) voltage model based rotor position estimation at high speed. Simulation and experimental results are presented at the end of Chapter 3.

Sensorless control algorithm of the induction machine over a wide speed range is investigated in Chapter 4. The system hardware design is introduced briefly. The transformation between Y-connection and Delta-connection is derived. Different control methods are designed for starting mode and generating mode respectively. The simulation and experimental results verify that the designed induction machine ISG system meets the required torque-speed characteristics.

Finally, in Chapter 5, the contributions of this dissertation are summarized and future work is also discussed.
CHAPTER 2

DYNAMIC MODEL OF ELECTRIC MACHINES

2.1 Reference Frame Theory

As is well known, the reference frame theory plays an important role to the analysis of different electric machines. All analysis presented in this dissertation are based on the reference frame theory.

In the late 1920s, Park proposed a new approach for electric machine analysis. He formulated a change of variables which replaced the variables (voltages, currents and flux linkages) associated with the stator windings of a synchronous machine with variables associated with fiction windings rotating with the rotor. In other words, he transformed, or referred, the stator variables to a frame of reference in the rotor. Park’s transformation, which revolutionized electric machine analysis, has the unique property of eliminating all time-varying inductance from the voltage equations of the synchronous machine which occur due to: (1) electric circuits in relative motion and (2) electric circuits with varying magnetic reluctance.

Later, in the late 1930s, H. C. Stanley showed that the time-varying inductances in the voltage equations of an induction machine due to electric circuits in relative motion could be eliminated by transforming the variables associated with the rotor windings (rotor
variables) to variables associated with fictitious stationary windings. In this case the rotor variables are transformed to a frame reference fixed in the stator.

G. Kron introduced a change of variables that eliminated the position of time-varying mutual inductances of a symmetrical induction machine by transforming both the stator variables and the rotor variables to a reference frame rotating in synchronism with the rotating magnetic field. This reference frame is commonly referred to as the synchronously rotating reference frame.

D. S. Brereton employed a change of variables that also eliminated the time-varying inductances of a symmetrical induction machine by transforming the stator variables to a reference frame fixed in the rotor. This is essentially Park’s transformation applied to induction machines.

Park, Stanley, Kron and Brereton et al. developed changes of variables, each of which appeared to be uniquely suited for a particular application. Consequently, each transformation was derived and treated separately in literature. In fact, it was shown in 1965 by Krause and Thomas that time-varying inductances can be eliminated by referring the stator and rotor variables to a common reference frame which may rotate at any speed, this so-called arbitrary reference frame.

Later, it was noted that the stator variables of a synchronous machine could also be referred to the arbitrary reference frame. However, we will find that the time-varying inductances of a synchronous machine are eliminated only if the reference frame is fixed in the rotor (Park); consequently the arbitrary reference frame does not offer the advantages in the analysis of the synchronous machines that it does in the case of induction machines [51, 52].
2.1.1 Equations of Transformation: Change of Variables

Although changes of variables are used in the analysis of AC machine to eliminate time-varying inductances, changes of variables are also employed in the analysis of various static, constant parameter power system components and control systems associated with electric drives. Fortunately, all known real transformations for these components and controls are also contained in the transformation to the arbitrary reference frame, the same general transformation used for the stator variables of the induction and synchronous machines and for the rotor variables of induction machines.

A change of variables that formulates a transformation of the three phase variables of stationary circuit elements to the arbitrary reference frame may be expressed as

\[ f_{qd0s} = K_s f_{abcs} \]  \hspace{1cm} (2.1)

where

\[ (f_{qd0s})^T = [f_{qs} \ f_{ds} \ f_{0s}] \]  \hspace{1cm} (2.2)

\[ (f_{abcs})^T = [f_{as} \ f_{bs} \ f_{cs}] \]  \hspace{1cm} (2.3)

\[
K_s = \frac{2}{3} \begin{bmatrix}
\cos(\theta) & \cos(\theta - \frac{2}{3}\pi) & \cos(\theta + \frac{2}{3}\pi) \\
\sin(\theta) & \sin(\theta - \frac{2}{3}\pi) & \sin(\theta + \frac{2}{3}\pi) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}
\]  \hspace{1cm} (2.4)

It can be shown that for the inverse transformation we have

\[
(K_s)^{-1} = \begin{bmatrix}
\cos(\theta) & \sin(\theta) & 1 \\
\cos(\theta - \frac{2}{3}\pi) & \sin(\theta - \frac{2}{3}\pi) & 1 \\
\cos(\theta + \frac{2}{3}\pi) & \sin(\theta + \frac{2}{3}\pi) & 1
\end{bmatrix}
\]  \hspace{1cm} (2.5)
In the above equations, $f$ can represent either voltage, current, flux linkage, or electric change. The superscript $T$ denotes the transpose of a matrix. The $s$ subscript indicates the variables, parameters, and transformation associated with stationary circuits.

The total instantaneous power may be expressed in $abc$ variables as

$$P_{abc} = v_{as}i_{as} + v_{bs}i_{bs} + v_{cs}i_{cs} \quad (2.6)$$

The total power expressed in the $qd0$ variables must equal total power expressed in the $abc$ variables; hence yields

$$P_{qd0s} = P_{abc} = \frac{3}{2}(v_{qs}i_{qs} + v_{ds}i_{ds} + 2v_{os}i_{os}) \quad (2.7)$$

The $3/2$ factor comes about due to the choice of the constant used in the transformation. Although the waveforms of the $qs$ and $ds$ voltages, currents, flux linkages, and electric charges are dependent upon the angular velocity of the frame of reference, the waveform of total power is independent of the frame of reference. In other words, the waveform of the total power is the same of the reference frame in which it is evaluated [51].

### 2.1.2 Stationary Circuit Variables Transformed to the Arbitrary Reference Frame

It is convenient to treat resistive, inductive, and capacitive circuit elements separately [51].

#### Resistive Elements

For three phase resistive circuit, we have

$$v_{abc} = r_s i_{abc} \quad (2.8)$$

So, we obtain

$$v_{dq0s} = K_s r_s (K_s)^{-1} i_{dq0s} \quad (2.9)$$
All stator phase windings of either a synchronous or a symmetrical induction machine are designed to have the same resistance. So, if the nonzero elements of the diagonal matrix \( r_s \) are equal, then
\[
K_s r_s (K_s)^{-1} = r_s
\]  
(2.10)

**Inductive Elements**

For three phase resistive circuit, we have
\[
v_{abcs} = p \lambda_{abcs}
\]  
(2.11)

where \( p \) is the operator \( d/dt \).

\[
v_{dq0s} = K_s p [(K_s)^{-1} \lambda_{dq0s}]
\]  
(2.12)

which can be written as
\[
v_{dq0s} = K_s p [(K_s)^{-1} \lambda_{dq0s}] + K_s (K_s)^{-1} p \lambda_{dq0s}
\]  
(2.13)

where
\[
K_s p [(K_s)^{-1}] = \omega \begin{bmatrix} 0 & 1 & 0 \\ -1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}
\]  
(2.14)

Then
\[
v_{dq0s} = \omega \lambda_{qds} + p \lambda_{dq0s}
\]  
(2.15)

where
\[
(\lambda_{qds})^T = [-\lambda_{qs} \ \lambda_{ds} \ 0]
\]  
(2.16)

For a linear magnetic system, the flux linkages may be expressed as
\[
\lambda_{abcs} = L_s i_{abcs}
\]  
(2.17)
whereupon the flux linkages in the arbitrary reference frame may be written as

\[ \lambda_{dq0s} = K_s L_s (K_s)^{-1} i_{dq0s} \]  

(2.18)

An inductance matrix that is common to synchronous and induction machines is of the form

\[
L_s = \begin{bmatrix}
  L_{ls} + L_{ms} & -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} \\
  -\frac{1}{2}L_{ms} & L_{ls} + L_{ms} & -\frac{1}{2}L_{ms} \\
  -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} & L_{ls} + L_{ms}
\end{bmatrix}
\]

(2.19)

where \( L_{ls} \) is a leakage inductance and \( L_{ms} \) is a magnetizing inductance. So,

\[
K_s L_s (K_s)^{-1} = \begin{bmatrix}
  L_{ls} + \frac{3}{2}L_{ms} & 0 & 0 \\
  0 & L_{ls} + \frac{3}{2}L_{ms} & 0 \\
  0 & 0 & L_{ls}
\end{bmatrix}
\]

(2.20)

2.1.3 Transformations in Electric Machines

(1) Clarke and Inverse Clarke Transformation [52, 53]

The Clarke transformation is used to project the three phase \( abc \) coordinates to the two phase \( \alpha \beta \) coordinates. The transformation is shown in Figure 2.1. The \( \alpha \) axis coincides with the phase \( a \) axis and \( \beta \) axis leads the \( \alpha \) axis by 90°.

\[
f'_{\alpha} = f_{an} - f_{bn} \sin(30^\circ) - f_{cn} \sin(30^\circ)
\]

(2.21)

\[
f'_{\beta} = f_{bn} \cos(30^\circ) - f_{cn} \cos(30^\circ)
\]

(2.22)

In the three phase system, \( f_{an} + f_{bn} + f_{cn} = 0 \). So \( f_{cn} = -f_{an} - f_{bn} \). By substituting \( f_{cn} \) into above equations, we obtain:

\[
f'_{\alpha} = \frac{3}{2} f_{an}
\]

(2.23)
In order to implement the transformation in DSP system with per unit system [53], the original Clarke transformation (Equation 2.21 – Equation 2.24) can be rescaled as follows.

Let \( f_\alpha = \frac{2}{3} f'_\alpha \), then

\[
f_\alpha = f_{an} \tag{2.25}
\]

\[
f_\beta = \frac{1}{\sqrt{3}} (f_{an} + 2f_{bn}) \tag{2.26}
\]

The matrix formation of the Clarke transformation is

\[
\begin{bmatrix}
    f_{an} \\
    f_{\beta n}
\end{bmatrix}
= \begin{bmatrix}
    1 & 0 \\
    \frac{1}{\sqrt{3}} & \frac{2}{\sqrt{3}}
\end{bmatrix}
\begin{bmatrix}
    f_{an} \\
    f_{bn}
\end{bmatrix} \tag{2.27}
\]
Where, \( f \) presents voltage, current and flux linkage. It means that voltage, current and flux linkage can be transformed between three phase \( abc \) coordinates and two phase \( \alpha \beta \) coordinates with Clarke and inverse Clarke transformation.

Then the inverse Clarke transformation is

\[
\begin{bmatrix}
  f_{an} \\
  f_{bn} \\
  f_{cn}
\end{bmatrix} =
\begin{bmatrix}
  1 & 0 \\
  -\frac{1}{2} & \frac{\sqrt{3}}{2} \\
  -\frac{1}{2} & -\frac{\sqrt{3}}{2}
\end{bmatrix}
\begin{bmatrix}
  f_{\alpha n} \\
  f_{\beta n}
\end{bmatrix}
\]  

(2.28)

(2) Park and Inverse Park Transformations [52, 53]

The Park and inverse Park transformation convert the quantities between the stationary reference frame and the synchronous reference frame. In Figure 2.2, project \( dq \) coordinates to \( \alpha \beta \) coordinates. Note that in the so-called \( dq \) system, the \( q \) axis leads the \( d \) axis by 90\(^\circ\).

![Diagram](image)

Figure 2.2: Project \( dq \) to \( \alpha \beta \) coordinates

\[
f_\alpha = f_d \cos(\theta) - f_q \sin(\theta)
\]  

(2.29)
\[ f_\beta = f_d \sin(\theta) + f_q \cos(\theta) \]  

(2.30)

So the matrix formation of the inverse Park transformation is:

\[
\begin{bmatrix}
  f_\alpha \\
  f_\beta 
\end{bmatrix} = \begin{bmatrix}
  \cos(\theta) & -\sin(\theta) \\
  \sin(\theta) & \cos(\theta) 
\end{bmatrix}
\begin{bmatrix}
  f_d \\
  f_q
\end{bmatrix}
\]  

(2.31)

Then the matrix formation of the Park transformation is:

\[
\begin{bmatrix}
  f_d \\
  f_q
\end{bmatrix} = \begin{bmatrix}
  \cos(\theta) & \sin(\theta) \\
  -\sin(\theta) & \cos(\theta)
\end{bmatrix}
\begin{bmatrix}
  f_\alpha \\
  f_\beta 
\end{bmatrix}
\]  

(2.32)

Some people define the \( q \) axis lags the \( d \) axis by 90°, which is the so called \( qd \) system.

The relationship of the rotating angles between the \( dq \) system and \( qd \) system is [54]:

\[ \theta_{qd} = \theta_{dq} + \frac{\pi}{2}. \]  

(2.33)

Substituting the relationship into above transformations with the trigonometric reduction equations 2.34 and 2.35. The results are basically the same except the order of the \( d \) and \( q \) variables.

\[ \cos \theta_{dq} = \cos(\theta_{qd} - \frac{\pi}{2}) = \sin \theta_{qd} \]  

(2.34)

\[ \sin \theta_{dq} = \sin(\theta_{qd} - \frac{\pi}{2}) = -\cos \theta_{qd} \]  

(2.35)

### 2.2 Dynamic Model of Synchronous Machines

In general, synchronous machines can be cataloged into two types according to air gaps between stator and rotor: (1) synchronous machine with uniform air gaps and concentric cylindrical rotor, such as surface mount permanent machine; (2) synchronous machine with nonuniform air gaps, such as salient pole synchronous machine, interior permanent machine (IPM), etc. The research presented in this dissertation will focus on the salient pole synchronous machine.
The model of a basic two pole salient pole synchronous machine with damper windings is shown in Figure 2.3. The \( d \) axis is aligned with the N-pole of the rotor and \( q \) axis is 90 degree apart from \( d \) axis.

![Figure 2.3: Basic representation of a salient pole synchronous machine with damper windings](image)

The voltage equations for the salient pole synchronous machine can be expressed as:

\[
\begin{align*}
v_{abc} &= r_s i_{abc} + \frac{d}{dt} \lambda_{abc} \\
v_{dq} &= r_i i_{dq} + \frac{d}{dt} \lambda_{dq}
\end{align*}
\]  

(2.36)

\[
\begin{align*}
\text{Where } v_{abc} &= \begin{bmatrix} v_{as} \\ v_{bs} \\ v_{cs} \end{bmatrix}; v_{dq} = \begin{bmatrix} v_{dr} \\ v_{qr} \\ v_{fd} \end{bmatrix}; r_s = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix} \text{ and } r_r = \begin{bmatrix} R_r & 0 & 0 \\ 0 & R_r & 0 \\ 0 & 0 & R_r \end{bmatrix}.
\end{align*}
\]

The subscripts \( s \) and \( r \) denote variables associated with stator and rotor windings, respectively.
The flux linkage equations can be expressed as:

\[
\begin{bmatrix}
\lambda_{abc}s \\
\lambda_{dqr}
\end{bmatrix} =
\begin{bmatrix}
L_s & L_{sr} \\
L^T_{sr} & L_r
\end{bmatrix}
\begin{bmatrix}
i_{abc}s \\
i_{dqr}
\end{bmatrix}
\]  

(2.38)

Where

\[
L_s =
\begin{bmatrix}
L_{ds} + L_{0s} + L_{2s} \cos(2\theta) & -\frac{1}{2}L_{0s} + L_{2s} \cos[2(\theta - \frac{\pi}{3})] & -\frac{1}{2}L_{0s} + L_{2s} \cos[2(\theta + \frac{\pi}{3})] \\
-\frac{1}{2}L_{0s} + L_{2s} \cos[2(\theta - \frac{\pi}{3})] & L_{ds} + L_{0s} + L_{2s} \cos[2(\theta - \frac{\pi}{3})] & -\frac{1}{2}L_{0s} + L_{2s} \cos[2(\theta + \pi)] \\
-\frac{1}{2}L_{0s} + L_{2s} \cos[2(\theta + \frac{\pi}{3})] & -\frac{1}{2}L_{0s} + L_{2s} \cos[2(\theta + \pi)] & L_{ds} + L_{0s} + L_{2s} \cos[2(\theta + \frac{2\pi}{3})]
\end{bmatrix}
\]  

(2.39)

\[
L_{sr} =
\begin{bmatrix}
L_{skd} \sin \theta & L_{sf d} \sin \theta & L_{skq} \cos \theta \\
L_{skd} \cos(\theta - \frac{2\pi}{3}) & L_{sf d} \cos(\theta - \frac{2\pi}{3}) & L_{skq} \cos(\theta - \frac{2\pi}{3}) \\
L_{skd} \cos(\theta + \frac{2\pi}{3}) & L_{sf d} \cos(\theta + \frac{2\pi}{3}) & L_{skq} \cos(\theta + \frac{2\pi}{3})
\end{bmatrix}
\]  

(2.40)

\[
L_r =
\begin{bmatrix}
L_{ld k} + L_{mk d} & L_{fd k d} & 0 \\
L_{fd k d} & L_{ld f} + L_{mf d} & 0 \\
0 & 0 & L_{l k q} + L_{mk q}
\end{bmatrix}
\]  

(2.41)

\(L_{ds}\) is the additional component due to the armature leakage flux; \(L_{0s}\) is self inductance due to the space fundamental air gap flux; \(L_{2s}\) is the component due to rotor position dependent inductance. The subscripts \(skq, skd, sf d\) denote mutual inductance between stator and rotor windings.

The above voltage equations could be used to describe the synchronous machine dynamics. However, it is still inconvenient because of the time-varying inductances. As aforementioned, change of variables between different reference frames was then introduced and commonly used for analysis. The transformations could be simply assigning the
rotating speed of the reference frame, the so-called arbitrary reference frame. The rotating speed of the reference frame could be zero, which is the so-called stationary reference frame.

Figure 2.4 shows the reference frames including the stationary reference frame circuit and arbitrary reference frame. The transformations between the stationary reference frame and arbitrary reference frame are:

\[
\begin{align*}
\mathbf{f}_{dq0} &= T_{abc \rightarrow dq0}(\theta) \mathbf{f}_{abcs} \\
T_{abc \rightarrow dq0}(\theta) &= \frac{2}{3} \begin{bmatrix}
\cos \theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta - \frac{4\pi}{3}) \\
-\sin \theta & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta - \frac{4\pi}{3}) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix} \\
T_{abc \rightarrow \alpha\beta0} &= T(0) = \begin{bmatrix}
1 & -\frac{1}{2} & -\frac{1}{2} \\
0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\
1 & 1 & 1
\end{bmatrix} \\
T_{abc \rightarrow \alpha\beta0}^{-1} &= T^{-1}(0) = \begin{bmatrix}
1 & 0 & 1 \\
-\frac{1}{2} & \frac{\sqrt{3}}{2} & 1 \\
-\frac{1}{2} & -\frac{\sqrt{3}}{2} & 1
\end{bmatrix}
\end{align*}
\]

If the transformations are between the \(abc\) reference frame and stationary reference frame, simply let \(\omega = 0\):
Figure 2.4: Reference frames
In addition, the transformations between stationary reference frame and arbitrary reference frame are:

$$T_{\alpha\beta_0 \rightarrow dq_0}(\theta) = \begin{bmatrix} \cos \theta & \sin \theta & 0 \\ -\sin \theta & \cos \theta & 0 \\ 0 & 0 & 1 \end{bmatrix}$$ (2.47)

and

$$T_{dq_0 \rightarrow \alpha\beta_0}(\theta) = \begin{bmatrix} \cos \theta & -\sin \theta & 0 \\ \sin \theta & \cos \theta & 0 \\ 0 & 0 & 1 \end{bmatrix}$$ (2.48)

According to Equation 2.36, apply the transformation from $abc$ frame to arbitrary reference frame:

$$T_{abc \rightarrow dq_0}(\theta) v_{abcs} = T_{abc \rightarrow dq_0}(\theta) r_s T_{abc \rightarrow dq_0}(\theta)^{-1} i_{dq0s} + T_{abc \rightarrow dq_0}(\theta) \frac{d}{dt} [T_{abc \rightarrow dq_0}(\theta) \lambda_{dq0s}]$$ (2.49)

where

$$T_{abc \rightarrow dq_0}(\theta) r_s T_{abc \rightarrow dq_0}(\theta)^{-1} i_{dq0s} = r_s$$

and

$$T_{abc \rightarrow dq_0}(\theta) \frac{d}{dt} [T_{abc \rightarrow dq_0}(\theta) \lambda_{dq0s}] = T_{abc \rightarrow dq_0}(\theta) \frac{d}{dt} [T_{abc \rightarrow dq_0}(\theta) \lambda_{dq0s}] + T_{abc \rightarrow dq_0}(\theta) T_{abc \rightarrow dq_0}(\theta) \frac{d}{dt} \lambda_{dq0s}$$

Because

$$T_{abc \rightarrow dq_0}(\theta) \frac{d}{dt} [T_{abc \rightarrow dq_0}(\theta)] = \omega \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}$$ (2.50)

Therefore, the voltage equations in arbitrary reference frame could be expressed as:

$$\begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix} = R_s \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + \omega \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \begin{bmatrix} \lambda_{ds} \\ \lambda_{qs} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \lambda_{ds} \\ \lambda_{qs} \end{bmatrix}$$ (2.51)
The time varying inductances are eliminated from the voltage equations. Therefore, for synchronous machine voltage equations, time varying inductances will not appear only if the reference frame is fixed in the rotor.

Figure 2.5: Salient pole synchronous machine equivalent circuit in $dq$ reference frame
Figure 2.5 shows the salient pole synchronous machine equivalent circuit in rotor reference frame. When transferred from the \(abc\) reference frame to the \(dq\) frame, the \(d\) axis and \(q\) axis inductances are:

\[
L_{dm} = \frac{3}{2}(L_{0s} + L_{2s}) \quad (2.52)
\]

\[
L_{qm} = \frac{3}{2}(L_{0s} - L_{2s}) \quad (2.53)
\]

\[
L_{ds} = L_{ls} + L_{dm} \quad (2.54)
\]

\[
L_{qs} = L_{ls} + L_{qm} \quad (2.55)
\]

The stator flux linkages in \(dq\) reference frame could be expressed as

\[
\frac{d}{dt} \begin{bmatrix} \lambda_{ds} \\ \lambda_{qs} \end{bmatrix} = \begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix} - R_s \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} - \omega \begin{bmatrix} 0 & -1 \\ -1 & 0 \end{bmatrix} \begin{bmatrix} \lambda_{ds} \\ \lambda_{qs} \end{bmatrix} \quad (2.56)
\]

That is

\[
\frac{d}{dt} \lambda_{ds} = v_{ds} - R_s i_{ds} + \omega \lambda_{qs} \quad (2.57)
\]

\[
\frac{d}{dt} \lambda_{qs} = v_{qs} - R_s i_{qs} - \omega \lambda_{ds} \quad (2.58)
\]

The rotor flux linkages in \(dq\) reference frame could be expressed as:

\[
\frac{d}{dt} \begin{bmatrix} \lambda_{dr} \\ \lambda_{qr} \\ \lambda_{fr} \end{bmatrix} = \begin{bmatrix} v_{dr} \\ v_{qr} \\ v_{fr} \end{bmatrix} - \begin{bmatrix} R_{dr} i_{dr} \\ R_{qr} i_{qr} \\ R_{rd} i_{fr} \end{bmatrix} \quad (2.59)
\]

The torque equation in rotor reference frame is:

\[
T_e = \frac{3}{2}P(\lambda_{ds}i_{qs} - \lambda_{qs}i_{ds}) \quad (2.60)
\]

The rotor speed is obtained by:

\[
\frac{J}{P} \frac{d}{dt} \omega_r = T_e - T_l \quad (2.61)
\]

Where \(P\) is the number of pole pairs and \(J\) is the system inertia.
2.3 Dynamic Model of Induction Machines

The induction machine is used in a wide variety of application as a mean of converting electric power to mechanical work. In this section, the dynamic models of the induction machine are presented with three different reference frames.

2.3.1 Dynamic Model in Synchronous Reference Frame

The following equations are referred to the synchronous rotating reference frame:

\[ v_{\alpha s}^e = r_s i_{\alpha s}^e + \frac{d\lambda_{\alpha s}^e}{dt} - \omega_e \lambda_{\beta s}^e \]  
(2.62)

\[ v_{\beta s}^e = r_s i_{\beta s}^e + \frac{d\lambda_{\beta s}^e}{dt} + \omega_e \lambda_{\alpha s}^e \]  
(2.63)

\[ \frac{d\lambda_{\alpha s}^e}{dt} = \frac{1}{2} \left( v_{\alpha s} - k_2 \lambda_{\alpha s}^e + k_3 \lambda_{\alpha r}^e + \omega_e \lambda_{\beta s}^e \right) \]  
(2.64)

\[ \frac{d\lambda_{\beta s}^e}{dt} = \frac{1}{2} \left( v_{\beta s} - k_2 \lambda_{\beta s}^e + k_3 \lambda_{\beta r}^e + \omega_e \lambda_{\alpha s}^e \right) \]  
(2.65)

where

\[ \lambda_{\alpha s}^e = i_{\alpha s}^e L_{is} + L_m (i_{\alpha s}^e + i_{\alpha r}^e) \]  
(2.66)

\[ \lambda_{\beta s}^e = i_{\beta s}^e L_{is} + L_m (i_{\beta s}^e + i_{\beta r}^e) \]  
(2.67)

\[ \lambda_{\alpha r}^e = i_{\alpha r}^e L_{ir} + L_m (i_{\alpha s}^e + i_{\alpha r}^e) \]  
(2.68)

\[ \lambda_{\beta r}^e = i_{\beta r}^e L_{ir} + L_m (i_{\beta s}^e + i_{\beta r}^e) \]  
(2.69)

also

\[ T_e = L_m \left( i_{\alpha r}^e i_{\beta s}^e - i_{\alpha s}^e i_{\beta r}^e \right) \]  
(2.70)

The rotor flux linkages in \( dq \) reference frame could be expressed as:

\[ \frac{d\lambda_{\alpha s}^e}{dt} = v_{\alpha s}^e - k_2 \lambda_{\alpha s}^e + k_3 \lambda_{\alpha r}^e + \omega_e \lambda_{\beta s}^e \]  
(2.71)
\[
\begin{align*}
\frac{d\lambda_{\beta s}^e}{dt} &= v_{\beta s}^e - k_2 \lambda_{\beta s}^e + k_3 \lambda_{\beta r}^e - \omega_e \lambda_{\alpha s}^e \quad (2.72) \\
\frac{d\lambda_{\alpha r}^e}{dt} &= v_{\alpha r}^e - k_4 \lambda_{\alpha r}^e + k_5 \lambda_{\alpha s}^e + (\omega_e - \omega_r) \lambda_{\alpha r}^e \quad (2.73) \\
\frac{d\lambda_{\beta r}^e}{dt} &= v_{\beta r}^e - k_4 \lambda_{\beta r}^e + k_5 \lambda_{\beta s}^e - (\omega_e - \omega_r) \lambda_{\alpha r}^e \quad (2.74)
\end{align*}
\]

where

\[
\begin{align*}
k_1 &= L_{ls} L_{lr} + L_m L_{ls} + L_m L_{lr}' \\ k_2 &= \frac{r_s}{k_1} (L_{lr} + L_m) \\ k_3 &= \frac{r_r}{k_1} L_m \\ k_4 &= \frac{r_r'}{k_1} (L_{ls} + L_m) \\ k_5 &= \frac{r_r'}{k_1} L_m
\end{align*}
\]

The currents of the stator and rotor are linear relationship with the flux linkages. The equations are shown as follows:

\[
\begin{align*}
i_{\alpha s}^e &= \frac{L_{lr}'}{k_1} \lambda_{\alpha s}^e + \frac{L_m}{k_1} \lambda_{\alpha r}^e \\ i_{\beta s}^e &= \frac{L_{lr}'}{k_1} \lambda_{\beta s}^e - \frac{L_m}{k_1} \lambda_{\beta r}^e \\ i_{\alpha r}^e &= \frac{L_{ls} + L_m}{k_1} \lambda_{\alpha r}^e - \frac{L_m}{k_1} \lambda_{\alpha s}^e \\ i_{\beta r}^e &= \frac{L_{ls} + L_m}{k_1} \lambda_{\beta r}^e - \frac{L_m}{k_1} \lambda_{\beta s}^e
\end{align*}
\]

2.3.2 Dynamic Model in Stationary Reference Frame

The following equations are referred to the stationary rotating reference frame:

\[
v_{\alpha s}^s = r_s i_{\alpha s}^s + \frac{d\lambda_{\alpha s}^s}{dt} \quad (2.84)
\]
\[ v_{s}^{\beta} = r_{s}^{s} i_{s}^{\beta} + \frac{d\lambda_{s}^{\beta}}{dt} \]  \hspace{1cm} (2.85)

\[ v_{s}^{\alpha} = r_{r}^{s} i_{r}^{s} + \frac{d\lambda_{r}^{s}}{dt} + \omega_{r} \lambda_{r}^{s} \]  \hspace{1cm} (2.86)

\[ v_{r}^{s} = r_{r}^{r} i_{r}^{s} + \frac{d\lambda_{r}^{s}}{dt} - \omega_{r} \lambda_{r}^{s} \]  \hspace{1cm} (2.87)

where

\[ \lambda_{s}^{\alpha} = i_{s}^{\alpha} L_{ls} + L_{m}(i_{s}^{\alpha} + i_{s}^{r}) \]  \hspace{1cm} (2.88)

\[ \lambda_{s}^{\beta} = i_{s}^{\beta} L_{ls} + L_{m}(i_{s}^{\beta} + i_{s}^{r}) \]  \hspace{1cm} (2.89)

\[ \lambda_{r}^{s} = i_{r}^{s} L_{lr} + L_{m}(i_{r}^{s} + i_{r}^{s}) \]  \hspace{1cm} (2.90)

\[ \lambda_{r}^{s} = i_{r}^{s} L_{lr} + L_{m}(i_{r}^{s} + i_{r}^{s}) \]  \hspace{1cm} (2.91)

also

\[ T_{e} = L_{m}(i_{s}^{s} i_{r}^{s} - i_{s}^{s} i_{r}^{s}) \]  \hspace{1cm} (2.92)

The rotor flux linkages in stationary reference frame could be expressed as:

\[ \frac{d\lambda_{s}^{\alpha}}{dt} = v_{s}^{\alpha} - k_{2} \lambda_{s}^{\alpha} + k_{3} \lambda_{r}^{s} \]  \hspace{1cm} (2.93)

\[ \frac{d\lambda_{s}^{\beta}}{dt} = v_{s}^{\beta} - k_{2} \lambda_{s}^{\beta} + k_{3} \lambda_{r}^{s} \]  \hspace{1cm} (2.94)

\[ \frac{d\lambda_{r}^{s}}{dt} = v_{r}^{s} - k_{4} \lambda_{s}^{s} + k_{5} \lambda_{s}^{s} - \omega_{r} \lambda_{r}^{s} \]  \hspace{1cm} (2.95)

\[ \frac{d\lambda_{r}^{s}}{dt} = v_{r}^{s} - k_{4} \lambda_{s}^{s} + k_{5} \lambda_{s}^{s} + \omega_{r} \lambda_{r}^{s} \]  \hspace{1cm} (2.96)

where

\[ k_{1} = L_{ls} L_{lr} + L_{m} L_{ls} + L_{m} L_{lr} \]  \hspace{1cm} (2.97)

\[ k_{2} = \frac{r_{s}}{k_{1}} (L_{lr} + L_{m}) \]  \hspace{1cm} (2.98)

\[ k_{3} = \frac{r_{r}}{k_{1}} L_{m} \]  \hspace{1cm} (2.99)
\[ k_4 = \frac{r_f}{k_1} (L_{ls} + L_m) \]  

\[ k_5 = \frac{r_f}{k_1} L_m \]  

The currents of the stator and rotor are linear relationship with the flux linkages. The equations are shown as follows:

\[ i_{\alpha s} = \frac{L'_{ls} + L_m}{k_1} \lambda_{\alpha s} + \frac{L_m}{k_1} \lambda'_{\alpha r} \]  

\[ i_{\beta s} = \frac{L'_{ls} + L_m}{k_1} \lambda_{\beta s} + \frac{L_m}{k_1} \lambda'_{\beta r} \]  

\[ i_{\alpha r} = \frac{L_{ls} + L_m}{k_1} \lambda'_{\alpha r} - \frac{L_m}{k_1} \lambda_{\alpha s} \]  

\[ i_{\beta r} = \frac{L_{ls} + L_m}{k_1} \lambda'_{\beta r} - \frac{L_m}{k_1} \lambda_{\beta s} \]  

### 2.3.3 Dynamic Model in Arbitrary Reference Frame

The following equations are referred to the arbitrary rotating reference frame:

\[ v_{\alpha s} = r_s i_{\alpha s} + \frac{d \lambda_{\alpha s}}{dt} - \omega \lambda_{\beta s} \]  

\[ v_{\beta s} = r_s i_{\beta s} + \frac{d \lambda_{\beta s}}{dt} + \omega \lambda_{\alpha s} \]  

\[ v'_{\alpha r} = r'_r i'_{\alpha r} + \frac{d \lambda'_{\alpha r}}{dt} - (\omega - \omega_r) \lambda'_{\beta r} \]  

\[ v'_{\beta r} = r'_r i'_{\beta r} + \frac{d \lambda'_{\beta r}}{dt} + (\omega - \omega_r) \lambda'_{\alpha r} \]  

where

\[ \lambda_{\alpha s} = i_{\alpha s} L_{ls} + L_m (i_{\alpha s} + i'_{\alpha r}) \]  

\[ \lambda_{\beta s} = i_{\beta s} L_{ls} + L_m (i_{\beta s} + i'_{\beta r}) \]  

\[ \lambda'_{\alpha r} = i'_{\alpha r} L_{lr} + L_m (i_{\alpha s} + i'_{\alpha r}) \]  

\[ \lambda'_{\beta r} = i'_{\beta r} L_{lr} + L_m (i_{\beta s} + i'_{\beta r}) \]
also

\[ T_e = L_m (i'_{\alpha r} i_{\beta s} - i_{\alpha s} i'_{\beta r}) \]  \hspace{1cm} (2.114)

The rotor flux linkages in arbitrary rotating reference frame could be expressed as:

\[ \frac{d\lambda_{\alpha s}}{dt} = v_{\alpha s} - k_2\lambda_{\alpha s} + k_3\lambda'_{\alpha r} + \omega\lambda_{\beta s} \]  \hspace{1cm} (2.115)

\[ \frac{d\lambda_{\beta s}}{dt} = v_{\beta s} - k_2\lambda_{\beta s} + k_3\lambda'_{\beta r} - \omega\lambda_{\alpha s} \]  \hspace{1cm} (2.116)

\[ \frac{d\lambda'_{\alpha r}}{dt} = v_{\alpha r} - k_4\lambda'_{\alpha r} + k_5\lambda'_{\alpha s} + (\omega - \omega_r)\lambda_{\alpha r} \]  \hspace{1cm} (2.117)

\[ \frac{d\lambda'_{\beta r}}{dt} = v_{\beta r} - k_4\lambda'_{\beta r} + k_5\lambda'_{\beta s} - (\omega - \omega_r)\lambda_{\alpha r} \]  \hspace{1cm} (2.118)

where

\[ k_1 = L_{ls} L'_{lr} + L_m L_{ls} + L_m L'_{lr} \]  \hspace{1cm} (2.119)

\[ k_2 = \frac{r_s}{k_1} (L'_{lr} + L_m) \]  \hspace{1cm} (2.120)

\[ k_3 = \frac{r_r}{k_1} L_m \]  \hspace{1cm} (2.121)

\[ k_4 = \frac{r'_r}{k_1} (L_{ls} + L_m) \]  \hspace{1cm} (2.122)

\[ k_5 = \frac{r'_r}{k_1} L_m \]  \hspace{1cm} (2.123)

The currents of the stator and rotor are linear relationship with the flux linkages. The equations are shown as follows:

\[ i_{\alpha s} = \frac{L'_{lr} + L_m}{k_1} \lambda_{\alpha s} - \frac{L_m}{k_1} \lambda'_{\alpha r} \]  \hspace{1cm} (2.124)

\[ i_{\beta s} = \frac{L'_{lr} + L_m}{k_1} \lambda_{\beta s} - \frac{L_m}{k_1} \lambda'_{\beta r} \]  \hspace{1cm} (2.125)

\[ i'_{\alpha r} = \frac{L_{ls} + L_m}{k_1} \lambda'_{\alpha r} - \frac{L_m}{k_1} \lambda_{\alpha s} \]  \hspace{1cm} (2.126)

\[ i'_{\beta r} = \frac{L_{ls} + L_m}{k_1} \lambda'_{\beta r} - \frac{L_m}{k_1} \lambda_{\beta s} \]  \hspace{1cm} (2.127)
2.4 Comparisons between Permanent Magnetic Synchronous Machine and Induction Machine

As ISG AC machine candidates, the advantages and disadvantages of the Permanent Magnetic Synchronous Machine (PMSM) and the Induction Machine (IM) are compared from three categories: mechanical, magnetic and electrical.

The permanent magnet synchronous machine has two electromagnetic components: the rotating magnetic field constructed with permanent magnets; and the stationary armature constructed with electrical windings located in a slotted iron core. The PMSM’s rotor is made by high-energy rare earth materials. The stationary iron core is made by laminated electrical grade steel. Electrical windings are made from high purity copper conductors.

Induction machine also has two electromagnetic components: the rotating magnetic field constructed with high conductivity, high strength bars located in a slotted iron core to form a squirrel cage; and the stationary armature similar to the one described in the PM technology.

The technology of induction machine is based on the relatively mature electric motor technology. Induction machines are perhaps the most common types of electric motors throughout the industry. Early developments in induction generators were made using fixed capacitors for excitation, since suitable active power devices were not available. This resulted in unstable power output since the excitation could not be adjusted as the load or speed deviated from the nominal values. With the availability of high power switching devices, induction generator can be provided with adjustable excitation and operate in isolation in a stable manner with appropriate controls.

Now we can compare two machines with the same power rating from three categories: mechanical, magnetic and electrical.
Mechanical:

PMSM’s size is smaller than IM’s because its stator windings only carry the load current while IM’s stator windings must carry both the load current and excitation current. PMSM’s inverter is smaller than IM’s because it only delivers load current while IM’s must supply additional excitation current to the squirrel cage rotor. IM’s air gap width is shorter than PMSM’s in order to minimize the excitation current. The thermal property for induction machines is good although the electromagnetic torque is lower at high temperatures. However, the rotor permanent magnet has poor thermal property for PMSM in high temperatures. PMSM’s price is typically higher than IM’s for the use of permanent magnets. It means that lower cost materials are used for induction machines. Obviously, Rare earth permanent magnets are substantially more expensive than the electrical steel used in electromagnets.

Magnetic:

PMSM’s rotor has the demagnetization risk by inrush load current. It is complex for the magnet handling in PMSM. It requires special machining operations and must be retained on the rotor structure by installation of the containment structure. The pre-charged handling of permanent magnets is generally difficult in production shops. These requirements increase the cost of labor. PMSM has higher losses than IM due to the higher stator current needed to compensate magnet flux and limit the induced voltage in flux weakening region. Also, PMSM has higher eddy current losses than IM.

Electrical:

IM need the additional reactive power to provide the rotor excitation current in constant flux zone. However, there is no additional reactive power source needed for PMSM. IM’s starting current is medium to high because the excitation current is needed. On the other
<table>
<thead>
<tr>
<th>Machine size</th>
<th>IM</th>
<th>PMSM</th>
<th>Remark</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inverter size</td>
<td>Medium</td>
<td>Small</td>
<td>Excitation needed by IM</td>
</tr>
<tr>
<td>Air gap width</td>
<td>Short</td>
<td>Medium</td>
<td>Excitation needed by IM</td>
</tr>
<tr>
<td>Thermal properties</td>
<td>Good</td>
<td>Poor</td>
<td>PM’s thermal high temp property</td>
</tr>
<tr>
<td>Cost of materials</td>
<td>Low</td>
<td>High</td>
<td>Rare earth PM is expensive</td>
</tr>
</tbody>
</table>

Table 2.1: Mechanical comparison between IM and PMSM

side, PMSM’s starting current is lower because it has permanent magnets in its rotor. If the ISG system fails in flux weakening region, the IM ISG system is de-excitation within several millisecond, preventing the hazardous situations. Unfortunately, for the PMSM ISG system, the back EMF is immense due to stator current is out of control in flux weakening region. The failed PMSM will continue to draw energy until the generator is stopped. For high-speed generators, it means a long enough duration during which further damage to electrical and mechanical components would occur.

The overall comparison for two machines is shown in Table 2.1, 2.2 and 2.3. Based upon the comparison, the induction machine is firstly chosen for its reliable, robust and safety for the 28V ISG system for aircraft application. A PMSM is chosen and tested for 270V ISG system. Details of the 28V ISG system with IM and 270V ISG system with PMSM will be discussed in Chapter 3 and Chapter 4.
### Magnetic Comparison

<table>
<thead>
<tr>
<th></th>
<th>IM</th>
<th>PMSM</th>
<th>Remark</th>
</tr>
</thead>
<tbody>
<tr>
<td>Demagnetization</td>
<td>N/A</td>
<td>High</td>
<td>PM’s property</td>
</tr>
<tr>
<td>Magnet handling</td>
<td>N/A</td>
<td>Difficult</td>
<td>Handling is complex</td>
</tr>
<tr>
<td>Flux weakening</td>
<td>High losses</td>
<td>Low losses</td>
<td>Higher Id is required for PMSM</td>
</tr>
<tr>
<td>Eddy current losses</td>
<td>High</td>
<td>N/A</td>
<td>PM’s property</td>
</tr>
</tbody>
</table>

Table 2.2: Magnetic comparison between IM and PMSM

### Electrical Comparison

<table>
<thead>
<tr>
<th></th>
<th>IM</th>
<th>PMSM</th>
<th>Remark</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reactive power demand</td>
<td>No</td>
<td>Yes</td>
<td>Excitation required by IM</td>
</tr>
<tr>
<td>Starting current</td>
<td>High</td>
<td>Low</td>
<td>Excitation required by IM</td>
</tr>
<tr>
<td>Failure at high speed</td>
<td>Safe</td>
<td>Unsafe</td>
<td>Uncontrollable Back-EMF for PMSM</td>
</tr>
</tbody>
</table>

Table 2.3: Electrical comparison between IM and PMSM
CHAPTER 3

SENSORLESS CONTROL FOR PERMANENT MAGNET SYNCHRONOUS MACHINE BASED INTEGRATED STARTER GENERATOR

In this chapter, the model of a basic salient pole synchronous machine with eddy current effects is presented. Then the sensorless control of permanent magnet synchronous machine is introduced. There are two parts for the sensorless control for a PMSM operating over a wide speed range: (1) high frequency injection method with magnetic pole identification in zero and low speed range; (2) voltage model sensorless control algorithm in high speed range.

3.1 High Frequency Injection Based Sensorless Control Technology for PM Synchronous Machine

In this dissertation, the high frequency signal injection (HFI) method is chosen for rotor position estimation of PMSMs with rotor saliency. When high frequency signal is injected into stator, it produces a rotating magnetic field with constant magnitude. Moreover, this rotating magnetic field will be modulated by the rotor saliency and then result in a high frequency current carrier wave including information of rotor position. Thus, through demodulation of this carrier wave, the rotor position can be obtained.
A high frequency component can be superimposed on the fundamental excitation to create a position-dependent carrier current signal. The location of the magnetic axis can be estimated by using the carrier signal current resulting from the interaction between the carrier signal voltage and the spatial saliency. This is the principle of the high frequency injection based sensorless control. The voltage and flux linkage equations in the stationary reference frame are given by:

\[
v_{\alpha\beta} = R_s i_{\alpha\beta} + \frac{d}{dt} \lambda_{\alpha\beta}
\]  

(3.1)

\[
\lambda_{\alpha\beta} = L_s (2\theta) i_{\alpha\beta} + \lambda_m \begin{bmatrix} \sin \theta \\ \cos \theta \end{bmatrix}
\]  

(3.2)

Where

\[
L_s(2\theta) = \begin{bmatrix} L_s + \frac{3}{2}L_0s + \frac{3}{2}L_2s \cos(2\theta) & \frac{3}{2}L_2s \sin(2\theta) \\ \frac{3}{2}L_2s \sin(2\theta) & L_s + \frac{3}{2}L_0s \frac{3}{2}L_2s \cos(2\theta) \end{bmatrix}
\]  

(3.3)

The injected high frequency component is:

\[
v_{\alpha\beta_{s,HFI}} = V_{HFI} \begin{bmatrix} -\sin(\omega_it) \\ \cos(\omega_it) \end{bmatrix}
\]  

(3.4)

Because \( \omega_i L \gg R_s \), the stator flux linkages can be obtained with Equation 3.5.

\[
\lambda_{\alpha\beta_{s,HFI}} = L_s(2\theta_{r,HFI}) i_{\alpha\beta_{s,HFI}} \approx \int v_{\alpha\beta_{s,HFI}} dt
\]  

(3.5)

The stator current is derived with Equation 3.6 and Equation 3.7.

\[
i_{\alpha\beta_{s,HFI}} = \left[ L_s(2\theta_{r,HFI}) \right]^{-1} \int v_{\alpha\beta_{s,HFI}} dt
\]  

(3.6)

\[
i_{\alpha\beta_{s,HFI}} = \frac{V_{HFI}}{\omega_i L_s(L_s + 3L_0s)} \begin{bmatrix} (L_s + \frac{3}{2}L_0s) \cos(\omega_it) - \frac{3}{2}L_2s \cos(2\theta_{r,HFI} - \omega_it) \\ (L_s + \frac{3}{2}L_0s) \sin(\omega_it) - \frac{3}{2}L_2s \sin(2\theta_{r,HFI} - \omega_it) \end{bmatrix}
\]  

(3.7)
The HFI diagram is shown in Figure 3.1. When a rotor position or saturation dependent variation in the stator transient inductance exists, its shape can be tracked by injecting a high frequency carrier signal voltage. This inherent rotor position spatial saliency makes it possible to use rotating or pulsating vector carrier frequency excitation to reliably identify and track the orientation of the $d$ and $q$ axes even when the rotor is at standstill.

For example, during HFI process, the demodulated process is implemented through two cascaded synchronous filters, a high pass filter (HPF) synchronized with the positive sequence and a low pass filter (LPF) with the negative sequence. By a high pass filter with the positive sequence current, high frequency components remain. Then through a cascaded synchronous low pass filter with the negative sequence current, components retaining saliency angle information are obtained. Figure 3.1 shows the estimated rotor angle by the demodulated method with the two synchronous filters.

Figure 3.1: Rotor position estimation based on high frequency injection method
The demodulation process of the high frequency carrier wave is simulated by MATLAB/SIMULINK. The frequency of signal injection is 200Hz and the PMSM is rotating in a very low speed (30rpm). As can be seen in Figure 3.2, high frequency currents have been superimposed on the fundamental excited current.

![Image of simulation results of \( i_\alpha \) and \( i_\beta \)](image)

**Figure 3.2**: Simulation results of \( i_\alpha \) and \( i_\beta \)

Figure 3.3, 3.4 and 3.5 show the intermediate results on different demodulation steps. \( i_{\alpha 1} \) and \( i_{\beta 1} \) are the stator current in the rotating reference frame synchronous with stator voltage vector \( u_c \). After filtered by the HPF, there is no positive current item left in \( i_{\alpha 2} \) and \( i_{\beta 2} \). Moreover, through the cascaded LPF, the double frequency fundamental currents \( i_{\alpha,dem} \) and \( i_{\beta,dem} \) shown in Figure 3.9 have been achieved. Then, the \( 2\dot{\theta}_r \) can be easily...
Figure 3.3: $i_{\alpha 1}$ and $i_{\beta 1}$

Figure 3.4: Details of $i_{\alpha 1}$ and $i_{\beta 1}$
Figure 3.5: $i_{\alpha 2}$ and $i_{\beta 2}$

Figure 3.6: Details of $i_{\alpha 2}$ and $i_{\beta 2}$
Figure 3.7: $i_{\alpha 3}$ and $i_{\beta 3}$

Figure 3.8: Details of $i_{\alpha 3}$ and $i_{\beta 3}$
Figure 3.9: $i_{\alpha\_dem}$ and $i_{\beta\_dem}$

Figure 3.10: $2\hat{\theta}_r$
obtained that shown in Figure 3.10. The simulation results show that the estimated angle matches the real angle very well.

The HPF and LPF are important parts in the HFI based rotor position estimation. The details for digital implementations HPF and LPF design are derived in Appendix A.

3.2 Modeling of PM Synchronous Machine Considering Eddy Current Effects

A PM synchronous machine is generally modeled as a synchronous machine with some rotor saliency. The eddy currents can be modeled as the currents circulating through a pair of short-circuited coils along the $d$ and $q$ axes. Figure 3.11 shows the basic structure of a PM synchronous machine with shorted coils emulating the eddy current effects. The $d$ axis is aligned with the N-pole of the rotor and $q$ axis 90 degrees apart from the $d$ axis.

The voltage and flux linkage equations in the stationary reference frame are given by:

\[ v_{\alpha\beta s} = R_s i_{\alpha\beta s} + \frac{d}{dt} \lambda_{\alpha\beta s} \]  
\[ (3.8) \]
\[ v_{\alpha\beta r} = R_r i_{\alpha\beta r} + \frac{d}{dt} \lambda_{\alpha\beta r} \]  
\[ (3.9) \]
\[ \lambda_{\alpha\beta s} = L_s(2\theta) i_{\alpha\beta s} + L_{sr}(i_{dqr}) + \lambda_m(\theta) \]  
\[ (3.10) \]
\[ \lambda_{dqr} = L_{sr}^T(2\theta) i_{\alpha\beta s} + L_{sr}(i_{dqr}) + \lambda_m(\theta) \]  
\[ (3.11) \]

where

\[ L_s(2\theta) = \begin{bmatrix} L_s + \frac{3}{2}L_{0s} + \frac{3}{2}L_{2s}\cos(2\theta) & \frac{3}{2}L_{2s}\sin(2\theta) \\ \frac{3}{2}L_{2s}\sin(2\theta) & L_s + \frac{3}{2}L_{0s} + \frac{3}{2}L_{2s}\cos(2\theta) \end{bmatrix} \]  
\[ (3.12) \]
\[ L_{sr}(\theta) = \begin{bmatrix} L_{dlr}\sin\theta & L_{qsr}\sin\theta \\ -L_{dqr}\cos\theta & -L_{qsr}\cos\theta \end{bmatrix} \]  
\[ (3.13) \]

Without considering eddy current effects, the $q$ and $d$ axes currents in the related short-circuited coils in Equation 3.8 through Equation 3.11 will be zero. Hence the voltage
Figure 3.11: PM synchronous machine with shorted coils accounting for eddy currents
equations without considering eddy current effects are reduced to Equation 3.14 and Equation 3.15.

\begin{align*}
    v_{\alpha\beta s} &= R_s i_{\alpha\beta s} + \frac{d}{dt} \lambda_{\alpha\beta s} \\
    \lambda_{\alpha\beta s} &= L_s (2\theta) i_{\alpha\beta s} + \lambda_m \begin{bmatrix} \sin \theta \\ \cos \theta \end{bmatrix}
\end{align*}

(3.14)  
(3.15)

However, in the HFI estimation method, a set of high frequency voltages, viewed as a voltage vector and not synchronized with the rotor rotation, is always applied. Therefore, substantial eddy currents will be induced on the stator and rotor cores. It is conceivable that the eddy currents will be reflected into the stator windings and cannot be neglected. As an approximation, a PM synchronous machine with short-circuited coils on the rotor can be utilized to account for eddy current effects. The improved rotor position estimation and sensorless control scheme based on HFI considering the eddy current effects are shown in Figure 3.12. The injected high frequency voltages are superimposed on the fundamental excitation and a signal demodulator is used to extract the rotor position information.

![Figure 3.12: Rotor position estimation with eddy current effect compensation](image)

Figure 3.12: Rotor position estimation with eddy current effect compensation
Note that the voltage and flux linkage equations for the PM synchronous machine with shorted coils are higher order and more complicated than those for a conventional PMSM. The additional components accounting for the eddy current are also rotor position dependent. Therefore, the eddy current effect compensation should be considered. The injected high frequency voltages are expressed in Equation 3.16. In the high frequency voltage equations, the stator and rotor resistance is negligible because for high frequency injection, $\omega_i L \gg R_s$. Equation 3.17 shows the relationship between the stator flux linkages, stator currents and rotor eddy currents with high frequency voltage excitation. The stator currents can be derived from Equation 3.18 and the components representing the eddy current effects are $i'_{dqr,HFI}$ shown in Equation 3.19. With the eddy current effects properly modeled and compensated as shown in Figure 3.13, we can obtain the desired components ($i'_{\alpha\beta,HFI}$) described by Equation 3.20 that are to be demodulated by the conventional HFI method. Then, the rotor position information can be extracted.

$$v_{\alpha\beta, HFI} = V_{HFI} \begin{bmatrix} -\sin(\omega_i t) \\ \cos(\omega_i t) \end{bmatrix}$$ (3.16)

$$\lambda_{\alpha\beta, HFI} = L_s (2\theta_{r, HFI}) i_{\alpha\beta, HFI} + L_{sr} (\theta_{r, HFI}) i_{dqr, HFI} \approx \int v_{\alpha\beta, HFI} dt$$ (3.17)

$$i_{\alpha\beta, HFI} = \left[ L_s (2\theta_{r, HFI}) \right]^{-1} \int v_{\alpha\beta, HFI} dt - \left[ L_s (2\theta_{r, HFI}) \right]^{-1} L_{sr} (\theta_{r, HFI}) i_{dqr, HFI}$$

$$= i'_{\alpha\beta, HFI} - i'_{dqr, HFI}$$ (3.18)
\[ i'_{dq_r_HFI} = [L_s(2\theta_{r_HFI})]^{-1}L_{sr}(\theta_{r_HFI}) \]

\[ = [L_s(2\theta_{r_HFI})]^{-1}[\lambda_{\alpha\beta_{s_HFI}} - L_s(2\theta_{r_HFI})i_{\alpha\beta_{s_HFI}}] \]

\[ = \frac{V_{HFI}}{\omega L_{ls}(L_{ls} + 3L_{0s})} \left[ (L_{ls} + \frac{3}{2}L_{0s}) \cos(\omega t) - \frac{3}{2}L_{2s} \cos(2\theta_{r_HFI} - \omega t) \right] - i_{\alpha\beta_{s_HFI}} \]

\[ (3.19) \]

\[ i'_{\alpha\beta_{s_HFI}} = \frac{V_{HFI}}{\omega L_{ls}(L_{ls} + 3L_{0s})} \left[ (L_{ls} + \frac{3}{2}L_{0s}) \cos(\omega t) - \frac{3}{2}L_{2s} \cos(2\theta_{r_HFI} - \omega t) \right] \]

\[ (3.20) \]

Figure 3.13: Eddy current effect compensation

3.3 Identifying Permanent Magnet Polarity

The principle of identifying the magnet pole based on magnetic saturation effects has been discussed in literature and is illustrated in Figure 3.14 (a), (b) and (c). Initially, the
flux increases in direct proportion to the increase of currents. Further increases in currents result in progressively smaller increases in flux because of magnetic saturation. As shown in Figure 3.14 (c), with the permanent magnet excitation alone, the original operating point is \((i_0, \lambda_0)\). However, with the stator excitation voltage applied in the same polarity with respect to that of the permanent magnet, the stator currents will increase from \(i_0\) to \(i_1\), and operating point from \((i_0, \lambda_0)\) to \((i_1, \lambda_1)\) due to the deep saturation. On the other hand, when the applied stator excitation voltage and magnetic pole are in the opposite directions, the magnetic path will not be saturated. The stator current will change from \(i_0\) to \(i_2\). At the same time, the operating point moves accordingly from point \((i_0, \lambda_0)\) to \((i_2, \lambda_2)\). In these two cases, the applied volt-seconds are the same but different in polarity. Due to core saturation nonlinearity, the magnitudes of current variations are different. Therefore, we can detect the polarity of permanent magnets mounted on the rotor.

![Figure 3.14](image.png)

Figure 3.14: Magnetic pole identification based on saturation effects

When the magnetic pole is in an arbitrary initial position as shown in Figure 3.14 (a), we have to detect the axis of the permanent magnet first, normally achieved by the high
frequency voltage injection method but the polarity of the magnet is left unknown. In the second step, according to the axis direction of the permanent magnet, positive and negative pilot voltages are applied sequentially in a controlled mode such that the volt-seconds are the same in both cases. When the positive voltage is applied, the current incremental is $+|i_1|$ and when negative the current incremental is $-|i_2|$. In the third step, we compare the magnitudes of the current incremental. If $|i_1|$ is larger than $|i_2|$, the magnetic pole that corresponds to $|i_1|$ is the North and the rotor position in the range from 0 to $\pi$. Similarly, when the magnetic pole is in the orientation as shown in Figure 3.14 (b), the corresponding current $|i_2|$ with the positive stator excitation applied will be smaller than current $|i_1|$ with the negative stator excitation applied. Therefore, the magnetic pole that corresponds with $|i_1|$ is the North Pole. In this case, the rotor position angle is actually an angle between $\pi$ and $2\pi$. The angle will be $\theta_r + \pi$.

In the proposed magnetic poles identification scheme, the principles discussed above are applied but the voltages are based on the space vector PWM that is readily available in the controller hardware and software. The system diagram for HFI rotor position estimation with magnetic pole identification is shown in Figure 3.15. The space vector is defined as shown in Figure 3.16. When we know the $d$ axis direction of the rotor, we can conveniently figure out the phase windings closest to the rotor poles. Then, a group of pilot voltage vectors will be applied to the phase windings closest to the rotor axis to detect the magnetic poles. The pilot voltages are composed of one positive pulse and one negative pulse alternatively. Normally, the rotor’s initial position could be arbitrary and the magnetic pole is not perfectly aligned with any phase windings. As an example, the $V_{NS}$ and $-V_{NS}$ shown in Figure 3.17 are one pair of pilot voltage vectors corresponding to an arbitrary initial rotor position. The applied SVPWM switching signals are shown in Figure 3.18. The
Figure 3.15: System diagram for rotor position estimation
sensed three phase currents through A/D circuits will be transformed into the synchronous reference frame first to obtain the relative information on the polarity of magnetic pole. If the current in the synchronous reference frame is larger when $V_{NS}$ applied than the current when $-V_{NS}$ is applied, the magnetic pole must be the North pole. Note that the applied voltage vectors should be sufficiently large to cause the magnetic saturation.
Figure 3.17: The pilot voltage pulses

Figure 3.18: Pilot voltage vectors in one PWM cycle for magnetic pole identification: $V_{NS}$ and $-V_{NS}$
3.4 Simulation and Experimental Results of High Frequency Injection and Permanent Magnet Polarity Identification

For comparison, a PM synchronous machine with and without considering eddy current effects is simulated with MATLAB/SIMULINK. To identify and investigate the eddy current effects, the rotor speed is maintained at a constant speed of 50\textit{rpm} and stator windings are excited with a high frequency voltage of 500Hz.

![Simulation results of \(\alpha\) axis current, \(2\hat{\theta}_{r-HFI}\) and \(\hat{\theta}_{r-HFI}\) for the synchronous machine without shorted coils](image)

Figure 3.19 shows the simulated phase current \(i_{\alpha-HFI}\), the estimated rotor angle \(2\hat{\theta}_{r-HFI}\) and \(\hat{\theta}_{r-HFI}\) for the PM synchronous machine without shorted coils. Figure 3.20 shows the
similar results based on the dynamic equations of PM synchronous machine considering the eddy current effects. As indicated by the high frequency current envelopes, the current responses in both Figure 3.19 and Figure 3.20 contain obvious information of the rotor positions. However, careful examination of the current responses also reveals the major differences in these two cases. In particular, the amplitude of high frequency current in Figure 3.19 is modulated by the rotor saliency strongly and it is quite easy to identify the rotor position for sensorless control. Nevertheless, in Figure 3.20, both the amplitude and phase angle of the phase current envelope are seriously interfered by the eddy current effects. As a result, the rotor position estimation has a difference of 48 degrees compared to that of the case in Figure 3.19.

Experimental verification of rotor position estimation including eddy current effects has been accomplished based on a six-pole 5HP PM synchronous machine. The verification has been focused on for the PM synchronous machine at zero and low speeds. In the experiment, the rotor of the machine is commanded rotating at very low speeds and subsequently reverses its rotating direction. Figure 3.21 shows the estimated rotor position waveforms of $\hat{\theta}_{r-HFI}$ and $2\hat{\theta}_{r-HFI}$ with the phase current in a very low speed range ($\pm20\text{rpm}$) in no load condition. Figure 3.22 shows the experimental results when the PM synchronous machine is running at a constant low speed (45rpm) with regulated currents of its fundamental frequency. The results show that the rotor position obtained through the proposed scheme works stably in the very low speed region.

In the experimental testing, the rotor position is compensated with the proposed eddy current effect compensator as shown in Figure 3.24. In the figure, $\theta_{r-\text{encoder}}$ is the real rotor position measured by the encoder. It is considered a reference for comparison. $\hat{\theta}_{r-HFI-cps}$ is the estimated rotor position using HFI with eddy current effect compensation while
Figure 3.20: Simulation results of $\alpha$ axis current, $2\hat{\theta}_{r-HF1}$ and $\hat{\theta}_{r-HF1}$ for the synchronous machine with shorted coils
Figure 3.21: Low speed ($\pm 1Hz$) waveforms without load (top trace is $\hat{\theta}_{r-HFI}$, middle is $2\hat{\theta}_{r-HFI}$ and bottom trace is the phase A current (10A/div))
Figure 3.22: Low speed (2.25Hz) with fundamental current waveforms (top trace is $2\dot{\theta}_{HF}$, middle is $\dot{\theta}_{HF}$ and bottom is the phase A current (10A/div))
without considering eddy current effects. As shown clearly, without compensation, the rotor position estimation error is as large as about 50 degrees, indicating the necessity of considering eddy current effects.

One of the experimental results for N-S pole identification is shown in Figure 3.23. The waveform shown in Channel A is the applied pilot voltage pulses and Channel D the stator current that transformed into the synchronous reference frame at zero speed (10A/div). The magnitude of positive current is 32.5A and the magnitude of negative current is 34A with the aforementioned pilot voltage vectors based on SVPWM ($V_{NS}$ and $-V_{NS}$). Obviously, the magnetic pole corresponding to applied negative current is North pole and accordingly the rotor position angle is between $\pi$ and $2\pi$.

The reference angle of the encoder mounted on the rotor shaft is used to calibrate the N-S pole identification method. The detected magnetic pole information is applied with $2\hat{\theta}_{r-HFI}$ to generate the estimated rotor position $\hat{\theta}_{r-HFI}$ as shown in Figure 3.25. The $\hat{\theta}_{r-HFI}$ matches the real rotor position $\theta_{r-encoder}$ very well. Figure 3.25 shows the estimated rotor position waveforms of $2\hat{\theta}_{r-HFI}$, $\hat{\theta}_{r-HFI}$ and phase current in a constant low speed range with regulated currents.

The results show that the proposed magnetic pole detection scheme works very well for the estimation of the arbitrary initial rotor position.
Figure 3.23: Magnetic pole identification (Channel A: applied voltage pulses; Channel D: the current in synchronous reference frame)

Figure 3.24: $\hat{\theta}_{r-HFI}$, $\hat{\theta}_{r-HFI-cps}$ and $\theta_{r-encoder}$
this section. The Back-EMF is easy to obtain from the synchronous machine model based on the Equation 3.21. The Back-EMF $\bar{E}_s$ is shown in Equation 3.22. We can see that $\bar{E}_s$ is related to $\bar{I}_d$ and $\bar{I}_q$. However, $\bar{I}_d$ and $\bar{I}_q$ cannot be achieved directly from the feedback system because the rotor position is unknown. So Equation 3.21 should be changed a little bit to get another equation to express the Back-EMF. All variables are known in the right of Equation 3.22. Also the left side of the equation is in the same direction with $E_s$. So, in Equation 3.23, $\bar{E}'_s$ is defined to replace the left of Equation 3.22. Note that the direction of

Figure 3.25: Compensated $\hat{\theta}_{r-HF}$ and NS pole identification

3.5 Synchronous Machine Voltage Model Sensorless Technology

Because of the limitation of the high frequency injection technology in the high speed range, the high speed rotor position estimation method based on Back-EMF is proposed in this section. The Back-EMF is easy to obtain from the synchronous machine model based on the Equation 3.21. The Back-EMF $\bar{E}_s$ is shown in Equation 3.22. We can see that $\bar{E}_s$ is related to $\bar{I}_d$ and $\bar{I}_q$. However, $\bar{I}_d$ and $\bar{I}_q$ cannot be achieved directly from the feedback system because the rotor position is unknown. So Equation 3.21 should be changed a little bit to get another equation to express the Back-EMF. All variables are known in the right of Equation 3.22. Also the left side of the equation is in the same direction with $E_s$. So, in Equation 3.23, $\bar{E}'_s$ is defined to replace the left of Equation 3.22. Note that the direction of
\( E'_s \) is same as \( E_s \).

\[
\vec{V}_s = \vec{E}_s + \vec{I}_s \vec{R}_s + j \vec{X}_d \vec{I}_d + j \vec{X}_q \vec{I}_q \\
= \vec{E}_s + \vec{I}_s \vec{R}_s + j \vec{X}_d \vec{I}_d + j \vec{X}_q \vec{I}_q - j \vec{X}_d \vec{I}_q + j \vec{X}_q \vec{I}_d \\
= \vec{E}_s + \vec{I}_s \vec{R}_s + j \vec{I}_d (X_d - X_q) + j (\vec{I}_q + \vec{I}_d) X_q \\
= \vec{E}_s + \vec{I}_s \vec{R}_s + j \vec{I}_d (X_d - X_q) + j \vec{I}_s X_q
\]  

Then reorganize the above equation with the following format:

\[
\vec{E}_s + j \vec{I}_d (X_d - X_q) = \vec{V}_s - \vec{I}_s - j \vec{I}_s X_q
\]  

(3.21)

We define

\[
\vec{E}'_s = \vec{E}_s + j \vec{I}_d (X_d - X_q)
\]  

(3.22)

So,

\[
\vec{E}'_s = \vec{V}_s - \vec{I}_s \vec{R}_s - j \vec{I}_s X_q = \vec{V}_s - \vec{I}_s \vec{R}_s - j \vec{I}_s \omega L_q = \vec{V}_s - \vec{I}_s \vec{R}_s - \frac{d \vec{I}_s}{dt} \omega L_q
\]  

(3.23)

In order to get the rotor position information from the vector \( \vec{E}'_s \), Equation 3.24 will be rewritten to Equation 3.25 and 3.26 in synchronous reference frame.

\[
\vec{E}'_\alpha = \vec{V}_\alpha - \vec{I}_\alpha \vec{R}_s - \frac{d \vec{I}_\alpha}{dt} L_q
\]  

(3.25)

\[
\vec{E}'_\beta = \vec{V}_\beta - \vec{I}_\beta \vec{R}_s - \frac{d \vec{I}_\beta}{dt} L_q
\]  

(3.26)

Because the rotor position is related to the flux linkages, after integrating Equation 3.25 and 3.26, the flux linkages are shown in Equation 3.27 and 3.28.

\[
\lambda'_\alpha = \int \vec{E}'_\alpha dt = \int [\vec{V}_\alpha - \vec{I}_\alpha \vec{R}_s - \frac{d \vec{I}_\alpha}{dt} L_q] dt = \int [\vec{V}_\alpha - \vec{I}_\alpha \vec{R}_s] dt - \vec{I}_\alpha L_q
\]  

(3.27)

\[
\lambda'_\beta = \int \vec{E}'_\beta dt = \int [\vec{V}_\beta - \vec{I}_\beta \vec{R}_s - \frac{d \vec{I}_\beta}{dt} L_q] dt = \int [\vec{V}_\beta - \vec{I}_\beta \vec{R}_s] dt - \vec{I}_\beta L_q
\]  

(3.28)

So the estimated rotor position is:

\[
\theta = \arctan \left( \frac{\lambda'_\beta}{\lambda'_\alpha} \right)
\]  

(3.29)
In order to realize the synchronous motor sensorless algorithm in the DSP system, all equations need to be discretized.

\[ e'_{\alpha-1} = v_\alpha - R_s i_\alpha \]  
\[ e'_{\beta-1} = v_\beta - R_s i_\beta \]

Equation 3.30 and Equation 3.31 are discretized as follows:

\[ e'_{\alpha-1}(k) = v_\alpha(k) - R_s i_\alpha(k) \]  
\[ e'_{\beta-1}(k) = v_\beta(k) - R_s i_\beta(k) \]

The integration part of Equation 3.27 and Equation 3.28 should be discretized:

\[ \lambda'_{\alpha-1} = \int e'_{\alpha-1} dt \]  
\[ \lambda'_{\beta-1} = \int e'_{\beta-1} dt \]

Equation 3.34 and Equation 3.35 are discretized by trapezoidal approximation as:

\[ \lambda'_{\alpha-1}(k) = \lambda'_{\alpha-1}(k-1) + \frac{T_s}{2} (e'_{\alpha-1}(k) + e'_{\alpha-1}(k-1)) \]  
\[ \lambda'_{\beta-1}(k) = \lambda'_{\beta-1}(k-1) + \frac{T_s}{2} (e'_{\beta-1}(k) + e'_{\beta-1}(k-1)) \]

Then the Equation 3.27 and Equation 3.28 can be rewritten as:

\[ \lambda'_\alpha = \lambda'_{\alpha-1} - L_q i_\alpha \]  
\[ \lambda'_\beta = \lambda'_{\beta-1} - L_q i_\beta \]

The discrete formats of Equation 3.38 and Equation 3.39 are:

\[ \lambda'_{\alpha}(k) = \lambda'_{\alpha-1}(k) - L_q i_\alpha(k) \]
\[ \lambda'_\beta(k) = \lambda'_{\beta-1}(k) - L_q i_\beta(k) \]  

(3.41)

So the estimated rotor position is

\[ \theta(k) = \arctan\left(\frac{\lambda'_\beta(k)}{\lambda'_\alpha(k)}\right) \]  

(3.42)

Finally, all discrete equations should be normalized into per-unit by the specified base quantities. Equation 3.32 and Equation 3.33 are normalized by dividing the base phase voltage \( V_{\text{base}} \):

\[ e'_{\alpha-1,\text{pu}}(k) = v_{\alpha,\text{pu}}(k) - \frac{I_{\text{base}} R_s}{V_{\text{base}}} i_{\alpha,\text{pu}}(k) \]  

(3.43)

\[ e'_{\beta-1,\text{pu}}(k) = v_{\beta,\text{pu}}(k) - \frac{I_{\text{base}} R_s}{V_{\text{base}}} i_{\beta,\text{pu}}(k) \]  

(3.44)

Equation 3.36 and Equation 3.37 are normalized by dividing the base flux linkage, where \( \lambda_b = L_{\text{base}} I_{\text{base}} \):

\[ \lambda'_{\alpha-1,\text{pu}}(k) = \lambda'_{\alpha-1,\text{pu}}(k-1) + \frac{T_s}{2 L_{\text{base}} I_{\text{base}}} \left( e'_{\alpha-1,\text{pu}}(k) + e'_{\alpha-1,\text{pu}}(k-1) \right) \]  

(3.45)

\[ \lambda'_{\beta-1,\text{pu}}(k) = \lambda'_{\beta-1,\text{pu}}(k-1) + \frac{T_s}{2 L_{\text{base}} I_{\text{base}}} \left( e'_{\beta-1,\text{pu}}(k) + e'_{\beta-1,\text{pu}}(k-1) \right) \]  

(3.46)

Equation 3.40 and Equation 3.41 are normalized by dividing the base flux linkage, where \( \lambda_b = L_{\text{base}} I_{\text{base}} \):

\[ \lambda'_{\alpha,\text{pu}}(k) = \lambda'_{\alpha-1,\text{pu}}(k) - \frac{L_q}{L_{\text{base}}} i_{\alpha,\text{pu}}(k) \]  

(3.47)

\[ \lambda'_{\beta,\text{pu}}(k) = \lambda'_{\beta-1,\text{pu}}(k) - \frac{L_q}{L_{\text{base}}} i_{\beta,\text{pu}}(k) \]  

(3.48)

So the estimated rotor position is:

\[ \theta_{\text{pu}}(k) = \arctan\left(\frac{\lambda'_{\beta,\text{pu}}(k)}{\lambda'_{\alpha,\text{pu}}(k)}\right) \]  

(3.49)

The discrete-time, per-unit equations are rewritten in terms of constants.

\[ e'_{\alpha-1,\text{pu}}(k) = v_{\alpha,\text{pu}}(k) - K_1 i_{\alpha,\text{pu}}(k) \]  

(3.50)
\[ e'_{\beta-1,pu}(k) = v_{\beta,pu}(k) - K_1 i_{\beta,pu}(k) \] (3.51)

\[ \lambda'_{\alpha-1,pu}(k) = \lambda'_{\alpha-1,pu}(k - 1) + K_2 (e'_{\alpha-1,pu}(k) + e'_{\alpha-1,pu}(k - 1)) \] (3.52)

\[ \lambda'_{\beta-1,pu}(k) = \lambda'_{\beta-1,pu}(k - 1) + K_2 (e'_{\beta-1,pu}(k) + e'_{\beta-1,pu}(k - 1)) \] (3.53)

\[ \lambda'_{\alpha,pu}(k) = \lambda'_{\alpha-1,pu}(k) - K_3 i_{\alpha,pu}(k) \] (3.54)

\[ \lambda'_{\beta,pu}(k) = \lambda'_{\beta-1,pu}(k) - K_3 i_{\beta,pu}(k) \] (3.55)

\[ \theta_{pu}(k) = \arctan \left( \frac{\lambda'_{\beta,pu}(k)}{\lambda'_{\alpha,pu}(k)} \right) \] (3.56)

Where in above equations:

- \( T_s \): Sampling period;
- \( K_1 = \frac{l_{base} R_s}{V_{base}} \);
- \( K_2 = \frac{T_s V_{base}}{l_{base} I_{base}} \);
- \( K_3 = \frac{L_q}{L_d} \).

### 3.6 Experimental Results of Voltage Model Sensorless Algorithm

Experimental verification of rotor position estimation with voltage model method has been accomplished based on the aforementioned six-pole 5HP PM synchronous machine. The verification has been focused on testing with the PM synchronous machine at high speeds. Figure 3.26 shows the estimated rotor position waveforms of \( \hat{\theta}_{rVM} \) and the phase current with a constant high speed (1200 rpm) with no load. The experimental results show that the rotor position obtained through the proposed scheme works well in the high speed region.
Figure 3.26: Waveforms of voltage model rotor position estimation (Channel C: estimated rotor position at 80Hz; Channel D: phase current (20A/div))
3.7 Full Speed Range Sensorless Control for PMSM

The high frequency injection based rotor position detection method is proper for the zero speed and lower speed range. In addition, the voltage model based rotor position estimation method works well in higher speed range. In order to implement the full speed range sensorless control for PMSM, the two methods should be combined together. A crossover function transition method is presented in the dissertation. The diagram of this method is shown in Figure 3.27. In the figure, the HFI and VM work together with two crossover function (XF) blocks. One XF block (XF_HFI) is for HFI and the other one (XF_VM) is for voltage model. When the rotor speed is lower than $\omega_{r,HFI_{XF}}$, XF_HFI’s coefficient should be 1.0 so that HFI’s rotor position can pass through the XF_HFI block without any change. In the meanwhile, the XF_VM’s coefficient should be 0.0 to disable the VM rotor position. When the rotor speed is between $\omega_{r,HFI_{XF}}$ and $\omega_{r,VM_{XF}}$, XF_HFI’s coefficient will decrease from 1.0 to 0.0 and XF_VM’s coefficient will increase from 0.0 to 1.0. The decreasing ratio and increasing ration are the same. When the rotor speed is equal and greater than $\omega_{r,VM_{XF}}$, the XF_HFI’s coefficient is 0.0 and XF_VM’s is 1.0. After $\omega_{r,VM_{XF}}$, the HFI rotor position is disabled completely. Consequently, the crossover functions (XF_HFI and XF_VM) guarantee the sensorless transition between two different sensorless methods.
Figure 3.27: Full speed range sensorless control with crossover function blocks
CHAPTER 4

SENSORLESS CONTROL FOR INDUCTION MACHINE BASED INTEGRATED STARTER GENERATOR

4.1 IM ISG System Hardware Design

Figure 4.1 shows the overall configuration of an ISG system, containing three major parts: a power converter, an AC machine and a DSP controller [46]. The constraint of the system design is that the DC bus voltage is in the range of 12-30 volts, not allowed to change [20].

The power circuit for the proposed ISG system is a three phase voltage source inverter (VSI) using power MOSFETs in parallel. Feedback signals such as currents and voltage are acquired and used by the DSP after the signal conditioning circuits. Based on the feedback information, advanced control algorithms embedded in DSP command the space vector PWM that in turn gates the power inverter for either starter or generator operation.

The specifications for the intended ISG inverter are listed in Table 4.1. According to Table 4.1, the ISG inverter is connected to a low DC bus voltage and it should have high AC current capability. The power switches should have very low on resistance for less $i^2R$ power loss. After many careful comparison studies on several types power MOSFETs, International Rectifier’s IRFP2907 with TO247 package is chosen for its best ratio.
Figure 4.1: The ISG overall structure
<table>
<thead>
<tr>
<th>DC voltage</th>
<th>12 – 30V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak torque at starting</td>
<td>41Nm</td>
</tr>
<tr>
<td>Continuous power for generating</td>
<td>10kW</td>
</tr>
<tr>
<td>Maximum current in starting</td>
<td>1500A&lt;sub&gt;peak&lt;/sub&gt;</td>
</tr>
<tr>
<td>Continuous current in generating</td>
<td>400A&lt;sub&gt;peak&lt;/sub&gt;</td>
</tr>
<tr>
<td>Cooling</td>
<td>Forced air</td>
</tr>
<tr>
<td>Dimensions (in. x in. x in.)</td>
<td>10x10x6.0</td>
</tr>
</tbody>
</table>

Table 4.1: ISG system specifications

of performance to cost. Specifically, IRFP2907 has the ratings of 75V, 209A<sub>peak</sub>, 175°C junction temperature and ultra low on-state resistance (4.5mΩ) [55]. The improved repetitive avalanche rating and the high switching frequency enable the kilo-amp inverter highly efficient and reliable. A 3-D drawing of the inverter is shown in Figure 4.2. The 36 power MOSFET devices are divided into three phases. Six IRFP2907 in each switch and twelve in each phase for a compact package. The intrinsic anti-paralleled body diode has the similar current and voltage ratings within the power MOSFET. No additional free-wheeling diode is needed. Thus, much space and cost is saved.

Passive components, especially the DC bus capacitors, are of critical importance for ISG inverter design. Switch mode power supply (SMPS) multilayer ceramic (MLC) capacitors with extremely low equivalent series inductance (ESL) and low (ESR) can offer the required performance for high frequency switching. In addition, SMPS MLC capacitors have very high ripple current capabilities, thermal stability and reliability, and compact
size. All capacitors are placed physically very close to power MOSFETs to minimize the parasitic inductance.

Thermal design is optimized to guarantee safe operation of the ISG system. The thermal design consideration has focused on reducing the overall system thermal resistance by enhancing contact between the device case and the heat sink surface. In addition, forced air cooling with sufficient air flow rate is also emphasized to dramatically reduce the thermal resistance by 1/5-1/15. In the prototype, a fan provides forced-air cooling with an air flow rate of 300\(m^3/hour\).

A prototype ISG inverter was built and tested. Figure 4.3 shows the inverter prototype with design specifications. The phase current capability is 1400A. More details for the inverter are described in [46].

Figure 4.2: 3-D drawing of the inverter
Figure 4.3: Prototype of the 10kW inverter
The IM ISG’s full speed range torque characteristics is defined in Figure 4.4. The load torque and required torque are shown in the figure. The IM ISG design should meet the torque requirement for the starting mode with sensorless control algorithm. The parameters of the prototype induction machine are shown in Table 4.2. Torque curves with different slip frequencies are calculated with parameters shown in Table 4.2. Likewise, the IM ISG should produce rated power output in the generating mode with sensorless control algorithm. The prototype induction machine is shown in Figure 4.5.

Figure 4.4: Induction machine ISG torque-speed characteristics
Figure 4.5: Prototype of induction machine in the 10kW IM ISG system

<table>
<thead>
<tr>
<th>IM ISG Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_s$</td>
<td>2.8 mΩ</td>
</tr>
<tr>
<td>$R_r$</td>
<td>1.6 mΩ</td>
</tr>
<tr>
<td>$L_{ls}$</td>
<td>5.3 μH</td>
</tr>
<tr>
<td>$L_{lr}$</td>
<td>5.3 μH</td>
</tr>
<tr>
<td>$L_m$</td>
<td>210 μH</td>
</tr>
</tbody>
</table>

Table 4.2: Parameters of the induction machine
4.2 Transformation Between Y-Connection And Delta-Connection

Why use Delta-connection? For the IM ISG system, the designer should consider two different modes: starting mode and generating mode. For starting mode, the torque produced by both Y-connection and Delta-connection may meet the torque requirements with a fixed low DC bus voltage. However, in the generating mode (7200rpm-12000rpm), the induction machine is working in the deep flux weakening region (constant power region). From torque-speed curve, the Y-connection system might not provide 10kW power in this region with the same DC bus voltage. In order to increase the power level with the fixed DC bus voltage, Delta-connection can be used to increase phase voltage $\sqrt{3}$ times larger with the fixed DC bus voltage. So, the power level will be increased accordingly. After using Delta-connection, the system meets the power requirements in the generating mode.

A benefit for the Delta-connection is that at starting mode, the engine can be started with lower DC bus voltage to meet starting mode torque requirements. The disadvantage for the ISG system is that the AC side starting currents will increase $\sqrt{3}$ times comparing with Y-connection. This incurs many design concerns, such as more paralleled power devices, higher thermal stress, stronger cooling system, and so on.

Note that in most drive systems, Y-connection is used. For a Delta-connection system, some modifications should be made. The basic rule of the modification is per phase equivalent. For both Y-connection and Delta-connection, the plants focus on per phase variables. So, based on this rule, the structure of the induction machine drive system should be changed in several aspects, such as: (1) magnitude and phase between line currents/voltages and phase current/voltages; (2) rotation angles for Park and inverse Park transformations; (3) rescale the starting mode parameters with lower DC bus voltage, and so on.
For the starting mode with Delta-connection, the parameters should be rescaled accordingly. For example, the DC bus voltage may be too high for Delta-connection in starting mode. Only half DC bus voltage is needed for the Delta-connection. The phase voltages and phase currents can meet the torque requirements. However, the DC bus cannot be changed during the starting mode and generating mode. In order to get required phase voltages and currents with full DC bus voltage \( V_{DC} \), the equivalent half DC bus voltage \( V_{DC_{half}} \) is applied in the starting mode. After parameters rescaling, the modulation index should be rescaled to 0.5 to get equivalent half DC bus voltage. Through these modifications, the ISG system can start the engine with required torque with equivalent half DC bus voltage and generate rated power with full DC bus voltage.

For the Delta-connection induction machine drive system, three parts should be paid attention to:

1. Additional Inverse Park transformation

Because all variables are based on per phase control in the induction machine drive system, the output of inverse Park transformation \( V_{\alpha 0} \) and \( V_{\beta 0} \) should be rotated 30 degrees to match the actual \( V_\alpha \) and \( V_\beta \). The relationship between \( V_{\alpha 0} \) and \( V_{\beta 0} \) and \( V_\alpha \) and \( V_\beta \) is shown in Figure 4.6. The additional \(-30^\circ\) inverse Park transformation is needed as follows:

\[
\begin{align*}
\begin{bmatrix} V_{\alpha} \\ V_{\beta} \end{bmatrix} &= \begin{bmatrix} \cos(-30^\circ) & -\sin(-30^\circ) \\ \sin(-30^\circ) & \cos(-30^\circ) \end{bmatrix} \begin{bmatrix} V_{\alpha 0} \\ V_{\beta 0} \end{bmatrix} \\
&= \begin{bmatrix} \cos(30^\circ) & \sin(30^\circ) \\ -\sin(30^\circ) & \cos(30^\circ) \end{bmatrix} \begin{bmatrix} V_{\alpha 0} \\ V_{\beta 0} \end{bmatrix}
\end{align*}
\]

(4.1)

2. Additional Park transformation

The line-line currents are acquired from ADC channels. They should be transformed to phase currents. The relationship between line-line currents and phase currents is shown in Figure 4.7. The transformation is:
Figure 4.6: The relationship between $V_{\alpha 0}$ and $V_{\beta 0}$ and $V_{\alpha}$ and $V_{\beta}$

\[
\begin{bmatrix}
i_{\alpha n} \\ i_{\beta n}
\end{bmatrix} = \frac{1}{\sqrt{3}} \begin{bmatrix}
\cos(30^\circ) & -\sin(30^\circ) \\
\sin(30^\circ) & \cos(30^\circ)
\end{bmatrix} \begin{bmatrix}
i_{\alpha,l-1} \\ i_{\beta,l-1}
\end{bmatrix}
\] (4.2)

(3) Phase voltage calculation based on SVPWM

Phase voltage can be derived with measured DC bus voltage and three upper switching functions $T_a$, $T_b$ and $T_c$ shown in Figure 4.8. First, the inverter output voltage $V_a$, $V_b$ and $V_c$ are expressed as follows:

\[
V_a = T_a V_{DC}
\] (4.3)

\[
V_b = T_b V_{DC}
\] (4.4)

\[
V_c = T_c V_{DC}
\] (4.5)

Then, the phase voltages are calculated as follows:

\[
V_{a,ph} = V_a - V_b = (T_a - T_b)V_{DC}
\] (4.6)

\[
V_{b,ph} = V_b - V_c = (T_b - T_c)V_{DC}
\] (4.7)
Figure 4.7: The relationship between $i_{\alpha_{-l-l}}$ and $i_{\beta_{-l-l}}$ and $i_{\alpha_n}$ and $i_{\beta_n}$

Figure 4.8: Phase voltage calculation
\[ V_{c,ph} = V_c - V_a = (T_c - T_a)V_{DC} \]  

(4.8)

So, the matrix formation is:

\[
\begin{bmatrix}
V_{a,ph} \\
V_{b,ph} \\
V_{c,ph}
\end{bmatrix} =
\begin{bmatrix}
1 & -1 & 0 \\
0 & 1 & -1 \\
-1 & 0 & 1
\end{bmatrix}
\begin{bmatrix}
T_aV_{DC} \\
T_bV_{DC} \\
T_cV_{DC}
\end{bmatrix}
\]

(4.9)

Then, the three phase voltages are converted to phase voltage \( V_{\alpha,ph,clac} \) and \( V_{\beta,ph,clac} \) in the stationary reference frame.

\[
\begin{bmatrix}
V_{\alpha,ph,clac} \\
V_{\beta,ph,clac}
\end{bmatrix} =
\begin{bmatrix}
1 & 0 \\
\frac{1}{\sqrt{3}} & \frac{2}{\sqrt{3}}
\end{bmatrix}
\begin{bmatrix}
V_{a,ph} \\
V_{b,ph}
\end{bmatrix}
\]

(4.10)

After above three modifications, the basic Delta-connection system diagram is shown in Figure 4.9. The final IM ISG system is based on this Delta-connection system. It works very well in both starting mode and generating mode.
Figure 4.9: The basic Delta-connection system diagram
4.3 Sensorless Control Algorithm for Starting Mode

Field orientation control ensures that the torque and flux is controlled independently. The voltage model flux observer will be discussed in this section for sensorless FOC induction machine drives. The IM model in stationary reference frame could be expressed as:

\[
\begin{bmatrix}
  i_{\alpha s} \\
  i_{\beta s} \\
  \lambda_{\alpha r} \\
  \lambda_{\beta r}
\end{bmatrix} =
\begin{bmatrix}
  -\frac{1}{\sigma L_s}(R_s + \frac{L_r^2}{L_m L_r}) & 0 & \frac{1}{\sigma L_m} \frac{L_m}{L_r} & \omega_r \frac{1}{\sigma L_s} \frac{L_m}{L_r} \\
  0 & -\frac{1}{\sigma L_s}(R_s + \frac{L_r^2}{L_m L_r}) & -\omega_r \frac{1}{\sigma L_s} \frac{L_m}{L_r} & \frac{1}{\sigma L_m} \frac{L_m}{L_r} \\
  \frac{L_m}{L_r} & 0 & -\frac{1}{L_r} & -\omega_r \\
  0 & \frac{L_m}{L_r} & \omega_r & -\frac{1}{L_r}
\end{bmatrix}
\begin{bmatrix}
  i_{\alpha s} \\
  i_{\beta s} \\
  \lambda_{\alpha r} \\
  \lambda_{\beta r}
\end{bmatrix} + \frac{1}{\sigma L_s}
\begin{bmatrix}
  v_{\alpha s} \\
  v_{\beta s} \\
  0 \\
  0
\end{bmatrix}
\]

where \( \sigma = 1 - \frac{L_m^2}{L_r L_s} \) and \( T_r = \frac{L_r}{K_r} \); subscripts \( \alpha \) and \( \beta \) represent stationary reference frame.

The stator flux linkages in the voltage model are computed by means of Back-EMF’s integration.

\[
\hat{\lambda}_{\alpha s} = \int (u_{\alpha s} - i_{\alpha s} R_s)dt \tag{4.12}
\]

\[
\hat{\lambda}_{\beta s} = \int (u_{\beta s} - i_{\beta s} R_s)dt \tag{4.13}
\]

Where \( R_s \) is the stator resistance, \( u_{\alpha s} \) and \( u_{\beta s} \) are stationary \( \alpha \beta \)-axis voltages. Then, the rotor flux linkages based on the voltage model are further computed.

\[
\hat{\lambda}_{\alpha r} = -\left(\frac{L_s L_r - L_m^2}{L_m}\right)i_{\alpha s} + \frac{L_r}{L_m} \hat{\lambda}_{\alpha s} \tag{4.14}
\]
The rotor position based on the voltage model is finally estimated as

$$\hat{\theta}_r = \arctan\left(\frac{\hat{\lambda}_{\beta r}}{\hat{\lambda}_{\alpha r}}\right)$$  \hspace{1cm} (4.16)$$

In the IM ISG system, the rotor position is estimated based on the above sensorless estimation algorithm. In general, the whole block diagram for rotor position sensorless algorithm is shown in Figure 4.10.

---

**Figure 4.10: Block diagram for rotor position sensorless algorithm**

---

### 4.4 Sensorless Control Algorithm for Generating Mode

Because the IM’s rotor shaft is coupled rigidly with the test stand, the rotor speed cannot be changed by the IM ISG itself. For example, when the load increases, the torque
will increase accordingly at 12000rpm. This also results in the increase of the slip at 12000rpm. It means that the applied stator flux’s speed should be decreased and be lower than 12000rpm. To keep the IM ISG works within the high efficient and stable zone, the maximum slip should be no more than the pullout slip of IM. The different operating curves in generating mode at 12000rpm are shown in Figure 4.11. The cross point of each curve with different load and 12000rpm dash line means that the ISG works at this operating point with corresponding load. So the operating point will always move on the 12000rpm dash line with different loads.

![Figure 4.11: ISG characteristics](image)

It is very different between sensorless control and sensor control in generating mode. Methods used in conventional sensor control may be not proper for sensorless control.
Here, three methods are implemented and tested with the sensorless control algorithm in generating mode.

4.4.1 Negative $I_{q\_ref}$ Command

The conventional system block diagram for the sensorless control of induction machine in flux weakening region is shown in Figure 4.12. In starting mode, the machine first accelerates from 0rpm to 5000rpm. The ISG system enters the generating mode when the machine speed is above 7200rpm. $I_{d\_ref}$ command is generated from the flux weakening profile block according to the rotor speed. $I_{q\_ref}$ is positive when the ISG is in starting mode. The $I_{q}$ PI controller will saturate when the ISG system enters the flux weakening region.

To investigate the two current PI controllers’ behavior in generating mode, $I_{d\_ref}$ should follow flux weakening profile and $I_{q\_ref}$ should change its value from a positive number to a negative number according to the induction machine’s torque equations (Equation 4.17 and Equation 4.18).

\[
T_e = \frac{3}{2} P (\psi_{ds} i_{qs} - \psi_{qs} i_{ds})
\]  
(4.17)

\[
\psi_{qs} = 0 \text{ for field orientation control, then}
T_e = \frac{3}{2} P (\psi_{ds} i_{qs})
\]  
(4.18)

Simulation results are shown in following Figure 4.13 through Figure 4.17. In Figure 4.13, $T_e$ is induction machine’s electromagnetic torque. $T_{load}$ is the test stand profile. Note that the machine speed is set at 7200rpm for all the simulation investigations presented in this document. Here, simulated $T_e$ is less than 0, showing the ISG system is
Figure 4.12: The conventional sensorless control of induction machine in flux weakening region
in generating mode. However, it is not true to say that the system is in generating mode according to the negative $T_e$.

The references and feedbacks feeding to two PI controllers are shown in Figure 4.14. The details of the transition waveforms between starting and generating are expended in the lower trace of Figure 4.14. It is observed that the $I_{q_{-fdb}}$ becomes negative, following $I_{q_{-ref}}$ very well. The two PI controllers’ outputs $V_d$ and $V_q$ are displayed in Figure 4.15. The lower trace is the expended simulation result after $-I_{q_{-ref}}$ is applied. $V_q$ saturates at higher speeds in starting mode. When $I_{q_{-ref}}$ becomes negative, because $|I_{q_{-ref}}|$ is greater than $|I_{q_{-fdb}}|$, PI controller’s output is decreasing from PI controller’s upper limit to negative value. When $I_{q_{-fdb}} = I_{q_{-ref}}$, the ISG system is in stable operation.

The estimated rotor flux angle is shown in Figure 4.16. The lower trace is the expended simulation result after $-I_{q_{-ref}}$ is applied. The estimated rotor flux angle direction is reversed when $V_q$ becomes negative as shown in the expanded simulation result. It means that the estimated rotor angle’s direction is different from the test stand’s rotating direction. That also means the control system gives the ISG a wrong command in direction. Obviously, the ISG is running into a wrong region. So phase currents shown in Figure 4.17 will be dissipated in the rotor’s circuit.

In short, all above simulation results tell us the ISG system cannot enter generating mode from starting mode by just changing $I_{q_{-ref}}$ from positive value to negative with sensorless vector control. The reason is the output of the $I_q$ PI controller becomes negative during the transition. Theoretically, $V_q$ should be changed a little in generating mode. Additionally, the synchronous speed is lower than the rotor speed. It is PI controller’s inherent behavior. Therefore the ISG system can not enter generating mode with this method.
The experimental results are shown in Figure 4.18. The estimated rotor flux angle \( \theta_{est} \)'s direction is reversed after the \(-I_{q,ref}\) command is applied. The system does not work correctly. \( \theta_{est} \) and \( I_{DC} \) are wrong so that the system is tripped after a while. Nevertheless, the system is still in starting mode rather than generating mode. It seems that the generating mode cannot be implemented by this method.

![Graph showing Torque (Nm) and Rotor Speed](image)

**Figure 4.13:** \( T_e \), \( T_{load} \) and rotor speed

### 4.4.2 \( I_{d,ref} \) Command Reduction in Generating Mode

The system block diagram of commanding \( I_{d,ref} \) instead of \( I_{q,ref} \) for sensorless control of induction machine in flux weakening region is shown in Figure 4.19. In the starting mode, the \( I^*_{d,ref} \) command is given by the flux weakening profile block according to the rotor speed. \( I_{d,ref,G} \) is regulated by DC bus voltage PI controller when the ISG is in the generating mode. The simulation results are shown in Figure 4.20 through Figure 4.25.
Figure 4.14: $I_d$ and $I_q$

Figure 4.15: $V_d$ and $V_q$
Figure 4.16: $\theta_{est}$

Figure 4.17: Phase currents $I_a, I_b, I_c$
Figure 4.18: The experimental results of $I_{DC}$ (CH3) and $\theta_{est}$ (CH4) with $-I_{q\_ref}$ command.
Figure 4.19: The system block diagram with commanding $I_{d_{\text{ref}}}$
In Figure 4.20, $T_e$ is the induction machine’s electromagnetic torque. $T_{load}$ is the test stand profile. As shown in the figure, the ISG is in generating mode when rotor speed is larger than 7200 rpm. However, $T_e$ is still positive in the "generating" mode. Actually, the ISG has not entered the generating mode yet. The references and feedbacks of the two PI controllers are shown in Figure 4.21. The transition waveforms of $I_{d\_ref}$ and $I_{d\_fdb}$ between starting and generating are expended in Figure 4.22. It is observed that the $I_{d\_fdb}$ can not follow $I_{d\_ref}$.

The two PI controllers’ outputs $V_d$ and $V_q$ are displayed in Figure 4.23. $V_q$ saturates at higher speeds in starting mode. $V_d$ also saturates when the ISG system enters the generating mode.

The estimated rotor flux angle is shown in Figure 4.24. The estimated rotor flux angle’s frequency is lower in generating mode than in starting mode. The controller is trying to command the ISG system to enter the generating mode. This method wants to adjust the $I_{d\_ref}$’s value to change the rotor speed. However, in the real system, the load is a prime mover. The rotor speed cannot be changed by IM ISG itself.

In general, from above simulation results, the ISG system cannot enter generating mode from starting mode by just changing $I_{d\_ref}$ command in sensorless vector control.

### 4.4.3 Slip Control in Generating Mode

The system block diagram of slip control for the sensorless control of induction machine in flux weakening region is shown in Figure 4.26. In the generating mode, AC phase voltages reach their limits because of the limited DC bus voltage. The maximum AC phase voltage is $\frac{2}{\pi} V_{DC}$. Therefore, in generating mode, only slip can be controlled through DC bus voltage PI controller. The DC bus voltage PI controller’s output is the slip speed. The
Figure 4.20: $T_e$, $T_{load}$ and rotor speed

Figure 4.21: $I_{d_{ref}}, I_{d_{fdb}}, I_{q_{ref}}$ and $I_{q_{fdb}}$
Figure 4.22: $I_{d_{ref}}$ and $I_{d_{fdb}}$

Figure 4.23: $V_d$ and $V_q$
Figure 4.24: Rotor angle

Figure 4.25: Phase currents $I_a, I_b, I_c$
system synchronous speed is the sum of the slip speed and the estimated rotor speed. As is known, all transformations depend on the synchronous angle for the field orientation control. In order to obtain synchronous flux rotating angle, the synchronous speed should be integrated. In the generating mode, the controller maintains the constant DC bus voltage through adjusting the slip speed.

Figure 4.26: Indirect field orientation control through slip control in the flux weakening region

Simulation results are shown in Figure 4.27 to Figure 4.31. In Figure 4.27, \( T_e \) is induction machine’s electromagnetic torque. It is observed that \( T_e \) is less than 0, showing that the ISG system is in generating mode.
Figure 4.27: $T_e$

Figure 4.28: DC bus current
Figure 4.29: The estimated rotor flux angle

Figure 4.30: The DC bus voltage reference and feedback
The estimated rotor flux angle is shown in Figure 4.29. The estimated rotor flux angle’s frequency is lower in generating mode than in starting mode at a same rotor speed. The controller is trying to command the ISG system to enter generating mode. By this method, the slip speed will be adjusted to change the rotor speed. In the real system, the load is a stiff prime mover making the ISG’s rotor speed a constant. The slip is negative if the ISG system’s synchronous speed is lower than the rotor speed. In addition, the DC bus current is negative, showing the ISG system is entering the generating mode. In general, from all above simulation results, the ISG system can enter generating mode from starting mode by adjusting the slip speed in sensorless vector control.

In order to verify the proposed sensorless control algorithm, the IM ISG was tested in the generating mode with this method. When IM ISG enters the generating mode, its velocity will be from 7200rpm to 12000rpm. During this period, the IM ISG will produce
Figure 4.32: Torque and speed waveforms in generating mode

Vdc = 30.0V, Idc = 312A - 44A - 312A
Speed: 7200rpm - 12000 rpm - 7200 rpm

-2.37 ft/lbs at 7200 with no load
-13.4 ft/lbs at 7200 rpm with load
-9.45 ft/lbs at 12000 rpm with load
Figure 4.33: Load testing waveforms for $V_{DC}$ (CH2), $I_{DC}$ (CH3: 100A/div) and $I_{phase}$ (CH4: 400A/div)

Figure 4.34: Load change from full to zero: $V_{DC}$ (CH2), $I_{DC}$ (CH3: 100A/div) and $I_{phase}$ (CH4: 400A/div)
Figure 4.35: Load change from zero to full: $V_{DC}$ (CH2), $I_{DC}$ (CH3: 100A/div) and $I_{phase}$ (CH4: 400A/div)

proper DC power output according to different DC side loads. In the meanwhile, the DC bus voltage should maintain constant 30V. The torque and speed experimental results recorded are displayed in Figure 4.32.

When the test stand’s speed is 7200rpm, the IM ISG enters the generating mode with sensorless control while the resistance load is switched to IM ISG’s DC bus. Then the test stand will run from 7200rpm to 12000rpm then back to 7200rpm with the DC side load and the DC bus voltage should be constant. In order to test the dynamic characteristics of the IM ISG system, different DC resistance loads are applied to DC bus at the end of the testing. The torque dynamic characteristics with and without load is shown in Figure 4.32. The DC bus voltage, DC bus current and AC line current for the dynamic testing are shown in Figure 4.33. There are two load changes: First load change is from full load to no
load; Second load change is from no load to full load. Figure 4.34 and Figure 4.35 are the detailed dynamic waveforms for the two load changes. The DC bus voltage could jump to 34.3V during the transition when load from full (312A) to zero (44A). After the transition, the DC bus voltage is back to 30V. The DC bus voltage, DC bus current and AC line current waveforms are shown in Figure 4.34. The DC bus voltage could drop to 13.2V during the transition when load from zero (44A) to full (312A). After the transition, the DC bus voltage is back to 30V. The DC bus voltage, DC bus current and AC line current waveforms are shown in Figure 4.35. All above performances meet the aircraft electric power characteristics standard [20].

From above experimental results, the ISG system can enter generating mode and the DC bus voltage is controllable in the testing by this method. It can be concluded that the slip control is a feasible control method for generating mode in flux weakening region.

4.5 Induction Machine ISG Torque-Speed Characteristics

To test IM ISG’s performance with real characteristics of a turbine, a test stand is set up to simulate the turbine engine characteristics. A 50HP test stand is shown in Figure 4.36. The ISG machine is coupled to the test stand shaft. There are one torque transducer and one speed transducer on its shaft. When the test stand is working, it records shaft torque, load torque and shaft speed information. Meanwhile, there is a DC machine in the test stand to simulate the characteristics of the turbine.

The details of the original DC machine starter torque-speed characteristics [56] are shown in Figure 4.37. The load torque and required torque in starting mode are drawn in the figure. The details of the induction machine ISG torque-speed characteristics are shown in Figure 4.38. The load torque and required torque are drawn in the figure.
During the starting mode, the test stand simulates a load torque curve according to shaft speed from 0 rpm to 5000 rpm. The starting mode load torque curve and required electromagnetic torque curve are shown in Figure 4.38. From the figure, the load torque is not constant. So there are several different work conditions in the whole process:

(1) At the beginning of the starting, the torque is 10 ft – lb with reverse direction rotation (−50 rpm). This torque is called windage of the engine. In this interval, the starter should provide 30 ft – lb to meet the required torque curve.

(2) Between −50 rpm and 1400 rpm, the load torque will increase from 10 ft – lb to 21 ft – lb. The starter will decrease from 30 ft – lb to 23 ft – lb.
Figure 4.37: Induction machine ISG torque-speed characteristics [56]
Figure 4.38: Induction machine ISG torque-speed characteristics
(3) When speed reaches 1400rpm, the load torque also reaches its maximum torque 21ft – lb. After this point, the load torque decreases quickly. Meanwhile, the starter’s torque also decreases according to load torque changes.

(4) When the speed reaches 2100rpm, the load torque is 0ft – lb. It means that the turbine is light now. After this point, the turbine is started. The starter’s torque continue to decreasing.

(5) However, the turbine still need the starter’s assistant until the engine’s speed is higher than 5000rpm. During this interval, the turbine produces negative load torque. The starter still provides positive torque in order to assist the turbine’s acceleration.

(6) At 5000rpm, the turbine is totally started by the starter.

(7) After 5000rpm, the starter is disabled and the turbine will run as its characteristic. The starting mode is completed. At 6000rpm, the turbine’s torque reaches 20ft – lb.

During the generating mode, the test stand simulates the turbine as a prime move with constant torque characteristics 7200rpm to 12000rpm shown in Figure 4.38. There are different work conditions in the generating process:

1. From 5000rpm to 7200rpm, the ISG system is in the idle state (No action). The turbine accelerates with its constant acceleration rate.

2. When the system reaches 7200rpm, the ISG system enters the generating mode. The engine will continue to accelerate from 7200rpm to 12000rpm. The ISG will work as a generator in this interval. The generator will provide negative torque for the generating mode and produce a constant 30V DC bus voltage and corresponding DC current according to DC side load.
(3) When the engine’s speed is 12000rpm, the engine works at its rated speed with normal condition. The generator also works in its rated condition: 30V DC bus voltage, 350A DC output with full load.

4.6 Induction Machine Performance Testing Results

(1) Starting mode with Y-connection

The starting mode with Y-connection were tested first with full DC bus voltage. The phase current of Y-connection is smaller than that of Delta-connection. Figure 4.39 shows the full load starting performance with Y-connection. It depicts that the torque produced by starter is met the required torque curve during the starting process.

(2) Generating mode with Y-connection

Because the full DC bus voltage is used in starting mode, it can not provide enough excitation voltage for the generating in higher speeds with Y-connection. The generator can not produce rated power with Y-connection with the 30V DC bus voltage. As aforementioned, a Delta-connected generator can solve this problem.

(3) Starting mode with Delta-connection

In order to overcome limitation of the Y-connection in the generating mode, the ISG system should be connected with Delta-connection. However, the line current is $\sqrt{3}$ times higher than that with the Y-connection. The high current capability of the power converter is limited by original thermal design. The testing result is shown in Figure 4.39. Only 75% load torque is applied to test the starting performance of the starter.

(4) Generating mode with Delta-connection

With Delta-connection, the generator can work at 91.4% rated power condition with 30V DC bus and 320A DC current generated shown in Figure 4.41. If the DC bus voltage
Figure 4.39: Induction machine ISG starting performance with Y-connection
Figure 4.40: Induction machine ISG starting performance with Delta-connection
is increased to 32V, the IM can generate rated current at 350A. This indicates that the problem is caused by machine design. In the second round machine design, the machine’s parameters are refined to provide the rated power output.

Figure 4.41: Induction machine ISG generating performance with Delta-connection
CHAPTER 5

CONCLUSIONS AND FUTURE WORK

5.1 Conclusions

High frequency current responses of a PM synchronous machine with and without considering the eddy current effects are considerably different. This dissertation identifies and investigates the eddy current effects on rotor position estimation and contributes to an improved high frequency injection based sensorless control of a PM synchronous machine at zero and low speeds. Both computer simulation and experimental results have proven that the proposed high frequency injection method with compensation works very well. It should also be pointed that for a conventional salient-pole synchronous machine, the proposed rotor position estimation and sensorless control approach is valid. This is because the currents in the damper windings of a conventional salient pole synchronous machine function identically as those eddy currents during the HFI mode.

In addition, an improved magnetic pole identification method based on space vector PWM for arbitrary initial rotor position estimation and sensorless control of PM synchronous machines is presented. Through applying vector controlled pilot voltages by SVPWM, the N-S pole can be identified at any initial rotor positions without rotor alignment actions. The proposed method can be combined with the conventional initial rotor
position and sensorless control scheme to ensure the effective estimation of initial magnetic pole position. The validity of the proposed identification method has been proven through experimental results based on a 5HP salient pole synchronous machine drive system.

In this dissertation, a voltage model based rotor position observer for induction machine ISG is presented and used for test validation. The method works well at very low frequency regions where full torque is required. Apparently, the voltage model flux observer can satisfy wide speed range ISG applications including zero speed starting. Test results with full starting torque over a wide speed range are presented based on the proposed rotor position observer.

A sensorless control structure in deep flux weakening region for an ISG system is also presented in this dissertation. AC phase voltages reach their limits in the generating mode because of the limited DC bus voltage. Only slip can be controlled through DC bus voltage PI controller. Accordingly, the controller maintains the constant DC bus voltage through adjusting the slip speed. The proposed methods can ensure the effective estimation of rotor position in the starting mode and generating mode. The validity of the proposed methods has been proven through experimental results based on a 10KW induction machine drive system.

### 5.2 Future Work

All test waveforms by the sensorless control of the PM synchronous machine are based on light load. The full load performance should also be tested to validate HFI and VM sensorless control algorithm for PM synchronous machine. In addition, the dynamic performance should be evaluated at low speed with HFI and high speed with VM. This algorithm also can be applied to other different AC machines whose rotors have an identified saliency,
such as interior permanent machines (IPM), induction machines with rotor saliency, and so on.

The IM ISG system should have an overload capability in generating mode. However, the existed design can not provide sufficient overload capability. Capacitance compensation could be considered to solve the overload capability in generating mode. Unfortunately, additional machine windings and external power converter may be needed for the capacitance compensation method.

Due to inherent conflicts imposed by low-speed high-torque in starting mode and high speed high efficiency in generating mode for the ISG system, it is much preferable to have a flexible DC voltage fed to the DC-AC conversion. To achieve an optimal ISG system by designing AC machine alone is almost impossible. In [57], three candidates of combined DC-DC-AC inverters are investigated for optimal overall ISG systems. The final answers to the optimized ISG system largely depend on how complicated and expensive the combined DC-DC-AC converter is. In general, there are several advantages with DC-DC-AC converter. (1) The ISG machine design will be flexible; (2) the DC and AC sides currents will dramatically dropped to a very lower range. (3) It is easier for the power converter thermal design.
The digital filters are very important parts in the HFI. The filter parameters and coefficients should be chosen carefully to avoid unnecessary phase shift.

1. Second Order High Pass Filter [58]

The digital domain form of the high pass filter is:

\[ H(z) = \frac{\alpha(1 - 2z^{-1} + z^{-2})}{\frac{1}{2} - \gamma z^{-1} + \beta z^{-2}} \]  

(A.1)

The coefficients are

\[ \beta = \frac{1 - \frac{d}{2} \sin(\theta_c)}{2[1 + \frac{d}{2} \sin(\theta_c)]}, \]

\[ \gamma = (\frac{1}{2} + \beta) \cos(\theta_c); \]

\[ \alpha = \frac{\frac{1}{2} + \beta + \gamma}{4}, \]

where \( \theta_c = \frac{2\pi f_c}{f_s}, \) \( d = 2. \)

\( d \) is the damping coefficients; \( \theta_c \) is the digital domain cutoff frequency. The equation of the second order digital low pass filter is shown in Equation A.2. This format is easily used for the DSP program with array or pointer technology. The digital implementation in DSP is shown in Figure A.1.

\[ y(n) = 2\alpha[x(n) - 2x(n-1) + x(n-2)] + \gamma y(n-1) - \beta y(n-2) \]  

(A.2)
2. Sixth Order Low Pass Filter [58]

The digital domain form of the Butterworth $k$th order low pass filter ($k/2$ second order filters cascaded) is:

$$H_k(z) = \frac{\alpha_k (1 + 2z^{-1} + z^{-2})}{\frac{1}{2} - \gamma_k z^{-1} + \beta_k z^{-2}}$$

(A.3)

The coefficients are

$$\beta_k = \frac{1 - d_k \sin(\theta_c)}{2[1 + \frac{d_k}{2} \sin(\theta_c)]},$$

$$\gamma_k = \left(\frac{1}{2} + \beta_k\right) \cos(\theta_c);$$

$$\alpha_k = \frac{\frac{1}{2} + \beta_k - \gamma_k}{4};$$

where $k = 1, 2, \ldots, \lfloor N/2 \rfloor$, $\theta_c = \frac{2\pi f_c}{f_s}$ and $d_k = 2 \sin\left(\frac{(2k-1)\pi}{2N}\right)$.

$d_k$ is the $k$th damping coefficient; $\theta_c$ is the digital domain cutoff frequency. The general equation of the $k$th order low pass filter is:

$$y_i(n) = 2\alpha_i[x_i(n) + 2x_i(n-1) + x_i(n-2)] + \gamma y_i(n-1) - \beta_i y_i(n-2)$$

(A.4)

where $x_{i+1}(n-k) = y_i(n-k), i = 1, 2, \ldots, k$
A sixth order low pass filter is used in HFI. It is a cascaded filter with three second order low pass filters. The digital implementation is shown in Figure A.2. The first second order low pass filter’s equation is:

$$y_1(n) = 2\alpha_1 [x_1(n) + 2x_1(n-1) + x_1(n-2)] + \gamma_1 y_1(n-1) - \beta_1 y_1(n-2) \quad (A.5)$$

where \( x_2(n-k) = y_1(n-k) \). It means that the output of the first LPF is the input of the second LPF. So, the input of the second LPF can be expressed with \( x_1 \) and \( x_2 \):

$$x_2(n) = 2\alpha_1 [x_1(n) + 2x_1(n-1) + x_1(n-2)] + \gamma_1 x_2(n-1) - \beta_1 x_2(n-2) \quad (A.6)$$

The second low pass filter is:

$$y_2(n) = 2\alpha_2 [x_2(n) + 2x_2(n-1) + x_2(n-2)] + \gamma_2 y_2(n-1) - \beta_2 y_2(n-2) \quad (A.7)$$

where \( x_3(n-k) = y_2(n-k) \). It means that the output of the second LPF is the input of the third LPF. So, the input of the third LPF can be expressed with \( x_2 \) and \( x_3 \):

$$x_3(n) = 2\alpha_2 [x_2(n) + 2x_2(n-1) + x_2(n-2)] + \gamma_2 x_3(n-1) - \beta_1 x_3(n-2) \quad (A.8)$$

The third low pass filter is:

$$y_3(n) = 2\alpha_3 [x_3(n) + 2x_3(n-1) + x_3(n-2)] + \gamma_3 y_3(n-1) - \beta_3 y_3(n-2) \quad (A.9)$$
where \( y(n - k) = y_3(n - k) \). It means that the output of the third LPF is the output of the sixth order low pass filter. So, the 6th order LPF’s output can be expressed with \( x_3 \) and \( y \):

\[
y(n) = 2\alpha_3 [x_3(n) + 2x_3(n - 1) + x_3(n - 2)] + \gamma_3 y(n - 1) - \beta_3 y(n - 2) \tag{A.10}
\]

For the DSP implementation, only three equations are used in the subroutine:

\[
x_2(n) = 2\alpha_1 [x_1(n) + 2x_1(n - 1) + x_1(n - 2)] + \gamma_1 x_2(n - 1) - \beta_1 x_2(n - 2) \tag{A.11}
\]

\[
x_3(n) = 2\alpha_2 [x_2(n) + 2x_2(n - 1) + x_2(n - 2)] + \gamma_2 x_3(n - 1) - \beta_2 x_3(n - 2) \tag{A.12}
\]

\[
y(n) = 2\alpha_3 [x_3(n) + 2x_3(n - 1) + x_3(n - 2)] + \gamma_3 y(n - 1) - \beta_3 y(n - 2) \tag{A.13}
\]
APPENDIX B

LIST OF ABBREVIATIONS

AC: Alternating Current
ADC: Analog to Digital Converter
A/D: Analog to Digital (Converter)
DC: Direct Current
DFOC: Direct Field Orientation Control
DSP: Digital Signal Processor
DTC: Direct Torque Control
EKF: Extended Kalman Filter
EMF: Electromagnetic Force
ESL: Equivalent Series Inductance
ESR: Equivalent Series Resistance
FOC: Field Orientation Control
GE: General Electric
HFI: High Frequency Injection
HP: Horsepower
HPF: High Pass Filter
ICE: Internal Combustion Engine
IFOC: Indirect Field Orientation Control
IM: Induction Machine
IPM: Interior Permanent Magnet Machine
ISG: Integrated Starter Generator
LPF: Low Pass Filter
MAC: Multiplier Accumulator Chip
MRAS: Model Reference Adaptive System
MCU: Micro Controller Unit
MIPS: Million Instructions Per Second
MLC: Multilayer Ceramic (Capacitor)
MOSFET: Metal Oxide Substrate Field Effect Transistor
PC: Personal Computer
PEEM: Power Electronics and Electric Machine Group
PI: Proportional-Integral (Controller)
PM: Permanent Magnet
PMSM: Permanent Magnet Synchronous Machine
PWM: Pulse Width Modulation
RPM: Revolutions Per Minute
SARAM: Signal Access Random Access Memory
SG: Starter Generator
SMO: Sliding Mode Observer
SMPM: Surface Mount Permanent Magnet Machine
SMPS: Switch Mode Power Supply
SVPWM: Space Vector Pulse Width Modulation
TI: Texas Instrument
VM: Voltage Model
VSI: Voltage Source Inverter
APPENDIX C

LIST OF SYMBOLS

\( \bar{e}'_{\alpha}(k), \bar{e}'_{\beta}(k) \): The discretized \( \bar{E}'_{\alpha}, \bar{E}'_{\beta} \)

\( \bar{e}'_{\alpha, pu}(k), \bar{e}'_{\beta, pu}(k) \): The normalized \( \bar{e}'_{\alpha}(k), \bar{e}'_{\beta}(k) \)

\( \bar{E}_s \): Back-EMF

\( \bar{E}'_s \): The vector that is in the same direction with \( \bar{E}_s \)

\( \bar{E}'_{\alpha}, \bar{E}'_{\beta} \): The \( \alpha \) and \( \beta \) axes vectors of \( \bar{E}'_s \)

\( f_{abc} \): Three phase variables

\( f_{an}, f_{bm}, f_{cn} \): Three phase variables

\( f_d, f_q \): Two phase variables in \( dq \) coordinates

\( f_{qd0s} \): \( qd0 \) variables in \( qd \) coordinates

\( f_{\alpha}, f_{\beta} \): Normalized two phase variables in \( \alpha\beta \) coordinates

\( f'_{\alpha}, f'_\beta \): Unnormalized two phase variables in \( \alpha\beta \) coordinates

\( i_{abc} \): Three phase stator current variables

\( i_{qd0s} \): Current variables in \( qd \) coordinates

\( i_{dqr} \): Rotor current variables in \( dq \) coordinates

\( i_{ds}, i_{qs} \): Stator current variables in \( dq \) coordinates

\( i_{\alpha n}, i_{\beta n} \): The phase currents in \( \alpha\beta \) coordinates

\( i_{\alpha j-1}, i_{\beta j-1} \): The line-line currents in \( \alpha\beta \) coordinates
\( i_{\alpha s}, i_{\beta s} \): Stator currents in \( \alpha \beta \) coordinates

\( i_{\alpha \beta \_HFI} \): Current vector after injecting high frequency voltage vector

\( i_{\alpha 1}, i_{\beta 1} \): The currents after first rotation with \( \omega t \) in HFI

\( i_{\alpha 2}, i_{\beta 2} \): The currents after HPFs in HFI

\( i_{\alpha 3}, i_{\beta 3} \): The currents after second rotation with \(-2\omega t \) in HFI

\( i_{\alpha \_dem}, i_{\beta \_dem} \): The currents after LPFs in HFI

\( i'_{dqr \_HFI} \): Undesired current vector considering eddy current effects in HFI

\( i'_{ar}, i'_{\beta r} \): Rotor currents that referred to stator in \( \alpha \beta \) coordinates

\( i'_{\alpha \beta \_HFI} \): Desired current vector considering eddy current effects in HFI

\( i'_{\alpha s}, i'_{\beta s} \): Stator currents in synchronous rotating reference frame

\( i'_{\alpha s}, i'_{\beta s} \): Stator currents in stationary rotating reference frame

\( i'_{ar}, i'_{\beta r} \): Rotor currents that referred to stator in synchronous rotating reference frame

\( i'_{ar}, i'_{\beta r} \): Rotor currents that referred to stator in stationary rotating reference frame

\( I_a, I_b, I_c \): Phase currents

\( I_{base} \): Base current for per-unit system

\( I_{DC} \): DC bus current

\( I_{d \_fdb}, I_{q \_fdb} \): Feedback currents in \( dq \) coordinates (The feedbacks of PI controllers)

\( I_{d \_ref}, I_{q \_ref} \): References currents in \( dq \) coordinates (The references of PI controllers)

\( I_{d \_ref \_G} \): \( d \) axis current regulated by DC bus voltage PI controller in generating mode

\( I^*_{d \_ref} \): \( d \) axis current command given by the flux weakening profile block according to the rotor speed

\( \bar{I}_d, \bar{I}_q \): The \( d \) and \( q \) axes current vectors

\( \bar{I}_s \): The stator terminal current vector

\( \bar{I}_\alpha, \bar{I}_\beta \): The \( \alpha \) and \( \beta \) axes current vectors of \( \bar{I}_s \)
\( J \): System inertia

\( k_1, k_2, k_3, k_4, k_5 \): Coefficients used for the three reference frames (Synchronous reference frame, stationary reference frame and arbitrary reference frame)

\( K_1, K_2, K_3 \): Per unit constants for the discrete time system

\( K_5 \): Transformation matrix from three phase variables to arbitrary reference frame

\( L_{dm}, L_{qm} \): \( d \) and \( q \) axes magnetizing inductances

\( L_{ds}, L_{qs} \): \( d \) and \( q \) axes stator inductances

\( L_{base} \): Base inductance for per-unit system

\( L_{ls} \): Leakage inductance

\( L_{ms} \): Magnetizing inductance

\( L_{sr} \): Mutual inductance matrix

\( L_s \): Stator inductance matrix

\( L_{skq}, L_{skd}, L_{sf} \): Mutual inductance between stator and rotor windings

\( L_s(2\theta) \): Stator inductance matrix for salient pole synchronous machine

\( L_{0s} \): Stator self inductance

\( L_{2s} \): Rotor position dependent inductance

\( P \): Number of pole pairs

\( P_{abc} \): The total instantaneous power for \( abc \) variables

\( P_{qd0} \): The total instantaneous power for \( qd0 \) variables

\( r_r \): Rotor resistance

\( r'_r \): Rotor resistance that referred to stator in synchronous rotating reference frame

\( r_s \): Stator resistance

\( R_{dr}, R_{qr} \): \( d \) and \( q \) axes rotor resistances

\( R_{rd} \): \( d \) axis excitation resistance
**Rs**: Stator resistance

**$T_{abc\rightarrow dq0}$**: Transformation from $abc$ coordinates to $dq0$ coordinates

**$T_{abc\rightarrow \alpha\beta0}$**: Transformation from $abc$ coordinates to $\alpha\beta0$ coordinates

**$T_{\alpha\beta0\rightarrow dq0}$**: Transformation from $\alpha\beta0$ coordinates to $dq0$ coordinates

**$T_a, T_b, T_c$**: Three phase upper switching functions

**$T_e$**: Electromagnetic torque

**$T_l$**: Load torque

**$T_s$**: Sampling time

**$v_{abcs}$**: Three phase stator voltage variables

**$v_{dq0}$**: Rotor voltage variables in $dq$ coordinates

**$v_{qd0s}$**: Voltage variables in $dq$ coordinates

**$v_{ds}, v_{qs}$**: Stator voltage variables in $dq$ coordinates

**$v_{\alpha s}, v_{\beta s}$**: Stator voltages in $\alpha\beta$ coordinates

**$v_{\alpha\beta_{HFI}}$**: Injected high frequency voltage vector

**$v'_{\alpha r}, v'_{\beta r}$**: Rotor voltages that referred to stator in arbitrary rotating reference frame

**$v^e_{\alpha s}, v^e_{\beta s}$**: Stator voltages in synchronous rotating reference frame

**$v^s_{\alpha s}, v^s_{\beta s}$**: Stator voltages in stationary rotating reference frame

**$v^e_{\alpha r}, v^e_{\beta r}$**: Rotor voltages that referred to stator in synchronous rotating reference frame

**$v^s_{\alpha r}, v^s_{\beta r}$**: Rotor voltages that referred to stator in stationary rotating reference frame

**$V_a, V_b, V_c$**: Calculated line voltages

**$V_{a_{ph}}, V_{b_{ph}}, V_{c_{ph}}$**: Calculated phase voltages

**$V_{base}$**: Base voltage for per-unit system

**$V_{DC}$**: DC bus voltage

**$V_{HFI}$**: The magnitude of the injected voltage vector for HFI
$V_{NS}, -V_{NS}$: Pilot voltages for magnetic pole identification

$V_{\alpha_0}, V_{\beta_0}$: The inputs of the additional inverse Park transformation

$V_{\alpha_{\text{ph}}}, V_{\beta_{\text{ph}}}$: Calculated phase voltages in $\alpha\beta$ coordinates

$\bar{V}_s$: The stator terminal voltage vector

$X_d, X_q$: The $d$ and $q$ axes reactances

$\theta_{dq}$: Rotating angle in $dq$ coordinates

$\theta_{est}$: Estimated rotor position

$\theta_{pu}(k)$: The normalized $\theta(k)$

$\theta_{qd}$: Rotating angle in $qd$ coordinates

$\theta_{r,HFI}$: Estimated rotor position with HFI

$\theta_{r_{-encoder}}$: The reference rotor angle of the encoder

$\theta(k)$: The discretized $\theta$

$\hat{\theta}_r$: Estimated rotor position

$\hat{\theta}_{r_{-VM}}$: Estimated rotor position based on voltage model

$\lambda_{abc\bar{s}}$: Three phase stator flux linkage variables

$\lambda_{d0q0}$: $dq0$ flux linkage variables

$\lambda_{dr}, \lambda_{qr}$: Rotor flux linkage variables in arbitrary reference frame

$\lambda_{ds}, \lambda_{qs}$: Stator flux linkage variables in arbitrary reference frame

$\lambda_{fr}$: Excitation flux linkage in arbitrary reference frame

$\lambda_{m}$: The magnitude of the rotor flux linkage

$\lambda_{\alpha s}, \lambda_{\beta s}$: Stator flux linkages in arbitrary rotating reference frame

$\lambda_{\alpha\beta_{-HFI}}$: Flux linkage vector after injecting high frequency voltage vector

$\lambda'_{\alpha r}, \lambda'_{\beta r}$: Rotor flux linkages that referred to stator in arbitrary rotating reference frame

$\lambda'_{\alpha s}, \lambda'_{\beta s}$: Stator flux linkages in synchronous rotating reference frame
\( \lambda_s^\alpha, \lambda_s^\beta \): Stator flux linkages in stationary rotating reference frame

\( \lambda_{te}^\alpha, \lambda_{te}^\beta \): Rotor flux linkages that referred to stator in synchronous rotating reference frame

\( \lambda_{sr}^\alpha, \lambda_{sr}^\beta \): Rotor flux linkages that referred to stator in stationary rotating reference frame

\( \bar{\lambda}_{\alpha r}, \bar{\lambda}_{\beta r} \): The \( d \) and \( q \) axes flux linkage vectors

\( \bar{\lambda}_{\alpha r}^\prime(k), \bar{\lambda}_{\beta r}^\prime(k) \): The discretized \( \bar{\lambda}_{\alpha r}, \bar{\lambda}_{\beta r} \)

\( \bar{\lambda}_{\alpha r,pu}(k), \bar{\lambda}_{\beta r,pu}(k) \): The normalized \( \bar{\lambda}_{\alpha r}^\prime(k), \bar{\lambda}_{\beta r}^\prime(k) \)

\( \hat{\lambda}_{\alpha r}, \hat{\lambda}_{\beta r} \): Estimated rotor flux linkages in \( \alpha \beta \) coordinates

\( \hat{\lambda}_{\alpha s}, \hat{\lambda}_{\beta s} \): Estimated stator flux linkages in \( \alpha \beta \) coordinates

\( \psi_{ds}, \psi_{qs} \): Stator flux in \( dq \) coordinates

\( \omega \): Rotating speed of the arbitrary reference frame

\( \omega_i \): Frequency for the high frequency injection method

\( \omega_e \): Synchronous rotating speed

\( \omega_r \): Rotor speed

\( \omega_{r-HFI_{XF}} \): Upper limit of the HFI crossover function block

\( \omega_{r-VM_{XF}} \): Lower limit of the voltage model crossover function block
BIBLIOGRAPHY


[40] ABB, Direct torque control - the world’s most advanced AC drive technology, [www.abb.com](http://www.abb.com), 2002.


