University of Technology, Sydney

Improved Direct Torque Control
of a Brushless Doubly-Fed
Reluctance Machine

William K Song

Supervisor:
Prof. David G Dorrell

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for the degree of Doctor of Philosophy
in the
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Faculty of Engineering and IT

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Certificate of Original Authorship

I certify that the work in this thesis has not previously been submitted for a degree nor has it been submitted as part of requirements for a degree except as fully acknowledged within the text.

I also certify that the thesis has been written by me. Any help that I have received in my research work and the preparation of the thesis itself has been acknowledged. In addition, I certify that all information sources and literature used are indicated in the thesis.

Signature of Student:

Date:
my parents,

my wife Julie,

my kids Doori, Ahri and Bohri

and

my research supervisor Prof. David G. Dorrell

William Song
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William Song
Contents

Certificate of Original Authorship i

Acknowledgements iii

Contents iv

List of Figures vii

List of Tables xi

Abbreviations and Symbols xii

Abstract xiv

1 Literature Review 1

1.1 Introduction ................................ 1
1.2 Control Structures ................................ 6
1.3 Thesis Outline .................................. 9
1.4 BDFRM Literature Review .......................... 10
1.5 Mathematical Equation of BDFRM .................. 13
1.6 Rotor Design for Brushless Doubly Fed Reluctance Machine .... 17
1.7 Conclusions ................................ 21

2 Direct Torque Control of the Brushless Doubly Fed Reluctance Machine 22
  2.1 Scalar Control of BDFRM ................................ 24
  2.2 Mathematical Derivation of DTC ........................ 30
  2.3 Direct Torque Control .................................. 34
  2.4 Equivalent circuit analysis ................................ 45
  2.5 Conclusions ............................................. 49

3 Improved DTC of BDFRM 51
  3.1 Space Vector Modulation ................................. 52
  3.2 Improved DTC method for BDFRM ...................... 60
  3.3 Conclusions ............................................. 62

4 Simulation Studies of the Proposed DTC 64
  4.1 Ideal Simulation Model of the Brushless Doubly Fed Reluctance Ma-
        chine .................................................. 65
  4.2 Simulation in asynchronous induction motor mode ........ 71
  4.3 Simulation with simple open-loop secondary frequency control .... 73
  4.4 Simulation Results of the Proposed DTC ................ 75
  4.5 Conclusions ............................................. 89

5 Experimental Results 90
  5.1 Experimental Test Bench Design ......................... 91
  5.2 Experimental Results – Step Changes in Speed .......... 95
  5.3 Comparison of Experimental and Simulations ............ 99
  5.4 Conclusions ............................................. 101
List of Figures

1.1 Cascaded induction machine connection .......................... 11
1.2 Concept of Brushless Doubly Fed Reluctance Machine ............ 12
1.3 Secondary winding frequency requirements for different pole number combinations and 50 Hz power winding frequency with 240V ... 15
1.4 d-q equivalent circuit of BDFRM ................................. 17
1.5 Design options for a 4-pole rotor; (a) traditional salient pole (b) axial laminations (c) segmented radial laminations ................. 18
1.6 Ducted rotor magnetic flux vector .................................. 19
1.7 Example of one segment of a 6-pole rotor .......................... 20
1.8 Machine laminations for assembly .................................. 20

2.1 Frequency and voltage relationship for $V/f$ control. The base speed is at $f_{rated}$ ............................................................... 26
2.2 Characteristics of speed and torque for $V/f$ control, constant maximum torque is possible regardless of the speed up to the base speed . 27
2.3 Basic diagram of voltage/frequency control ......................... 29
2.4 Relationship of the reference frames for the BDFRM .............. 33
2.5 Basic diagram of the conventional DTC ............................ 35
2.6 Basic circuit diagram of the Voltage Source Inverter .............. 36
2.7 All voltage vectors and six sectors for the 3-phase VSI .......... 37
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.8</td>
<td>Basic concept diagram of DTC</td>
<td>41</td>
</tr>
<tr>
<td>2.9</td>
<td>Trajectories of the secondary flux $\lambda_s$ and its reference $\lambda_s^*$</td>
<td>41</td>
</tr>
<tr>
<td>2.10</td>
<td>Simple equivalent circuit as a generator</td>
<td>45</td>
</tr>
<tr>
<td>2.11</td>
<td>Simple equivalent circuit as a motor and frequencies referred to grid</td>
<td>46</td>
</tr>
<tr>
<td>2.12</td>
<td>Doubly-fed per-phase circuit referred to the power winding</td>
<td>48</td>
</tr>
<tr>
<td>3.1</td>
<td>Eight switching state topologies of a voltage source inverter</td>
<td>53</td>
</tr>
<tr>
<td>3.2</td>
<td>Reference vector movement</td>
<td>54</td>
</tr>
<tr>
<td>3.3</td>
<td>Modulation algorithm</td>
<td>55</td>
</tr>
<tr>
<td>3.4</td>
<td>Ripple current with regards to the position of the non-zero voltage vector</td>
<td>58</td>
</tr>
<tr>
<td>3.5</td>
<td>Optimal switching pattern of SVM</td>
<td>58</td>
</tr>
<tr>
<td>3.6</td>
<td>All switching patterns</td>
<td>59</td>
</tr>
<tr>
<td>3.7</td>
<td>Control block diagram of the proposed DTC method - this is a modified diagram when compared with Fig 2.5</td>
<td>60</td>
</tr>
<tr>
<td>3.8</td>
<td>Flow chart of the space vector modulation: modified $\alpha$ to $\theta$ when compared with Fig 3.7</td>
<td>62</td>
</tr>
<tr>
<td>4.1</td>
<td>Ideal simulation model of the BDFRM for Matlab/Simulink®</td>
<td>67</td>
</tr>
<tr>
<td>4.2</td>
<td>Variation of control winding frequency with speed when the power winding connected to 50 Hz grid. Two machines (2-6-4, 4-8-6) are displayed</td>
<td>71</td>
</tr>
<tr>
<td>4.3</td>
<td>Speed and torque transient run-up for shorted control winding</td>
<td>72</td>
</tr>
<tr>
<td>4.4</td>
<td>Results for 50 Hz secondary control winding – run-up from -50 Hz to 50 Hz</td>
<td>74</td>
</tr>
<tr>
<td>4.5</td>
<td>Converting process among the three reference frames</td>
<td>75</td>
</tr>
<tr>
<td>4.6</td>
<td>Simulation responses for a speed change from 0 to 50 rpm with the proposed DTC</td>
<td>77</td>
</tr>
</tbody>
</table>
List of Figures

4.7 Simulation responses for a speed change from 0 to 50 rpm with the conventional DTC ........................................... 78
4.8 Simulation responses for a speed change from 200 to 400 rpm with the proposed DTC ........................................... 80
4.9 Simulation responses for a speed change from 200 to 400 rpm with the conventional DTC ........................................... 81
4.10 Simulation responses for a speed change from 0 to 800 rpm with the proposed DTC ........................................... 83
4.11 Simulation responses for a speed change from 0 to 800 rpm with the conventional DTC ........................................... 84
4.12 Simulation response for a load torque change from 0 to 10 Nm to -10 Nm with the proposed DTC ........................................... 85
4.13 Primary(top) and secondary(bottom) winding currents for a speed change from 0 to 800 rpm with the proposed DTC ........................................... 87
4.14 Primary(top) and secondary(bottom) winding currents for a speed change from 0 to 800 rpm with the conventional DTC ........................................... 88
5.1 Diagram of the experimental platform scheme ........................................... 92
5.2 Experimental Test Bench – Prototype BDFRM and DC machine ........................................... 93
5.3 Basic Functional Diagram of TMS320F28335 [80] ........................................... 94
5.4 DSP controller, Circuit breaker and pre-charge resistors and Semicron IGBT inverter kit ........................................... 95
5.5 Low speed response from 0 to 50 rpm (1 s per horizontal division and 10 rpm per vertical division) ........................................... 97
5.6 Medium speed response from 200 rpm to 400 rpm (with the initial step from 0 to 200 rpm – 1 s per horizontal division and 50 rpm per vertical division) ........................................... 97
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.7</td>
<td>High speed response from 0 to 800 rpm (1 s per horizontal division and 200 rpm per vertical division)</td>
<td>98</td>
</tr>
<tr>
<td>A.1</td>
<td>Stator Dimensioning</td>
<td>118</td>
</tr>
<tr>
<td>A.2</td>
<td>Rotor Dimensioning</td>
<td>119</td>
</tr>
<tr>
<td>A.3</td>
<td>Stator Drawing</td>
<td>120</td>
</tr>
<tr>
<td>A.4</td>
<td>Rotor Mounting</td>
<td>121</td>
</tr>
<tr>
<td>A.5</td>
<td>Bed Plate and Stator Mountings</td>
<td>122</td>
</tr>
<tr>
<td>A.6</td>
<td>Actual Machine</td>
<td>123</td>
</tr>
<tr>
<td>B.1</td>
<td>TI TMS320F28335 DSP Board</td>
<td>124</td>
</tr>
<tr>
<td>B.2</td>
<td>DSP I/O Interface Module</td>
<td>125</td>
</tr>
<tr>
<td>B.3</td>
<td>BDFRM with torque transducer mounted on the shaft</td>
<td>126</td>
</tr>
<tr>
<td>B.4</td>
<td>The overall photos of the hardware system</td>
<td>126</td>
</tr>
</tbody>
</table>
List of Tables

2.1 Binary codes for all voltage vectors ................................. 38
2.2 Selected vector definition ............................................. 39
2.3 Optimum switching voltage vector look-up table .................... 44
3.1 Inverter switching states and output voltages ....................... 56
4.1 Simulation parameters .................................................. 69
Abbreviations and Symbols

$V_{pd}, V_{pq}$ primary winding d and q voltage;
$V_{sd}, V_{sq}$ secondary winding d and q voltage;
$i_{pd}, i_{pq}$ primary winding d and q current;
$i_{sd}, i_{sq}$ secondary winding d and q current;
$\omega$ reference frame angular velocity;
$\omega_r$ rotor speed of the machine;
$p$ differential operator;
$p_g$ primary winding pole number;
$p_s$ secondary winding pole number;
$p_r$ rotor pole number;
$R_p, R_s$ primary winding resistance and the secondary winding resistance;
$L_p, L_s$ primary winding inductance and the secondary winding inductance;
$L_m$ mutual inductance;
$\lambda_p, \lambda_s$ primary and secondary flux;
$\lambda_{pd}, \lambda_{pq}$ primary d axis and q axis flux;
$\lambda_{sd}, \lambda_{sq}$ secondary d axis and q axis flux;
$\lambda_{s0}$ secondary original flux value;
$f_p, f_s$ grid frequency and secondary winding electrical frequency;
$f_r$ rotor mechanical rotational frequency;
<table>
<thead>
<tr>
<th>Abbreviations and Symbols</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_e$</td>
<td>electro-magnetic torque;</td>
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<tr>
<td>*</td>
<td>reference signal indicator;</td>
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<tr>
<td>BDFRM</td>
<td>Brushless Doubly-Fed Reluctance Machine;</td>
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<tr>
<td>DTC</td>
<td>Direct Torque Control;</td>
</tr>
<tr>
<td>SVM</td>
<td>Space Vector Modulation;</td>
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<tr>
<td>DFIG</td>
<td>Doubly-Fed Induction Machine;</td>
</tr>
<tr>
<td>BDFIM</td>
<td>Brushless Doubly-Fed Induction Machine;</td>
</tr>
<tr>
<td>MMF</td>
<td>Magneto-Motive Force;</td>
</tr>
<tr>
<td>V/f Control</td>
<td>Voltage/Frequency Control;</td>
</tr>
<tr>
<td>FOC</td>
<td>Field Oriented Control;</td>
</tr>
<tr>
<td>TPIA</td>
<td>Torque Per Secondary Inverter Ampere;</td>
</tr>
<tr>
<td>MTPIA</td>
<td>Maximum Torque Per Secondary Inverter Ampere;</td>
</tr>
<tr>
<td>VSI</td>
<td>Voltage Source Inverter;</td>
</tr>
<tr>
<td>THD</td>
<td>Total Harmonic Distortion;</td>
</tr>
<tr>
<td>HRPWM</td>
<td>High Resolution Pulse Width Modulation;</td>
</tr>
</tbody>
</table>
An improved Direct Torque Control algorithm for the brushless doubly-fed reluctance machine (BDFRM) has been proposed in this thesis. The conventional DTC method has some well-known advantages such as simple implementation, fast torque and speed response, and independence from machine parameters. However, despite these advantages, the conventional DTC method also has some drawbacks including high torque and speed ripple, variable switching frequency, and poor performance during the transient periods. The main motivation of this thesis is to overcome the disadvantages of the conventional DTC method.

The main focus of the work is to adapt the Space Vector Modulation method to the control problem instead of the vector selection method. SVM is a well-known method which can reduce the torque and speed ripple, and provide a fix to the problem to the variation of switching frequency and possible over-frequency. The proposed theory is implemented with simple modification, so that all the advantages
of the conventional DTC, such as easy implementation and fast response, still remain, but the issues concerned with torque and current ripple are reduced.

This thesis is essentially divided into three main parts. The first part is the theory development which compares the fundamental principles of DTC and the proposed new method. In the second part, the proposed theory is verified as giving improved performance. This was done by comparing the simulation results of the conventional DTC algorithm and the proposed algorithm. The proposed method is shown to significantly reduce the torque and speed ripple, and improve the speed response during the transient period. The third part contains a set of experimental results for verification of the simulation results, the proposed theory and the real time control system are implemented using a TI DSP module, a conventional IGBT inverter kit and a specially constructed prototype which has a radially-laminated rotor with ducts in the form of 4-8-6 pole BDFRM. Transient steps in speed were used to test the control and it was found to work well at low speed and thought the synchronous speed point where drive responses can be challenging.
Chapter 1

Literature Review

1.1 Introduction

In recent years the types and topologies of rotating electrical machines have expanded and diversified. This is mainly due to the development of power electronic converters which can deliver variable voltage and frequency over a wide range and with high efficiency. Other factors are improved design and analysis techniques, using computational electromagnetic and control simulation methods, and the development of many new applications, mostly in the renewable energy and automotive drives areas which require high-efficiency variable-speed operation.

Prior to the development of power electronic converters, the DC machine was the main variable speed drive. Indeed, in some very high power applications, such as the mining and steel industries, they are still extensively used although they may now be driven from an AC-to-DC or DC-to DC power electronic converter rather than a more classical Ward-Leonard motor-generator set (if supplied from an AC power system), or from a diesel/DC generator with simple field control [1]. They are
still used in small simple applications, especially with permanent magnet field operation. The DC machine has good characteristics in terms of wide torque operation and good speed control. The DC machine is easy controlled using either a voltage source controller or current source controller. However, they have several disadvantages which are related to the brush and commutator structure. This makes them larger and more expensive to manufacture; they require protection because they can produce sparks, and the brush gear requires regular maintenance, replacement, and cleaning. Over recent years inverter-fed AC machines have replaced them [2, 3]. These replacement machines are mostly squirrel cage induction machines (particularly in traction motors) but other machines include brushless permanent magnet machines (high efficiency machines, and with the use of high energy rare earth magnets, which are also high power dense machines suitable for electric and hybrid-electric vehicles), various types of reluctance machines (switched reluctance and more recently synchronous reluctance machines, for actuators, servos and power drives), and wound-rotor (doubly-fed) induction machines (derived from slip energy recovery schemes and now often used as wind turbine generators) [4].

The squirrel cage induction machine is the most widely used AC machine because it has a simple and brushless structure, so they are low cost in terms of both manufacture and maintenance. In the most basic form they are grid-fed constant speed machines that can drive pumps and fans; the speed is function of the grid frequency, pole number, and slip (which must be close to zero for good efficiency) [5]. The development of the power electronic inverter controller and various control strategies have now made the variable-speed induction machine now a cheaper alternative to the DC machine drive in most power drive applications. There are some disadvantages to the induction machine drive. The primary disadvantage is that the speed control is complex because it has non-linear characteristics and the rotor has no saliency
so tracking the rotor and the rotor MMF is not straightforward, and there is mutual interference between the machine parameters. This means the machine speed is more difficult to control compared to a DC machine, particularly at low speed, and all the power must flow through a multiple phase inverter rather than a simpler DC-to-DC chopper. This it will make the controller cost higher than an equivalent DC machine, although this cost difference is now becoming more marginal.

As already mentioned, another disadvantage is the need for higher starting current, even with a variable frequency inverter, since the machine always must start with unity slip. In a grid-connected machine this is 5 to 10 times full load current but under inverter control this will be only maybe up to about 50% higher; but his is very dependent on the motor and control. Careful control may minimise this.

The final disadvantage of the induction machine is the need to supply the field to the machine through the main winding which may give low lagging power factor when the machine is under lightly loaded conditions; the need to supply reactive power to the machine will make the controller system larger and the power losses potentially higher. In a DC machine the flux is produced either through separate DC windings or permanent magnets so that the losses can be minimized and there are no power factor issues.

The brushless permanent magnet machine can be the most efficient machine available [6]. However, in very large machines they require a large amount of rare earth permanent magnet material which becomes expensive. In smaller machines then the cost becomes more manageable although the sourcing of rare earth materials can be problematic since most is sourced from China and subject to issues related to sourcing material from a single country. Ferrite magnets can be used in smaller
drives but they have lower efficiency and can be subject to demagnetization. The machines can either be AC or DC controlled. This means that the input current is either sinusoidal, requiring an encoder for full position sensing, or trapezoidal which requires Hall effect position sensing for current switching points. The machines are current fed via a power electronic inverter. Again, they can suffer from poor power factor requiring higher inverter device ratings. These machines make excellent servo drive machines but are more scarce in terms of high power drives and generators.

Switched reluctance machines are still used in many actuators and drives. They are simple machines with concentrated windings and double saliency. However, they have not found extensive markets because their control can be difficult, and they have power factor, and noise and vibration issues. The current is supplied from a unipolar converter and the phase current is approximately trapezoidal. More recently synchronous reluctance machines have come on the market. These have sinusoidal current and ducted reluctance rotors [7]. Again, they require magnetization which leads to lower efficiency and low power factor. They also need full position feedback from an encoder and their control can be complex.

The final machine mentioned above is the wound rotor induction machine. These were used as drive motors that required high starting torque and/or limited starting current but this demand has died away with the development of the inverter-fed cage induction motor system. However, they have enjoyed a renaissance in the renewable energy industry as a generator, particularly in wind turbines. They offer generation over a wide speed range. While all the previous machines can be used as generators, all the generated power needs to be conditioned via a rectifier, or for reactive power control, a controlled rectifier or pulsed wave modulated (PWM) inverter operating as a rectifier. A wound rotor induction machine can generate over a reasonably wide
speed range by connecting the stator to the grid and the rotor via a bidirectional inverter. Only up to about 25% of the power then has to pass through a power electronic converter which significantly reduces the cost of the power electronics. This is termed a doubly-fed induction generator (DFIG) [8–10]. However, the rotor winding, which is a three-phase distributed winding, is still fed via slip rings and brushes, which again leads to additional cost and maintenance. It would therefore be advantageous to develop a machine with similar DFIG properties, but without the slip rings and brushes. One such machine, which in recent years has attracted some interest in the research community, is the Brushless Doubly Fed Reluctance Machine (BDFRM). [11, 12]

The BDFRM is one of a family of slip power recovery machines being investigated for a wide range of energy applications [13]. It can be operated in several different modes but one is as a slip energy recovery machine which is similar to the DFIG. The BDFRM offers higher frequency operation at a given speed which can mean that this machine may offer higher efficiency, higher power density, smaller size and lower cost, as illustrated in [14]. This machine, as already stated, has a brushless structure which is an advantage over BDFM. This combination creates the enticing prospect for a robust, controllable, low cost and low cost maintenance machine.

The BDFRM has two sets of 3-phase windings in the stator. These have different pole numbers. One is connected directly to the grid which is called the primary winding or power winding and another one is connected to grid via a bi-directional inverter which is called the secondary winding or control winding. The windings are magnetically linked via a reluctance rotor that modulates the stator MMF waves to produce rotating flux waves of different speed and pole number. The machine
is controlled via a bi-directional inverter. There are similarities to the DFIG machine; however, the two windings have much weaker linkage and the machine has two different rotating reference frames and the rotor has a third rotating reference frame. This represents a more complex control requirement than the DFIG, and indeed most other electrical machines. Because the BDFRM only requires a partially rated converter it is possible to reduce the cost of the drive system and the brushless nature of the machine increases its reliability. This is especially beneficial in large power applications such as wind turbines and large pumps. Therefore various control methods for the BDFRM have been investigated [15–19].

1.2 Control Structures

In this section, various electrical machine control issues are reviewed before the final structure is selected for the BDFRM.

The separately excited DC Motor is controlled via voltage control of the field and commutator; however, AC machines such as the induction machine, brushless permanent magnet machine, or DFIG require control of the flux and current. There are two kind of control techniques that have developed for these types of machines: field oriented control and vector control. The similarities in the structures between the DFIG and BDFRM mean that the vector control method has been extensively used for BDFRM control. Vector control leads to good control characteristics; using torque and flux control methods requires the rotor magnetic flux and speed; i.e., magnetic flux and rotor speed information is essential for vector control [20–24]. To
obtain the magnetic flux, it may be necessary to use a magnetic flux sensor. However, using a magnetic flux sensor such as hall probe in the airgap, will reduce the structural robustness for BDFRM machine. Hence, this is not normally done and flux and its orientation is estimated from the current and machine parameters. For obtaining the rotor speed (and position), it is necessary to use a speed sensor such as an encoder or resolver. Using sensors will increase the system cost and the control algorithm will be more complex. Therefore, for overcoming the disadvantages of vector control, many papers have been published on studies which aim to get fast torque and flux response.

Direct Torque Control (DTC) is one of several different drive vector control methods. It is considered to be one of the most high performance BDFRM drive control strategies [25, 26]. [25] considers a brushless doubly-fed induction machine which has a wound rotor rather than reluctance rotor, and [26] considers a small BDFRM with axial laminations which is a prototype machine which can generate torque to some extent through eddy currents in the rotor [27]; these represent more robust machines less likely to lose synchronism. DTC offers simple implementation and fast torque response, as well as better speed control and little dependence on machine parameters. This reduces the drive hardware cost when compared to vector controlled drives [28]. In the DTC method, the selection of optimum inverter switching is made in order to keep the flux and torque variations to within the hysteresis controller. Because of the absence of coordinate transformations and the lower computational burden on the microprocessor, this method generally allows higher control rates and faster response. Using the switching table for the control of a bi-directional inverter means there is no pulse width modulation module. These attractive features make it undoubtedly one of the best methods among the control methods; however, the conventional DTC strategy produces high torque and flux ripple, current harmonics,
variable switching frequency and low performance during transient torque periods. This is because of the use of the switching table selection method.

This thesis proposes an improved DTC method for controlling the BDFRM. This is the main claim to originality in this thesis since it has not been applied to this form of electrical machine. It is developed and implemented experimentally. The method is applied to a prototype machine that is specially developed and originality is claimed since it is an untested design and it is a loosely coupled machine in terms of the two windings which is somewhat different from the DFIG machine. DTC has been used in several Doctoral studies of cage induction machines [27], DFIGs [27, 29, 30], brushless permanent magnet machines [31], as well as several others. In a divergence from the conventional DTC method, the improved DTC method is implemented using a Space Vector Modulation (SVM) technique. The SVM technique is the best well-known way to reduce torque and flux ripple, and current harmonics [32, 33]. Thus, this proposed method can provide low torque and flux ripple and constant switching frequency, and also gives good performance during transient torque periods. The proposed control method will be analysed using MATLAB Simulink® then implemented on a new prototype BDFRM using an IGBT (Insulated Gate Bipolar Transistor) Voltage Source Inverter and a TI micro controller.
1.3 Thesis Outline

The remainder of this thesis is structured in the following way:

Chapter 1 puts forward a critical review of the BDFRM in general and the design aspects of the prototype BDFRM. This includes a per-phase equivalent circuit analysis.

Chapter 2 provides a basic understanding of the DTC principles when applied to the BDFRM. This chapter also presents some basic control schemes.

Chapter 3 presents an improved DTC scheme by modification of the traditional switching strategies. Special attention is paid to the improvement in machine performance due to the use of these strategies.

Chapter 4 describes in detail of the main components of the improved DTC controller in simulation when applied to the BDFRM.

Chapter 5 addresses the real time implementation issues and describes the development of an experimental controller. The hardware components are introduced. A set of experimental results generated by executing the improved DTC algorithm is shown.

Chapter 6 concludes proposed thesis with presented results and discusses some recommendations for future work.

Appendix A shows the details of new designed prototype BDFRM.
Appendix B presents the experimental hardware system and photographs of its development.

Appendix C lists the published papers related to this project.

1.4 BDFRM Literature Review

Various Doubly Fed Machines (DFMs) have been developed; their concepts can be derived from two cascaded induction machines as shown Fig 1.1. The primary 2p-pole induction machine is connected to power grid and is supplied power at constant frequency and the secondary 2q-pole induction machine is connected to a variable frequency inverter for machine speed control. The main advantage of this arrangement is the speed can be controlled to some rated speed by varying the secondary machine frequency so that only part of the system power is via the inverter. However, this cascaded induction machine has the major disadvantage that the primary machine is operating at very high slip so that it is very inefficient; the operating speed range is likely to be very narrow for all practical purposes. It also uses two machines which is costly. If the two machines were electrically and magnetically linked, and put into the same frame, then this would be advantageous. To do this, a new type of cascaded induction machine was researched and simulated in the late 60s and early 70s and the self-cascade induction machine was developed [34–38].
The self-cascaded machine targeted medium and high power applications with limited speed variation. The main concept was to combine the two machines into same frame by placing two separated windings of different pole into same stator.

This basic concept of mixed-pole windings in a single stator was extended with a special cage-type rotor structure. This machine concept became known machine as the Brushless Doubly Fed Induction Machine (BDFIM). The BDFIM is a development of the classic cascaded induction machine [39]. This machine brings together all the advantages of the cascaded induction machine in order to achieve high torque with some variable speed range. It also has a brushless structure which has a clear advantage over the conventional DFIG (which can motor as well as generate). The BDFIM has good low-speed high-torque characteristics with some speed variation, and the machine structure is brushless. However, the BDFIM has some disadvantages; the control system is complex and there is a stability performance problem around the synchronous speed where there is no torque. There are three electrical circuits – two in the stator and one on the rotor and the rotor current cannot be
DC since it relies on the voltage being induced into it by the stator windings.

Broadway and Burbridge investigated a cageless axially-laminated rotor and established the fundamental operating principles of a new type of machine - the Brushless Doubly Fed Reluctance Machine [40, 41]. The BDFRM uses a synchronous reluctance machine type of rotor structure, so the machine potentially has better efficiency and easier control than the BDFIM since the rotor position can be fed back and no rotor MMF position needs to be estimated. The BDFRM should have more stable performance over all the speed range. This machine should be able to overcome the disadvantages of the BDFIM. Fig 1.2 shows the basic concept of connection for BDFRM.

![Figure 1.2: Concept of Brushless Doubly Fed Reluctance Machine](image)

Although the BDFM has many advantages, the various types of BDFM were largely ignored until the 1990s because of the lack of reliable power electronic converters and the cost of the main controller hardware. A fast and accurate hardware system is essential for the implementation of a BDFM controller. This means the controller
cost was higher than other machine controllers. In the 1990s, due to the fast development of the personal computer and improvement in microprocessor performance, the BDFM was again the focus of research studies. Some new control algorithms were developed for controlling of the BDFIM; these used a field oriented control theory and strategy aimed at possible implementation in a practical system [42]. After successful development of new control algorithms, several theoretical studies were published for use in a BDFIM drive [43]. Interest has also moved to the BDFRM. Basic concepts were developed by Broadway and Burbridge. This led to general basic theory for the BDFRM which has been analysed and simulated by researchers [44, 45].

1.5 Mathematical Equation of BDFRM

The basic mathematical equations of the BDFRM allow an understanding of the machine and they are necessary for the analysis and simulation of the control algorithms for the BDFRM. This section will explain the mathematical BDFRM model. A per-phase analysis is given in a later section.

The BDFRM stator is similar to induction machine in several ways, but it has two sets of three phase windings with different pole numbers. The magnetic coupling between the two sets of windings relies on the reluctance of the rotor. The rotor pole number should be equal to the total number of poles of the stator windings. More detailed numerical assumptions and analyses are put forward in previous research papers [46–48]. This detailed basic analysis of BDFRM will not be put forward in
this thesis but essentially the rotor reluctance modulates the stator MMF waves to produce flux waves with different pole number leading to cross coupling between windings of different pole number. It should be noted that popular machine combinations have been the 2/6 pole machine with a 4 pole rotor, and the 4/8 pole machine with a 6 pole rotor. A 4/6 pole machine with a 5 pole has been tried [49] but having flux waves with pole-pairs differing by 1 is ill advised because it can lead to unbalanced magnetic pull as illustrated by [50, 51]. This generates considerable noise and vibration.

It also needs to be clearly stated that the machine speed needs to be synchronized to the primary and secondary windings. The primary winding is supplied from a constant 50 Hz grid supply while the secondary winding frequency varies with machine speed. Taking the high pole winding to be the primary winding, the synchronizing requirement for the inverter-fed secondary winding is:

\[
f_s = Pf_r \pm f_p \tag{1.1}
\]

where \(f_s\) represents the secondary winding electrical frequency, \(f_r\) is the rotor mechanical rotational frequency, and \(f_p\) is the grid frequency; \(P\) is the number of poles on the rotor. The frequency variation of the secondary winding for variation of the speed for different pole number machines is shown in Figure 1.3. It can be seen that the secondary frequency goes through a zero point. At lower speeds the frequency is negative which is effectively a backwards rotating MMF wave.
The main concept of the basic mathematical model is that it consists of two reference frames which both use d-q models. Using an equivalent two phase model of the three phase machine windings makes the machine more straightforward to analyse and simulate because the d-q model has less variables than the three phase model. The d-q equations of BDFRM model are given in (1.2). There are four expressions; two equations for the primary winding which connect to the grid and two equations for the secondary winding which connect to the control inverter.

\[
\begin{bmatrix}
V_{pd} \\
V_{pq} \\
V_{sd} \\
V_{sq}
\end{bmatrix} =
\begin{bmatrix}
R_p + L_{pp} & -\omega L_p & L_{mp} & \omega L_m \\
\omega L_p & R_p + L_{pp} & \omega L_m & -L_{mp} \\
L_{mp} & (\omega_r - \omega) L_m & R_s + L_{sp} & (\omega_r - \omega) L_s \\
(\omega_r - \omega) L_m & -L_{mp} & (\omega_r - \omega) L_s & R_s + L_{sp}
\end{bmatrix}
\begin{bmatrix}
i_{pd} \\
i_{pq} \\
i_{sd} \\
i_{sq}
\end{bmatrix}
\]  

(1.2)

where,

V_{pd} and V_{pq} are the primary winding d and q voltages;

V_{sd} and V_{sq} are the secondary winding d and q voltages;
Chapter 1. Literature Review

\(i_{pd}\) and \(i_{pq}\) are the primary winding \(d\) and \(q\) currents;
\(i_{sd}\) and \(i_{sq}\) are the secondary winding \(d\) and \(q\) currents;
\(\omega\) and \(\omega_r\) are the reference frame angular velocity and the rotor speed of the machine;
\(p\) is the differential operator;
\(L_p, L_s\) and \(L_m\) are the primary, secondary and mutual inductances; and
\(R_p\) and \(R_s\) are the primary and secondary winding resistances.

From (1.2), the \(d-q\) equations, expressions can be developed for the equivalent circuit of the BDFRM as illustrated in Fig 1.4:

\[
\begin{align*}
v_{pd} &= R_p i_{pd} + \frac{d}{dt}(L_p i_{pd} - L_m i_{sd}) + \frac{d}{dt}[L_m(i_{pd} + i_{sd})] - \omega \lambda_{pq} \\
v_{sd} &= R_s i_{sd} + \frac{d}{dt}(L_s i_{sd} - L_m i_{pd}) + \frac{d}{dt}[L_m(i_{sd} + i_{pd})] - (\omega_r - \omega) \lambda_{sq} \\
v_{pq} &= R_p i_{pq} + \frac{d}{dt}(L_p i_{pq} - L_m i_{pq}) + \frac{d}{dt}[L_m(i_{pq} - i_{sq})] - \omega \lambda_{pd} \\
v_{sq} &= R_s i_{sq} + \frac{d}{dt}(L_s i_{sq} - L_m i_{sq}) + \frac{d}{dt}[L_m(i_{sq} - i_{pq})] + (\omega_r - \omega) \lambda_{sd}
\end{align*}
\] (1.3)

where,
\[
\begin{align*}
\lambda_{pd} &= L_p i_{pd} + L_m i_{sd} \\
\lambda_{sd} &= L_s i_{sd} + L_m i_{pd} \\
\lambda_{pq} &= L_p i_{pq} - L_m i_{sq} \\
\lambda_{sq} &= L_s i_{sq} - L_m i_{pq}
\end{align*}
\]

Hence (1.3) represents the \(d-q\) equivalent circuits of the BDFRM as shown Figure 1.4.
1.6 Rotor Design for Brushless Doubly Fed Reluctance Machine

The main principles of operation of the BDFRM have been researched and many papers have been published on the analysis of the BDFRM operation [45–48]. Also, the rotor structure of BDFRM has been addressed for the efficiency of machine [49, 50]. The salient rotor structure has had recent attention in order to improve the
efficiency in order to move the machine from a laboratory proof-of-concept prototype towards being a realistic industrial machine. While the optimum rotor design is not the focus in this thesis, it will be described briefly in this section.

Figure 1.5 shows three alternative options for the rotor design. These design options have been described in previous research papers [14, 50]. As stated earlier, the purpose of the salient pole rotor is to modulate the stator Magneto-Motive Force (MMF) waves to produce flux waves with different pole number. This is in order to produce coupling between the two stator windings. This is different from traditional induction machines; the BDFRM has an air-gap flux distribution which is split between the different pole number waves generated by the rotor saliency as well as the MMF distribution. Thus, the cross coupling will be created between magnetizing and torque producing currents. As a result, the flux linkages have to be considered in the BDFRM: the self and mutual flux linkages. Leakage inductance also has to be considered. This is similar to a transformer as well as an induction machine.

An optimised rotor design for the BDFRM should have a high saliency ratio [51].
Comparing the rotor structures in Figure 1.5 (which have been used in previous studies), the segmented radial lamination rotor can achieve very high saliency ratio value. The segmented radial lamination rotor is made of multiple thin air ducts, so it is called a ducted rotor [52, 53]. The interleaved laminations between the ducts act as a flux guides as shown in the flux plot in Figure 1.6.

![Ducted rotor magnetic flux vector](image)

\textbf{Figure 1.6: Ducted rotor magnetic flux vector}

In this thesis, the 6-pole segment rotor is used for the experimental prototype as opposed to the 4-pole design. An example of one segment of a 6-pole rotor design is shown in Figure 1.7 and the actual rotor lamination is shown in Figure 1.8.

The BDFRM in this thesis is assembled with this lamination which were made specifically for this project; the stator shown in Figure 1.8 is used in conjunction with it and 4 and 8 pole windings are wound in this.
Figure 1.7: Example of one segment of a 6-pole rotor

Figure 1.8: Machine laminations for assembly
1.7 Conclusions

This chapter has introduced the BDFRM and put forward a literature review. It has described the possible controls of the machine and the basic machine structures. In addition, it has described the development of the machine. This review helps in the understanding of why the BDFRM offers good future potential for commercialization. To briefly summarise, the BDFRM has the advantages of the cage induction machine in terms of simple brushless structure and variable speed operation. In addition, the machine operates with a de-rated inverter and is suitable for replacement of a DFIG, and even as a drive motor in some applications.

The prototype BDFRM used in this project has been designed to offer higher efficiency using a ducted rotor as explained in Section 1.6. It is expected that the BDFRM will have a 30% better efficiency compare with the conventional BDRFM. However, a suitable control has to be set up and the aim of this work is to investigate DTC when applied to this machine.

The Direct Torque Control algorithm has been successfully adapted for the BDFM in other studies but not for this design of the machine, with a ducted reluctance rotor. DTC is a more efficient control method for many different types of machine. However, DTC for the BDFRM will still have the same potential issues as DTC for any other type of machine. In this study, the advantages and disadvantages for the existing DTC algorithm is discussed, especially in relation to the BDFRM. The next chapter will address this. An adapted solution to overcome the disadvantages of the conventional DTC algorithm is then suggested.
Chapter 2

Direct Torque Control of the Brushless Doubly Fed Reluctance Machine

In Chapter 1, the basic mathematical model of the BDFRM was provided. A clear understanding of the BDFRM model is needed in order to simulate and build a controller for the BDFRM. There are three control technique categories that can be used for controlling the general BDFM; these are Based Scalar Control Techniques, Vector Control Techniques (or also called Field Oriented Control Techniques) and Direct Control Techniques. Amongst these control techniques, Direct Control is considered an alternative control technique for the BDFM. Many researchers have studied this control technique and several different methods of Direct Control have been developed. The most well-known and representative direct control technique is Direct Torque Control (DTC) [54]. In this chapter, DTC will be explained carefully. This is based on the mathematical model of BDFRM developed in Chapter 1. This chapter is structured as follows.
In the initial section of this chapter, the scalar control technique (also called voltage/frequency control) will be briefly explained for comparison with DTC of a BD-FRM [55]. This is a crude open-loop control and the disadvantages of scalar control are discussed in Section 2.1; however, scalar control techniques are still widely used in industry because they are easy to understand and establish. Scalar control is not very suitable for the BDFRM despite its simple nature. While DTC is more complicated, this technique is more easily developed compared to vector control techniques in terms of BDFRM control. This is because there are two rotating MMF and flux waves of different pole number and rotational velocity to consider, and a rotor with yet another pole number and velocity. These need to be synchronized, hence the complex vector control strategy that is needed to maintain this synchronism. Therefore, this thesis will not discuss vector control techniques to any great depth, but it can be stated that the vector control technique was developed to overcome the disadvantages of the scalar control technique. It provides more robust control performance than scalar control.

Many researchers have focused on finding more straightforward controls rather than resorting to full vector control, and as already stated DTC is one of these simpler strategies. While it is more complicated than scalar control, it is closed loop control and provides better performance than scalar control but is easier to establish than vector control. DTC has the following advantages: fast dynamic response, reduced controller work requirement, robust control against machine parameter variation, and simplicity of control method. In Section 2.2, a basic mathematical derivation for DTC will be put forward, while Section 2.3 will show the basic structure and concept of DTC.

In Section 2.4 a shift of emphasis is made. A per-phase circuit model is put forward
and related to the induction machine, and, in particular, the DFIG. This is then
developed to illustrate the range of operation of the machine and the power flow in
the machine. Comparison is made to the DFIG and it is shown that the machine
is loosely coupled and that the control is more arduous than for a DFIG and hence
there is a need to prove DTC is effective in this machine.

2.1 Scalar Control of BDFRM

A scalar control technique means that the machine is only controlled by the varia-
tion of various variable magnitudes. It is usually open loop although speed feedback
can be used as well as overcurrent protection. Coupling effects, and space and time
vectors of the machine, are ignored in this technique. Therefore, all controlled pa-
rameters are scalar and are independent. This is why the machine driver, when using
a scalar control technique, is easy and low cost to implement. It often works well in
simple power drives where precision is not needed so it is widely used in industrial
applications for pump and fan drives [56].

In an induction motor, the two parameters that can be varied are the 3-phase voltage
magnitude and the frequency. The voltage appearing across the magnetizing induct-
tance in the equivalent circuit is a function of the rate of change of flux linkage;
therefore, to maintain an almost constant air-gap flux, the voltage should increase
with the frequency. Hence, this control is called voltage/frequency control (V/f con-
trol), and the ratio of V/f is kept constant to maintain the stator flux linkage. This
is apart from when at low speed, where the frequency is low so that the equivalent
circuit resistances are dominant. There is usually a boost in voltage to overcome this. Many studies have been reported which try to solve this and also produce high torque and smooth speed. At standstill, the slip of the machine is always unity no matter what the frequency of supply is. This can create starting problems in variable speed inductions motors. Once running the slip reduces considerably to a more efficient operating point. The flux and torque of the machine are functions of the line voltage and frequency respectively; the basic concept of scalar control ignores the coupling between the flux and torque, thus the scalar control drivers provide somewhat poor performance. The vector control technique was developed to overcome the disadvantages of scalar control [57, 58].

Stator frequency control is a more efficient method for controlling the speed of an induction machine. Limited speed control is possible using voltage control, which is now very much redundant as a technique, and slip energy methods, which have been extensively developed for controlling DFIGs in winding turbines. The speed of induction machine can be controlled by the adjustable voltage magnitude and frequency supplied to the stator. The relationship between the stator frequency and synchronous speed is

\[ \omega = \frac{2\pi \cdot f_s}{p} \] (2.1)

where \( \omega \) is synchronous rotational speed of the air-gap flux wave (synchronous speed), \( f_s \) is the stator frequency and \( p \) is pole number. The machine is often called an asynchronous machine because it does not operate at the synchronous speed, it operates at a speed just below it (for motoring) or just above it (for generating). The difference between the synchronous speed and rotor rotational speed is termed the slip. Unless slip energy recovery methods are used, the slip will be below about
4% for a small machine and much less than 1% for very large machines at grid frequencies. The slope of the speed droop at a set frequency with increasing load should be constant so below grid frequency, the full-load p.u. slip is higher than this, and at frequencies above grid frequency the full-load p.u. slip will be load; i.e., at 50 Hz the full load slip is 2 %, at 100 Hz it will be 4 % and at 25 Hz it will be 4 %.

This is in the maximum torque range which exists up to the base speed, at higher speeds then field weakening is used and the full load p.u. slip will be constant at these higher speeds. When the frequency is changed in order to change the speed, the generated torque also changes to drive the speed change but should settle to deliver the correct torque to drive the load. To minimize the electric loading, and hence the losses, the machine needs to be fluxed and corrected level. This is done up to the base speed before the field weakening region is entered.

![Figure 2.1: Frequency and voltage relationship for V/f control. The base speed is at \( f_{\text{rated}} \)](image)
To avoid the problem of changing flux levels using the frequency control method, the stator voltage should change with frequency. This is shown in Figure 2.1; the ratio of $V/f$ will be remained constant for any change of $f_s$ over the linear range. At low speed there is a voltage boost, as already explained, and at high speed there is a “field weakening” region. This is for two reasons: at high frequency the iron losses increase so that the flux may have to be reduced to limit these; also the inverter voltage may have reached a maximum so that only the frequency can be increased. Over the linear region maximum torque is available. Over the field weakening region maximum power is available.

Figure 2.2 shows a set of torque/speed curves for different values of frequency. The curve up to rated $T_L$ is used. If the torque increases above $T_L$, and hence the current is an overcurrent, the frequency is reduced in an attempt to find a torque below the maximum. The advantages of the constant $V/f$ control are that the torque and the rotor current are frequency independent and the speed is proportional to frequency; hence, the control block is easily implemented.
This technique was originally developed for induction machines, but it can be adapted for brushless permanent magnetic machines, and can be used to control of BDFRM because the BDFRM model is similar to the induction machine model [59]. The scalar control technique can be easily implemented in the BDFRM because the equivalent circuit is very similar. This technique does not produce a fast dynamic response of the BDFRM. Therefore, it has a limited frequency change range which means the speed range is limited to within certain areas.

Figure 2.3 shows the basic control strategy concept of the $V/f$ control technique as applied to the BDFRM, which has a speed feedback signal which can be obtained from measurement or estimation [60]. There are two types of scalar control technique; open loop control and closed loop control. The BDFRM, since it is generally considered a synchronous machine, does require speed feedback. Only the reference speed is fed to the controller in order to control the voltage and frequency signals. The grid frequency is also required. These are applied to machine through the secondary or control connection. There is no information about the MMF vector, or rotor orientation.

The voltage magnitude is a function of the control frequency, which in turn is a function of the speed as illustrated in the Figure 1.3 in Chapter 1. The controlling equation for synchronism is (1.1). The lack of vector orientation or control is the main disadvantage in this control; the machine speed feedback signal is compared with the reference speed and then the error value between the actual speed and the reference speed is applied to the PI controller. The output of PI controller is used for control signal of the Voltage Source Inverter. The controller should be PI because obviously there can be no steady-state error in the frequencies since it is
Chapter 2. Direct Torque Control of Brushless Doubly Fed Reluctance Machine

29

Figure 2.3: Basic diagram of voltage/frequency control

synchronous. This control algorithm has more improved stability, faster dynamic response and a cleaner waveform than an open-loop $V/f$ control.

The control described above is synchronized; to use open-loop $V/f$ control on the BDFRM implies that it cannot be synchronized so that the power winding would have to be shorted rather than connected to a grid. If this is done, the machine can be run from the control winding via the inverter in a similar manner to an inverter-fed cage induction machine. This is the asynchronous operating mode of the BDFRM which has not been studied in the literature.
2.2 Mathematical Derivation of DTC

Although the scalar control technique is widely used for machines in industry, it cannot be used for robust and accurate torque control because of the mutual interference between the flux and torque and it also has a time delay in its response. The decoupling of the torque and the magnetizing flux is the weakness of the scalar control technique. The key for the robust control of a machine is the ability to control both the flux and torque from the stator current. Hence, the scalar control technique is not good for robust control. In order to achieve this, the currents will have transformed on to d-q axes. The d axis is the direct axis which represents the flux or field current, and the q axis is the quadrature axis represents the load current or torque. When it is using the d-q axis conversion for control, the machine is controlled by individual control of the flux and torque directly. This control scheme is called a vector control or field oriented control (FOC).

The vector control technique gives solutions for the issues with scalar control and provides better control quality. Vector control techniques have proved popular since being developed. The first vector control technique was introduced and developed by Blaschke, Hasse and Leonhard [51, 61–64] in 1968 and in the early 1970s. But vector control is complicated; to use this method, it is necessary to utilize several mathematical transforms. With the development of high performance microprocessors, which can calculate these mathematical transformations expeditiously, vector control has developed very quickly and now it is used widely.

DTC is a more specific form of the vector control technique. DTC has been developed for high performance control of induction machines [26, 65–69]. With the recent developments in the brushless permanent magnet machine, and because the
technique uses few machine parameters, DTC has been successfully used to control all types of brushless machine, and this includes doubly fed machines. The technique used only two control parameters; the flux and torque. The BDFRM has similar characteristics with the induction machine: two electrical circuits with AC linked by a magnetic circuit.

However, the BDFRM does have a different arrangement; it has two stator windings with different pole numbers and with different frequency voltages applied. The revolving fields are not mutually stationary. As explained in the previous section, two flux waves are produced for each winding with only one of them coupling with the other winding due to a reluctance modulation effect of the rotor structure. In this section, a mathematical derivation of DTC using the dynamic model will be developed for the BDFRM.

An expression for the torque for an electrical machine has been developed in several previous research works [16, 70–72]. This is a particularly suitable equation for explaining the DTC concept for the BDFRM:

\[
T_e = \frac{3}{2} p_r \frac{L_m}{L_p} \lambda_p i_{sq}
\]

(2.2)

where,

- \( T_e \) is the electro-magnetic torque,
- \( p_r \) is the rotor pole number,
- \( L_p, L_m \) are the primary and mutual inductances,
- \( \lambda_p \) is the primary flux,
- \( i_{sq} \) is the secondary stator current along the q-axis
\( \frac{L_m}{L_p} \lambda_p \) is the primary coupling with the secondary winding; this is a flux term which is rotating with same speed as the secondary mmf at \( \omega_s \), it is denoted as \( \lambda_{ps} \). The primary flux, \( \lambda_p \), is almost constant because the primary winding is connected to the grid. The machine is operating under fully fluxed conditions; this means the variation of the inductance is sufficiently small so that it can be ignored so that the primary and mutual inductances, \( L_p, L_m \), can be considered constant. The secondary stator q-axis current can be replaced with \( i_{sq} = i_s \sin \alpha_p \) with the secondary current vector position angle given by \( \alpha_p \). Therefore, (2.2) can be rewritten as

\[
\frac{T_e}{i_s} = \frac{3}{2} \frac{L_m}{L_p} \lambda_p \sin \alpha_p
\]  

(2.3)

This equation defines the torque per secondary inverter ampere (TPIA); and the maximum torque per secondary inverter ampere (MTPIA) is clearly achieved when \( \alpha_p = \frac{\pi}{2} \). This MTPIA is important for selecting the inverter rating and improving the efficiency for a given torque output. Figure 2.4 shows the basic relationships and the reference frames for the BDFRM.

The d-q flux mathematical expressions are given below for the dynamic model equations of the BDFRM given in (1.2) in Chapter 1:

\[
\begin{align*}
\lambda_{pd} &= L_p i_{pd} + L_m i_{sd} \\
\lambda_{pq} &= L_p i_{pq} - L_m i_{sq} \\
\lambda_{sd} &= L_s i_{sd} + L_m i_{pd} \\
\lambda_{sq} &= L_s i_{sq} - L_m i_{pq}
\end{align*}
\]  

(2.4)

where the subscripts ‘p’, ‘s’ and ‘m’ are the primary, secondary and mutual quantities respectively.
In Figure 2.4, the primary reference frame d-axis rotates at \( \omega_p \) and is aligned with \( \lambda_p \). The secondary reference frame d-axis rotates at \( \omega_s \) and is aligned with \( \lambda_{ps} \), which is \( \frac{L_m}{L_p} \lambda_p \), so the mathematical expression can be simplified to

\[
\begin{align*}
\lambda_{pd} &= L_p i_{pd} + L_m i_{sd} = \lambda_p \\
\lambda_{pq} &= L_p i_{pq} - L_m i_{sq} = 0 
\end{align*}
\]  

The last line in (2.4) can be re-written as:

\[
\begin{align*}
\lambda_{sq} &= L_s i_{sq} - L_m i_{pq} \\
&= (L_s - \frac{L_m^2}{L_p}) i_{sq} \\
&= \lambda_s \sin \alpha_s 
\end{align*}
\]  

where \( \alpha_s \) is the secondary flux angle between the secondary flux and the mutual flux \( \frac{L_m}{L_p} \lambda_p \). Finally, the torque expression (2.2) for the DTC when applied to a BDFRM
can be further defined, where \( K = L_s - \frac{L_p^2}{L_m} \), so that

\[
T_e = \frac{3}{2} \frac{p_r}{K} \frac{L_m}{L_p} \lambda_p \lambda_s \sin \alpha_s
\]  

(2.7)

This torque equation is generally used for the fundamental DTC theory. This is a straightforward way to understand the DTC method for the BDFRM. \( p_r, K, L_p, L_m \) and \( \lambda_p \) are all constant values. The machine torque then only varies with \( \lambda_{sq} = \lambda_s \sin \alpha_s \) which is obtained directly from \( i_{sq} \). Thus, the machine torque will increase or decrease with \( \lambda_{sq} \), which will increase or decrease by \( \alpha_s \).

### 2.3 Direct Torque Control

DTC was proposed by Takahashi and Noguchi in 1986 [73]. It has the following advantages: fast dynamic and accurate response, simple implementation, minimised driver operation, reliability, and robustness against machine parameter variation. DTC is based on the direct control of two variables: the electromagnetic torque and the flux amplitude. For high efficiency control of the BDFRM, DTC is the best method for controlling the machine. However, the varying nature of the switching frequency is the main drawback of DTC. This non-constant switching frequency behaviour in the inverter driver will generate non-uniform semiconductor switching losses in the inverter. In addition, this will affect the torque, flux and current, where high ripple and harmonics will be generated. Therefore, in this Chapter, the basic concepts and background theory of DTC will be described in detail, while methods that will overcome the DTC drawbacks will be discussed in Chapter 3.
For distinction between basic DTC and the proposed DTC in this thesis, basic DTC will be referred to as “Conventional DTC”. The configuration of a conventional DTC is much simpler than full vector control. Only the torque and flux are controlled directly in a conventional DTC. The control block diagram of a conventional DTC is illustrated in Figure 2.5. The controller consists of five blocks: the torque and stator flux estimators, the torque and flux controllers (called hysteresis comparators), and a switching table for voltage vector selection for the voltage source inverter [71–73].

![Diagram of conventional DTC](image)

**Figure 2.5**: Basic diagram of the conventional DTC

The basic concept of operation for a conventional DTC is very simple. The speed
of the BFRDM is measured by the speed sensor. This value is compared with the reference speed and the reference torque is then obtained. The flux and torque are calculated and estimated by measuring the voltages and currents from the machine. The estimated flux and torque are compared with the reference flux and torque during every sampling period. The error values between the estimated and reference flux and torque is re-defined by the hysteresis comparators with the desired accuracy. The outputs of the hysteresis comparators are used as the inputs to the switching look up table where the most suitable voltage vector in the switching table is selected for the voltage source inverter (VSI). The controlled 3 phase voltages are supplied to the BDFRM through the secondary winding of machine. Since the flux is assessed from the secondary voltage, and the secondary current is measured, the rotor position and primary current are not needed because these will appear in the secondary flux as one of its components, both in terms of magnitude and orientation.

![Basic circuit diagram of the Voltage Source Inverter](image)

**Figure 2.6:** Basic circuit diagram of the Voltage Source Inverter

Referring to the machine in Figure 2.5, there are two windings, a primary and secondary. The primary winding called power winding supplies grid power to the BDFRM and the secondary winding called controlled winding is to control the machine
through the VSI. The conventional VSI can be controlled by six non-zero voltage vectors and two zero voltage vectors. Figure 2.6 shows the conventional VSI circuit diagram. The non-zero voltage vectors are formulated below in the stationary reference frame. The non-zero voltage vector magnitude is defined by the DC link voltage so that

\[ v_k = \frac{2}{3} V_{dc} e^{j(k-1)\pi/3} \]  

(2.8)

where, \( k \) is an integer from 1 to 6.

Figure 2.7: All voltage vectors and six sectors for the 3-phase VSI

Figure 2.7 shows the all voltage vectors of the conventional VSI in a stationary reference frame. There are three legs to the VSI and the binary numbers represent the switch status of each leg. ‘1’ means the upper switch of the leg is ON and the lower switch is OFF, ‘0’ means the upper switch of the leg is OFF state and the lower switch is ON. For example, the voltage vector \( v_1 \) is defined 100, the most significant bit is leg A of the VSI, the middle bit is leg B, and the least significant bit is leg
C. So the first 1 means switch S1 is ON and switch S4 is OFF, the middle 0 means switch S3 is OFF and switch S6 is ON and the last 0 means switch S5 is OFF and switch S2 is ON. The binary coding of each vector is summarised in Table 2.1 [74].

<table>
<thead>
<tr>
<th></th>
<th>$v_0$</th>
<th>$v_1$</th>
<th>$v_2$</th>
<th>$v_3$</th>
<th>$v_4$</th>
<th>$v_5$</th>
<th>$v_6$</th>
<th>$v_7$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Leg A</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Leg B</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>Leg C</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

Table 2.1: Binary codes for all voltage vectors

The secondary flux magnitude can be obtained by rearrangement of the voltage equation where

$$V = RI + \frac{d}{dt}\lambda$$ \hspace{1cm} (2.9)

$$\lambda_s = \int (v_s - R_s i_s) dt$$
$$= V_s t - R_s \int i_s dt$$ \hspace{1cm} (2.10)

In (2.10), $R_s$ is small and can be neglected so that the secondary flux is proportional to the voltage vector and has same vector direction as the voltage vector. Hence it can be simplified to:

$$\Delta \lambda_s \approx V_s \Delta t$$ \hspace{1cm} (2.11)

where $\Delta t$ is a sampling period.
The secondary flux changes its orientation with the voltage vector since they are in the same secondary reference frame. When a non-zero voltage vector is applied, \( \lambda_s \) will be controlled the same direction and speed as the applied non-zero vector. When a zero vector is applied, it maintains almost the same position. Therefore \( \lambda_s \) will be controlled by a combination of non-zero voltage vectors and zero vectors.

Table 2.2 shows the selected voltage vector when \( \lambda_s \) is in the sector \( N \). If \( \lambda_s \) is in the sector \( N \) (where \( N = 1, 2 \ldots 6 \)), its magnitude can be increased or decreased by switching vectors with directions. A voltage vector in advance of the direction of \( \lambda_s \) will increase the electrical torque while the voltage vector behind the direction of \( \lambda_s \) will decrease the instantaneous torque.

<table>
<thead>
<tr>
<th>( v_N )</th>
<th>Radial Positive Voltage Vector</th>
</tr>
</thead>
<tbody>
<tr>
<td>( v_{N+1} )</td>
<td>Forward Positive Voltage Vector</td>
</tr>
<tr>
<td>( v_{N+2} )</td>
<td>Forward Negative Voltage Vector</td>
</tr>
<tr>
<td>( v_{N+3} )</td>
<td>Radial Negative Voltage Vector</td>
</tr>
<tr>
<td>( v_{N-1} )</td>
<td>Backward Positive Voltage Vector</td>
</tr>
<tr>
<td>( v_{N-2} )</td>
<td>Backward Negative Voltage Vector</td>
</tr>
<tr>
<td>( v_0 )</td>
<td>Zero Voltage Vector</td>
</tr>
</tbody>
</table>

**Table 2.2: Selected vector definition**
From (2.10), the secondary flux can be re-written by the voltage vector

\[ \lambda_s(t) = v_n t + \lambda_{s0} \]  

(2.12)

where the \( \lambda_{s0} \) is the original value of flux at the very beginning of the time interval \( t \) and \( \lambda_s(t) \) is the final value of flux at the end of the time interval \( t \).

Figure 2.8 shows the relationship between the voltage vectors and the secondary flux, this is the simple concept diagram of DTC. A suitable combination of non-zero voltage vectors and zero voltage vectors can be applied to the VSI so that a certain voltage will be supplied to the secondary winding in order to obtain the required flux-linkage locus. The flux will rotate and follow a circle with some ripple as shown in Figure 2.9. The hysteresis comparator can limit this ripple as a user demand. The two circles in Figure 2.9 illustrate the boundary of the hysteresis [75]
Figure 2.8: Basic concept diagram of DTC

Figure 2.9: Trajectories of the secondary flux $\lambda_s$ and its reference $\lambda_s^*$
The hysteresis band of the flux is defined by

\[
(|\lambda_s|^* - \frac{\Delta}{s} |\lambda_s|) \leq \lambda_s \leq (|\lambda_s|^* + \frac{\Delta}{s} |\lambda_s|)
\]  
(2.13)

where, $\lambda_s^*$ is the flux reference value.

The selection of the voltage vector describes the magnitude inside the hysteresis band and also the rotational direction. For example, if the flux is rotating in the clockwise direction in the hysteresis band, when the flux approaches the inner hysteresis band limit then $v_6$ is selected, when the flux approaches the outer hysteresis band limit then $v_5$ is selected. If the flux is rotating with anti-clockwise direction, then $v_2$ and $v_3$ are selected. When the flux is controlled, the torque can calculate from (2.7).

When $T_e$ is approaching to the torque reference $T_e^*$, the $T_e$ has to slowly reduced using the zero voltage vector. This can be done in two ways depending on the rotating direction:

\[
T_e^* - \Delta T_e \leq T_e \leq T_e^* : \text{when } \lambda_s \text{ is rotating with clockwise}
\]

\[
T_e^* \leq T_e \leq T_e^* + \Delta T_e : \text{when } \lambda_s \text{ is rotating with anti-clockwise}
\]

If $\lambda_s$ is rotating in the clockwise direction, the zero voltage vector will be selected for reducing $T_e$ and stopping $\lambda_s$ when the $T_e$ approaches $T_e^*$. When $T_e$ is approaching $T_e^* - \Delta T_e$, i.e., it is too low, the non-zero voltage vector is selected to make $\lambda_s$ rotate in to clockwise direction.

The secondary flux has a positive value, so the output of hysteresis comparator
consists of two levels as defined by (2.14). When the difference between $\lambda_s^*$ and $\lambda_s$ is higher than the hysteresis band limit, $d\lambda = 1$; when the difference is smaller than the negative hysteresis band limitation, $d\lambda = 0$:

$$d\lambda = \begin{cases} 
1, & \lambda_s^* - \lambda_s \geq \Delta \lambda \\
0, & \lambda_s^* - \lambda_s \leq \Delta \lambda 
\end{cases} \tag{2.14}$$

The torque hysteresis comparator output has three levels as defined by (2.15). This is because the reference torque has clockwise and anti-clockwise directions:

$$dT = \begin{cases} 
1, & T_{re}^* - T_e \geq \Delta T \\
0, & T_{re}^* - T_e = 0 \\
-1, & T_{re}^* - T_e \leq \Delta T 
\end{cases} \tag{2.15}$$

A suitable switching voltage vector loop-up table needs to know the angular position of the secondary flux for calculating the optimum voltage vector. This angular position can be calculated from the A-axis and B-axis of the secondary flux in the stationary reference frame:

$$\begin{align*}
\lambda_s &= \lambda_{As} + j\lambda_{Bs} \\
&= \int (u_{As} - R_si_{As})dt + j \int (u_{Bs} - R_si_{Bs})dt \\
|\lambda_s| &= \sqrt{\lambda_{As}^2 + \lambda_{Bs}^2} \\
\theta_s &= \tan^{-1}\left(\frac{\lambda_{As}}{\lambda_{Bs}}\right) \tag{2.16}
\end{align*}$$

where $u_{As}$, $u_{Bs}$, $i_{As}$ and $i_{Bs}$ are the measured voltage and current values of the secondary winding.
Table 2.3 is the optimum switching voltage look up table which allows the minimum switching frequency and related loss. The two hysteresis comparator outputs, which are the torque and flux values, are used to select the sector together with the calculation of the angle. These can be used to select a suitable reference vector in the optimum switching look up table. Once the vector is selected, the controlled switching voltage will be generated and supplied to the VSI as shown the Table 2.1.

<table>
<thead>
<tr>
<th>Hysteresis Output</th>
<th>Sector which is selected by angle</th>
</tr>
</thead>
<tbody>
<tr>
<td>$d\lambda$</td>
<td>$dT$</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>-1</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>-1</td>
</tr>
</tbody>
</table>

Table 2.3: Optimum switching voltage vector look-up table
2.4 Equivalent circuit analysis

The chapter so far has studied the application of DTC to the brushless doubly-fed induction machine. This has been the subject of several studies; however, it is further developed and modified in this study as part of a claim to originality for the work. At this point it is worth considering the per-phase equivalent circuit of the machine. Fig. 2.10 shows a simple coupled circuit for the machine. This has the generating convention used.

![Simple equivalent circuit as a generator](image)

**Figure 2.10:** Simple equivalent circuit as a generator

The grid frequency is fixed so this can be modified to the circuit in Fig. 2.11 with the rotor angle $\delta$ set to zero and the direction of the power winding current redefined. Here we can introduce the slip term in the same way as with the induction machine. At standstill $s = 1$ if the secondary MMF is forwards rotating, as it is in a DFIG. In a BDFRM, [60] illustrated that the secondary MMF wave is actually rotating in the reverse direction at stand still so this is where the BDFRM diverges from the DFIG.
and the slip is defined as \( s = -1 \) at standstill. At the synchronous speed, where the control winding frequency is zero, the slip is zero; above the synchronous speed the slip is negative for the DFIG and positive for the BDFRM. We can then derive the airgap power components which represent the mutual coupling:

\[
P_p(\text{airgap}) = 3Re[j\omega_p M\bar{I}_s\bar{I}_p^*]\text{(flowing into airgap)}
\]

\[
P_s(\text{airgap}) = 3Re[-j\omega_p M\bar{I}_s\bar{I}_p^*] = 3Re[j\omega_p \bar{I}_s\bar{I}_p^*]\text{(flowing out of airgap)}
\]

\[
P(\text{mech out}) = P_p(\text{airgap}) - P_s(\text{airgap}) = 3(1-s)Re[j\omega_p M\bar{I}_s\bar{I}_p^*]
\]

(2.17)

There are several conclusions that can be drawn from these equations. First, consider the rotor angle \( \delta = 0 \), as defined in Fig 2.10. We can observe that

- When \(-1 < s < 1\) then the magnitude of the power flowing in or out of

\[
\begin{align*}
\text{Control winding} & \quad \bar{I}_s & R_s \\
\text{Power winding} & \quad R_P & \bar{I}_p \\
\text{Variable frequency and terminal voltage (4 pole winding)} & \quad j\omega_p L_s \\
\text{Fixed frequency and terminal voltage (8 pole winding)} & \quad j\omega_p L_P \\
\end{align*}
\]

Figure 2.11: Simple equivalent circuit as a motor and frequencies referred to grid
the control winding at the airgap is less than the airgap power for the power winding. In order to restrict the power rating of the inverter in a DFIG the operating range in is often limited to the range $s = \pm 0.3$. For the BDFRM a similar regime should be adopted if it is used as a winding turbine generator.

- When the machine is motoring below the base speed for the DFIG or motoring above base speed for the BDFRM, $0 < s < 1$, and power should flow into the machine via the power winding and out of the control winding.

- When the machine is generating below the base speed for the DFIG or generating above base speed for the BDFRM, $0 < s < 1$, then with the positive slip power flows out of the power winding but in to the control winding. Control of the terminal voltages can do this.

- When the machine is motoring above the base speed for the DFIG or below base speed for the BDFRM, $-1 < s < 0$, then power still flows into the power winding and into the control winding due to the negative slip.

- When the machine is generating above the base speed for the DFIG or below base speed for the BDFRM, $-1 < s < 0$, power flows out of the power winding, and out of the control winding.

These points are general for the doubly fed machines, with issues with the effects of direction of the control winding MMF being addressed for the DFIG and BDFRM. These issues are often not discussed in analyses of doubly fed machines that are grid tied on one winding. Further theoretical studies are required on the rotor angle rotor angle $\delta$ to see if this can be used to control the real power and reactive power flow in the BDFRM although this is not the focus of this work.

We can further refine this by referring the control winding to the grid frequency.
Chapter 2. Direct Torque Control of Brushless Doubly Fed Reluctance Machine

This is shown in Fig 2.12. This is now the familiar equivalent circuit for the induction machine with the rotor circuit left open. The energy conversion occurs in the rotor circuit where

\[
P_{\text{mech out}} = 3 \left( \frac{1-s}{s} \right) |\bar{I}_s|^2 R_s' + 3 \frac{(1-s)}{s} \text{Re}[\bar{V}_s \bar{I}_s^*]
\]

\[
P_s(\text{output}) = 3 \text{Re}[\bar{V}_s \bar{I}_s^*] \quad \text{(Power flowing out of actual control windings)}
\]

This again, means that when \(0 < s < 1\), for motoring operation, power flows out of the rotor circuit for the DFIG but into the rotor circuit for the BDFRM. To make the machine generate then power must flow into the rotor for the DFIG but out of the rotor for the BDFRM. Note that the first term for the mechanical output power on the right-hand side can only ever be positive below the synchronous speed so that the rotor input power must overcome this to generate in the DFIG; for the BDFRM the power flows out of both the stator and rotor. Again, these points are general for the DFIG and BDFRM.

**Figure 2.12:** Doubly-fed per-phase circuit referred to the power winding
Another point that differentiates the circuit in Fig. 2.12 for the DFIG and BDFRM are the relative magnitudes of the circuit components. For a DFIG, $L_{PL}$ and $L_{SL} \ll L_M$ (where $L_M$ is the magnetizing inductance). Indeed, in the standard running light test for calculating the equivalent circuit parameters, $L_{PL}$ is ignored when calculating $L_M$, and in the locked rotor test, $L_M$ is ignored when calculating $L_{PL} + L_{SL}$. However, for the BDFRM, these tests start to break down and $L_{PL}$, $L_{SL}$ and $L_M$ can all have similar magnitudes. In Chapter 4, Table 1, $L_{PL} = 46mH$, $L_{SL} = 87mH$ and $L_M = 62mH$. This issue needs to be investigated to validate the control since it illustrates that the BDFRM has a much more loosely coupled magnetic circuit than the DFIG which may affect the efficacy of the DTC control.

2.5 Conclusions

This chapter has reviewed the basic operation of the scalar control and traditional DTC control for the BDFRM. The traditional DTC when applied to the BDFRM has been addressed in more detail in this chapter because this control method has been shown to be an effective way to control for most types of machines. The traditional DTC provides fast torque and speed response, simple implementation, minimised driver operation, reliability, and robustness against machine parameter variation. However, in Section 2.4, the BDFRM per-phase equivalent circuit was developed. It was used to compare to the DFIG and it is illustrated that the power and control windings are much more loosely coupled than for a DFIG, and the slip is defined differently, with the slip $= -1$ at standstill for the BDFRM, and this warrants the work that is put forward in this study which addresses DTC when applied to the BDFRM.
This chapter is shown the theory for DTC control. However, for application to the BDFRM then further development of the DTC method is necessary and this is put forward in Chapter 3 before the method is examined via simulation in Chapter 4 and experimentally verified in Chapter 5.
Chapter 3

Improved DTC of BDFRM

In Chapter 2, the direct torque control method for the BDFRM was discussed. DTC has a simple structure and has fast torque response. These are major properties so it has received a lot of attention from academia and industry in recent years. The voltage vector selection method is the basic key concept of DTC. When approaching the point of selection, it needs three input parameters: the torque and the flux errors between the estimates and references, and the angular position of the secondary flux. As already stated, it offers simple implementation and fast torque response, as well as better speed control and little dependence on machine parameters. However, there are some issues associated with the DTC method which need to be addressed. These are high torque and flux ripple, current harmonics, variable switching frequency and low performance during transient torque periods. This is because the non-zero voltage vectors are using a look up table with 6 sectors, each spanning 60° [76], and it is not using the zero voltage vector.

The well-known method for reducing the issues related to the conventional DTC method is to use the space vector modulation method. This chapter will discuss
how to apply the space vector modulation to the conventional DTC method and show how to improve the performance of the control. In Section 3.1, the basic concepts of space vector modulation are explained. Section 3.2 shows how it can be applied to the conventional direct torque control method. The control algorithm will also be discussed in Section 3.2.

3.1 Space Vector Modulation

Space vector modulation (SVM) is a well-known technique for reducing torque and flux ripple. This section will introduce the basic concepts of SVM and also discuss the symmetry of the SVM method which is utilized when implemented with the conventional DTC method.

The basic diagram of the voltage source inverter when using SVM is the same as that in Figure 2.7 in Chapter 2. There are three legs in the voltage source inverter and the SVM has eight distinct statuses as shown Figure 3.1. These eight statuses produce six non-zero output voltage states and two zero output voltage states.
The three phase voltages are represented by a reference vector $V^*$ in the d-q plane. It is rotating in the counter-clockwise direction as shown in Figure 3.2(a). This vector will complete one revolution is the same time as one period of the three phase output voltages as shown Figure 3.2(b). The reference vector in Figure 3.2(a) can be made from the vector summation of two close non-zero voltage vectors and two zero voltage vectors with modulation period $T_s$; there are six modulation periods in one rotation. This is expanded upon below.

Let us consider the situation where we wish to drive the machine with the reference vector $V^*$ in sector I as shown in Figure 3.2(a). This vector can be represented by the summation of two non-zero vectors $V_1$ and $V_2$ for appropriate time periods, which
will give direction, with the addition of zero vectors to scale the vector. The modulation algorithm is made up of three steps over the switching period $T_s$ as shown Figure 3.3. The first step applies the non-voltage vector $V_1$ for a period $T_1$. The second step applies a non-zero vector $V_2$ for a period $T_2$ until the voltage has the same phase as $V^*$. The resulting vector till now will be same with the reference vector $V^*$. The last step applies a zero vector for time $T_0 \ (= T_s - T_1 - T_2)$ until time $T_s$ is reached.

This three step modulation algorithm can be represented by the mathematical expression:

$$
\int_0^{T_s} V^* dt = \int_0^{T_1} V_n dt + \int_{T_1}^{T_1 + T_2} V_{n+1} dt + \int_{T_1 + T_2}^{T_s} V_{7,8} dt \quad (3.1)
$$

where $V_{7,8}$ are the zero voltage vectors. The first and second terms in the expression are shown to form the reference vector $V^*$ with two close non-zero vectors and the third expression does not produce any change in direction during the rest of the time $T_0$, although it will scale the vector.
Now we can calculate the time periods $T_1$, $T_2$ and $T_0$. The reference vector $V^*$ amplitude and phase are constant during the $T_s$, (3.1) can be re-written as

$$V^*T_s = V_nT_1 + V_{n+1}T_2$$

(3.2)

since the third term will integrate to zero. Since the reference vector $V^*$ is assumed to exist in sector I ($0 \leq \theta \leq 60^\circ$), as shown as the Figure 3.2, (3.2) can be represented by

$$T_s \vert V^* \vert \cos \theta = T_1(\frac{2}{3}V_{dc}) + T_2(\frac{2}{3}V_{dc})\cos 60^\circ$$

$$T_s \vert V^* \vert \sin \theta = T_2(\frac{2}{3}V_{dc})\sin 60^\circ$$

(3.3)

where $V_{dc}$ is the input DC voltage (the DC rail voltage), $\vert V^* \vert$ is the magnitude of the reference vector $V^*$, and $\theta$ is the phase of the reference vector $V^*$. The time periods $T_1$, $T_2$ and $T_0$ can be calculated using (3.3) so that
Chapter 3. Improved DTC of BDFRM

\[ T_1 = \sqrt{3} T_s \frac{|V^*|}{V_{dc}} \sin\left(\frac{\pi}{3} - \theta\right) \]

\[ T_2 = \sqrt{3} T_s \frac{|V^*|}{V_{dc}} \sin\theta \]

\[ T_0 = T_s - (T_1 + T_2) \]

(3.4)

Hence the time periods for non-zero vectors are used to scale voltage vector as well as obtain the phase angle. Table 3.1 shows the magnitude, phase and switch states of the all eight voltage vectors.

<table>
<thead>
<tr>
<th></th>
<th>$S_a$</th>
<th>$S_b$</th>
<th>$S_c$</th>
<th>$V^*$</th>
<th>$v_{as}$</th>
<th>$v_{bs}$</th>
<th>$v_{cs}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_1$</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>$\frac{2}{3} V_{dc} / 0^\circ$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$-\frac{1}{3} V_{dc}$</td>
<td>$-\frac{1}{3} V_{dc}$</td>
</tr>
<tr>
<td>$V_2$</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>$\frac{2}{3} V_{dc} / 60^\circ$</td>
<td>$\frac{1}{3} V_{dc}$</td>
<td>$\frac{1}{3} V_{dc}$</td>
<td>$-\frac{2}{3} V_{dc}$</td>
</tr>
<tr>
<td>$V_3$</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>$\frac{2}{3} V_{dc} / 120^\circ$</td>
<td>$-\frac{1}{3} V_{dc}$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$-\frac{1}{3} V_{dc}$</td>
</tr>
<tr>
<td>$V_4$</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>$\frac{2}{3} V_{dc} / 180^\circ$</td>
<td>$-\frac{2}{3} V_{dc}$</td>
<td>$\frac{1}{3} V_{dc}$</td>
<td>$\frac{1}{3} V_{dc}$</td>
</tr>
<tr>
<td>$V_5$</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>$\frac{2}{3} V_{dc} / 240^\circ$</td>
<td>$-\frac{1}{3} V_{dc}$</td>
<td>$-\frac{1}{3} V_{dc}$</td>
<td>$\frac{2}{3} V_{dc}$</td>
</tr>
<tr>
<td>$V_6$</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>$\frac{2}{3} V_{dc} / 360^\circ$</td>
<td>$\frac{1}{3} V_{dc}$</td>
<td>$\frac{2}{3} V_{dc}$</td>
<td>$\frac{1}{3} V_{dc}$</td>
</tr>
<tr>
<td>$V_7$</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0 / 0$^\circ$</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>$V_8$</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0 / 0$^\circ$</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

Table 3.1: Inverter switching states and output voltages

Sometimes the reference vector $V^*$ is found to be outside of the six voltage vector hexagon shape as shown in Figure 3.2(a), this is called over modulation. The numerical over modulation condition is $T_1 + T_2 > T_s$. If the reference vector $V^*$ does not exist inside the hexagon, only the two non-zero vectors are used to form the over
modulation reference vector $V^*$. Thus, the time periods are recalculated as

\[ T_1' = \frac{T_1 T_s}{T_1 + T_2} \]
\[ T_2' = \frac{T_2 T_s}{T_1 + T_2} \]
\[ T_0' = 0 \] \hspace{1cm} (3.5)

During the switching period $T_s$, the voltage vectors can be applied any order, and the reference vector $V^*$ will have the same magnitude and phase. However, the harmonics, switching frequency and current ripple will be affected by the order of the applied voltage vectors. An important key feature is the process for deciding the switching sequence of the voltage source inverter and this related to the non-zero vector location during the switching period.

Figure 3.4 shows how the switching sequence affects the ripple current and the switching frequency. Figure 3.4(a) and 3.4(b) look similar, but Figure 3.4(a) has a smaller ripple current and faster switching frequency than Figure 3.4(b). The smaller ripple current also means it has the lowest total harmonic distortion (THD). From the results of Figure 3.4, it can be seen that a symmetry in the switching waveform over the switching period is a good technique for reducing ripple current and minimizing the THD. This technique is called Symmetrical SVM.

In Figure 3.4, it can be observed that to get a central location for the non-zero voltage vectors, the zero voltage vector has to divided into two and applied for time $\frac{T_0}{2}$ before and after the non-zero voltage vectors. For example, the reference vector $V^*$ is in sector I, for this vector, the switching pattern will be in the sequence $V_8$ (000) $\rightarrow$ $V_1$ (100) $\rightarrow$ $V_2$ (110) $\rightarrow$ $V_7$ (111). In the next modulation period, the switching
Figure 3.4: Ripple current with regards to the position of the non-zero voltage vector pattern is in the opposite sequence as shown Figure 3.5.

Figure 3.5: Optimal switching pattern of SVM

Hence, two adjacent sequences (the On Gate Sequence and the Off Gate Sequence) are joined to form one switching period. Furthermore, symmetrical switching can also reduce the number of switches in one period. If successive switching vectors have one leg changing state, or no changes in leg state, then there is less switching
stress and losses. This criterion can be used to decide on whether to use $V_7$ or $V_8$ for the zero vector. It can be seen in Figure 3.5 that each leg only changes state twice over the $2T_s$ switching period.

Figure 3.6 shows all switching patterns in the six sections and these can be set as a pre-calculation. It can be seen that all the switches in the voltage source inverter will be switched twice only (on and off) during the one period as discussed before. The section vectors are controlled using these sequences. The magnitudes and angle of the section vectors are controlled by the non-zero vector switching periods relative to each other and to $T_s$ as described earlier.

![Figure 3.6: All switching patterns](image)
3.2 Improved DTC method for BDFRM

Figure 3.7 shows a block diagram of the proposed DTC method. Instead of the switching selection method in the traditional DTC, this thesis proposes the use of the SVM module using the symmetry switching technique for reducing the current ripple and faster switching frequency as discussed the previous section in this chapter. The symmetrical SVM technique also reduces the THD and torque ripple, and provides a fixed switching frequency for the inverter.

![Control block diagram of the proposed DTC method](image)

**Figure 3.7:** Control block diagram of the proposed DTC method - this is a modified diagram when compared with Fig 2.5

In Figure 3.7, the reference torque $T_e^*$ can be obtained from the difference between the actual speed $\omega_r$ and the reference speed $\omega_r^*$. The torque difference $\Delta T_e$ can
be obtained from the difference between the reference torque $T_e^*$ and the estimated torque $T_e$. The $q$-axis voltage $v_{qs}$ which is used as one of inputs to the SVM module is obtained from the torque PI controller which has torque difference $\Delta T_e$ as the input. The $d$-axis voltage $v_{ds}$ can be obtained in a similar fashion from the flux PI controller where the flux difference $\Delta \lambda_s$ is the input, and this is the difference between the reference flux $\lambda_s^*$ and the estimated flux $\lambda_s$. The reference vector $V^*$ phase is $\theta$ and this can be estimated from the flux calculator.

The SVM module can select one of the voltage vectors using $v_{ds}$, $v_{qs}$ and $\theta$ and calculate the time periods $T_1$, $T_2$ and $T_0$. Finally, the voltage source inverter switching signals are generated by the SVM module, and these are used to control the 2-level inverter to drive the BDFRM.

Figure 3.8 shows the flowchart for the SVM modulation. This is the basic control algorithm for the proposed DTC method in this thesis. At the beginning of the switching period, the phase of the reference vector $V^*$ will be calculated by using the two BDFRM references. It will select the sector of reference vector then find the amplitude of the reference vector $V^*$.

After finding the amplitude, it is possible to calculate the time periods $T_1$, $T_2$ and $T_0$ using (3.4). The next step is to inspect the value of $T_0$ to see if it greater than zero. If it is greater than 0, all values have been calculated for selecting the correct voltage vector and generating the actual switching signals for the voltage source inverter. If $T_0$ is less than 0, this means $T_1 + T_2$ is greater than $T_3$, so that over modulation is occurring. $T_1$ and $T_2$ will be recalculated using (3.5). After recalculating, the switching signals for the voltage source inverter can be generated. This sequence takes place for every switching period.
3.3 Conclusions

It has been shown in this chapter that modifications can be made to the traditional DTC strategy using SVM to improve its effectiveness in terms of controlling BDFRM. The fundamental SVM-DTC algorithm has been modified for application to BDFRM. The existing DTC as applied to an induction machine is the basis for developing the algorithm for the BDFRM; however, as illustrated in Chapter 2, the
Chapter 3. Improved DTC of BDFRM

BDFRM is a different machine with different characteristics. The reason for considering the SVM-DTC method for this particular machine is to improve its transient response. The studies of SVM-DTC algorithm applied to electrical machines such as induction machines and permanent-magnet synchronous machines have been extensively published[77, 78]. However, this control method for the BDFRM has not been studied in the literature so it is important to apply it both in simulation (Chapter 4) and experiment (Chapter 5).

The main contribution of this chapter is the development of the SVM-DTC scheme for the BDFRM. The traditional DTC for the BDFRM has been shown to have high torque ripple, current harmonics, variable switching frequency and low performance during transient torque periods. The Space Vector Modulation is well-known method to reduce ripple and harmonics using a fixed switching frequency and is therefore applied to the DTC method.

The performance of the proposed theory will be discussed and confirmed by both simulation and experimental work in Chapters 4 and 5.
Chapter 4

Simulation Studies of the Proposed DTC

In this chapter a computer simulation analysis is carried out in order to analyse the method and to probe and show the performance of the proposed theory before testing on an experimental prototype. Using the computer simulation, the performance can be compared to existing controls and any potential performance improvement assessed. There are numerous computer simulation tools available, but perhaps the most flexible and one of the most popular computer simulation tool is Matlab/Simulink®. This programme consists of functional blocks from the libraries of Matlab/Simulink® and this feature provides a flexible simulation environment. It was decided to use this simulation tool and show the results in this chapter.

There are a large number of research papers that describe the results of traditional and improved DTC methods using Matlab/Simulink®. This work focuses on the simulation for the proposed DTC operation. To obtain accurate simulation results for the proposed control method for BDFRM, it is important to use an accurate
machine model. In this chapter the ideal dynamic simulation model of the BDFRM will be explained which is couched in terms of the basic mathematical analysis from Chapter 1. It is designed with Matlab/Simulink® control blocks. The proposed DTC method is simulated and the results of the performance simulation are presented in this chapter.

The control block diagram of the proposed DTC was shown as Figure 3.7 in Chapter 3. The diagram is divided into three basic parts: the BDFRM model; the control block; and the controlled inverter. The control theory of the conventional DTC and the proposed DTC method were fully discussed in Chapters 2 and 3, so the methods are not reviewed in this section.

### 4.1 Ideal Simulation Model of the Brushless Doubly Fed Reluctance Machine

The aim of this section is to help with the correct understanding of the BDFRM. The basic mathematical equations of the BDFRM are necessary in order to analyse and simulate the machine. Using (1.1) in Chapter 1, the basic mathematical equations can be rewritten as

\[
\begin{align*}
v_{pd} &= R_{p}i_{pd} + \frac{d}{dt}L_{p}i_{pd} + \frac{d}{dt}L_{m}i_{sd} - \omega \lambda_{pq} \\
v_{pq} &= R_{p}i_{pq} + \frac{d}{dt}L_{p}i_{pq} - \frac{d}{dt}L_{m}i_{sq} + \omega \lambda_{pd} \\
v_{sd} &= R_{s}i_{sd} + \frac{d}{dt}L_{s}i_{sd} + \frac{d}{dt}L_{m}i_{pd} - (\omega_{r} - \omega)\lambda_{pq} \\
v_{sq} &= R_{s}i_{sq} + \frac{d}{dt}L_{s}i_{sq} - \frac{d}{dt}L_{m}i_{pq} + (\omega_{r} - \omega)\lambda_{sd}
\end{align*}
\] (4.1)
These will help build the ideal BDFRM model for a Matlab/Simulink® block. The primary and secondary d-q current equations can be expressed in a manner suitable for use in a Matlab/Simulink® where

\[
\begin{align*}
    i_{pd} &= \frac{1}{L_p} \int (v_{pd} - R_p i_{pd} + \omega \lambda_{pq}) - L_m i_{sd} \\
    i_{pq} &= \frac{1}{L_p} \int (v_{pq} - R_p i_{pq} - \omega \lambda_{pd}) + L_m i_{sq} \\
    i_{sd} &= \frac{1}{L_s} \int (v_{sd} - R_s i_{sd} + (\omega_r - \omega) \lambda_{sq}) - L_m i_{pd} \\
    i_{sq} &= \frac{1}{L_s} \int (v_{sq} - R_s i_{sq} - (\omega_r - \omega) \lambda_{sd}) + L_m i_{pq}
\end{align*}
\]

(4.2)

The machine torque can be calculated using (2.2) in Chapter 2. This equation can be written as

\[
    T_e = \frac{3}{2} p_r L_m (i_{sd} i_{pq} + i_{sq} i_{pd})
\]

(4.3)

where \( p_r = p_p + p_s \). The speed of the machine can be obtained from

\[
    J \frac{d}{dt} \omega_r + K_d \dot{\omega}_r = T_e - T_L
\]

(4.4)

and this can be re-arranged using \( \omega_r \) in the simulation blocks so that

\[
    \omega_r = \int \left[ \frac{1}{J} (T_e - T_L - K_d \dot{\omega}_r) \right]
\]

(4.5)

Therefore, based on the mathematical model defined by (4.2), (4.3) and (4.5), the ideal simulation model of the BDFRM can be developed and this is shown in Figure...
4.1.

**Figure 4.1:** Ideal simulation model of the BDFRM for Matlab/Simulink®
As previously discussed, the BDFRM has two windings supplying power to the machine. These are primary or power windings, which are connected to the grid, and the secondary or control windings, which connected via an inverter and have the additional role of controlling the machine. Using the basic machine parameters as given in Table 4.1, the machine was simulated. The machine parameters were obtained from modified running light and locked rotor tests as done on induction machines. They had to be modified because the leakage inductances are of a similar magnitude as the magnetizing inductance. This was discussed in Section 2.4. The rated BDFRM current is 30 A. The inertia is very low because this is to test the robustness of the control. The inertia was estimated for simulation as 3.2 kgm$^2$ and further calculation work was carried out in Chapter 5 in order to calculate the torque. The inertia of the system can be difficult to assess because the combined drive and load inertia needs to be included. Increased inertia will slow the system response to speed change.

The machine requires speed feedback not only for the speed terms in impedance matrix in (1.1) in Chapter 1 but also for the synchronization. The primary can be short circuited to produce an induction machine operating mode (as discussed in Section 2.1) but to obtain true synchronous operation through the control winding then speed feedback is required in order to set the correct frequency and phase. This is shown in Figure 4.1. It can be seen in (4.3) that with no secondary current there should be no torque. Even with secondary current, if the winding currents and speed are not synchronized then the torque should simply oscillate.

The synchronizing requirement of the machine was discussed in Section 1.4. To gain
<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pole Contribution</td>
<td>( p_p = 8, p_s = 4, p_r = 6 )</td>
</tr>
<tr>
<td>Primary Resistance</td>
<td>1.2 ohm</td>
</tr>
<tr>
<td>Primary Inductance</td>
<td>46 mH</td>
</tr>
<tr>
<td>Secondary Resistance</td>
<td>0.9 ohm</td>
</tr>
<tr>
<td>Secondary Inductance</td>
<td>87 mH</td>
</tr>
<tr>
<td>Mutual Inductance</td>
<td>62 mH</td>
</tr>
<tr>
<td>Grid Power Supply</td>
<td>240V / 50 Hz</td>
</tr>
<tr>
<td>Secondary Power Supply</td>
<td>100V / 50 Hz</td>
</tr>
</tbody>
</table>

**Table 4.1: Simulation parameters**

an insight into the frequency and speed relationship then taking the high pole winding to be the power winding, there is a synchronizing requirement for the inverter-fed control winding [76]

\[
f_c = P f_f \pm f_p
\]  

(4.6)

where \( f_c \), \( f_r \) and \( f_p \) represent the control winding electrical frequency, rotor mechanical rotational frequency and grid (power winding) frequency. This is a repetition of (1.1) but it is important establish the synchronization requirement.

For the 4-8 stator pole machine (which is used here), if the machine is in induction motor mode with the secondary shorted, then the speed will be close to 500 rpm. Figure 4.2 shows the speed and control winding frequency relationship. This
is development of Figure 1.3 and shows the speed down to zero. It can be seen that the control winding at zero speed is the same as the grid frequency. The negative frequency means that the phase rotation is reversed.
Figure 4.2: Variation of control winding frequency with speed when the power winding connected to 50 Hz grid. Two machines (2-6-4, 4-8-6) are displayed

4.2 Simulation in asynchronous induction motor mode

The machine model is implemented as a simulation for the simple case where the secondary windings are shorted with the primary windings grid connected. Figure 4.3 shows the torque and speed against time for the short circuit control winding scenario. This shows a simple asynchronous speed run up in a similar manner to an induction motor direct on-line start. It will run up to just below 500 rpm which is the point where the secondary current would be DC. At this point the primary will not induce a voltage into the secondary so there will be no secondary current and hence no torque.

According to the relationship between the control winding frequency and the machine speed in Fig 4.2, when the control winding has zero frequency, the machine speed should be reached to 500 rpm. However, when the machine is operating in asynchronous mode with shorted control windings, there is no EMF induced into the control windings at 500 rpm. However, the simulation has no load at steady-state so
once this point is reached in the simulation the machine is running at 500 rpm with no torque generation, i.e., there is no control current. The simulation results in Fig 4.3 shows that the machine speed has reached 500 rpm when the control winding has no current. This is validated well the simulation model of BDFRM under asynchronous induction motor mode. In a real machine, there is always some loading so the machine would be operating just below 500 rpm.

![Graphs showing speed and torque transient run-up for shorted control winding](image-url)

**Figure 4.3:** Speed and torque transient run-up for shorted control winding
4.3 Simulation with simple open-loop secondary frequency control

If the machine has 50 Hz on both windings, then the speed will be 1000 rpm and this is shown in Figure 4.4. Since this is double the previous speed then the time to reach this speed is longer. To run dynamically up to this point then the control winding has to start at -50 Hz (i.e., backwards rotation) then then run up to this speed by varying the frequency. It appears to do this successfully [79]. This is open-loop control with the rotor being pulled into synchronism at start. It can be seen that there is substantial transient torque oscillation.

To compare with Fig 4.2, when the control winding has 50 Hz frequency the machine speed should be 1000 rpm. Thus, these simulation results in Fig 4.3 and Fig 4.4 help validate the BDFRM simulation model developed in Matlab/Simulink®.
Figure 4.4: Results for 50 Hz secondary control winding – run-up from -50 Hz to 50 Hz
4.4 Simulation Results of the Proposed DTC

This section puts forward the simulation results for the proposed DTC developed in this thesis. The functional block diagram for the system is shown in Figure 3.7. As mentioned earlier, the dynamic model of the BDFRM has been used in the simulation and the control theory, which is described in Chapter 3, has been applied.

For much simpler and more convenient numerical analysis of the machine dynamics, a Parks transformation is used in the simulation. The Parks transformation is a well-known method for obtaining a rotating reference frame from the stationary reference frame that is used by the DTC. Figure 4.5 shows the relationship between the three reference frames: the abc frame, the stationary frame, and the rotational frame. The first conversion from the abc frame to the stationary frame is called the Clarks transformation and the second conversion from the stationary frame to the rotating frame is called the Parks transformation.

\[
\begin{bmatrix}
  f_a \\
  f_b \\
  f_c 
\end{bmatrix} = \frac{2}{3} \begin{bmatrix}
  \cos \theta & \cos(\theta - \frac{2}{3} \pi) & \cos(\theta + \frac{2}{3} \pi) \\
  -\sin \theta & -\sin(\theta - \frac{2}{3} \pi) & -\sin(\theta + \frac{2}{3} \pi)
\end{bmatrix} \begin{bmatrix}
  f_a \\
  f_b \\
  f_c 
\end{bmatrix}
\]

\[
\begin{bmatrix}
  f_a \\
  f_b \\
  f_c 
\end{bmatrix} = \begin{bmatrix}
  \cos \theta & \sin \theta & f_a \\
  -\sin \theta & \cos \theta & f_b \\
  0 & 0 & f_c
\end{bmatrix}
\]

**Figure 4.5:** Converting process among the three reference frames
Figure 4.6 displays the speed and torque response of the machine for a speed change from 0 rpm to 50 rpm using the proposed DTC method and Figure 4.7 shows the same results with the conventional DTC method for comparison. When comparing the results, the low-speed torque ripple is reduced using the proposed theory. This illustrates the improvement in performance as discussed earlier. This simulation is at low speed where many drive systems struggle to operate smoothly.
Figure 4.6: Simulation responses for a speed change from 0 to 50 rpm with the proposed DTC
Figure 4.7: Simulation responses for a speed change from 0 to 50 rpm with the conventional DTC.
Figure 4.8 and Figure 4.9 show a comparison in results for the two methods with a speed change from 0 rpm to 200 rpm, then to 400 rpm. The results show that the speed response and torque ripple of the proposed DTC again give improved performance when compared to the conventional DTC. This time this is for the mid speed range.
Figure 4.8: Simulation responses for a speed change from 200 to 400 rpm with the proposed DTC
Figure 4.9: Simulation responses for a speed change from 200 to 400 rpm with the conventional DTC
Figure 4.10 and Figure 4.11 show a comparison of results for a speed change from 0 rpm to 800 rpm. This is a big step change. These results confirm that the proposed DTC has improved performance in high speed range too. One point to note is that the speed goes through the synchronous speed point at 500 rpm where the control current goes through the 0 Hz value. This could be a point of poor control however the results seem good with no discernible variance in the control.
Chapter 4. *Simulation Studies of the Proposed DTC*  

**Figure 4.10:** Simulation responses for a speed change from 0 to 800 rpm with the proposed DTC
Figure 4.11: Simulation responses for a speed change from 0 to 800 rpm with the conventional DTC
Fig 4.12 shows a speed response for changes in load conditions. This is carried out at 400 rpm which is a mid-speed point. Initially there is no load, then 10 Nm is applied and then -10 Nm is applied. This result confirms that the proposed DTC has good performance in steady state with load variations.

**Figure 4.12**: Simulation response for a load torque change from 0 to 10 Nm to -10 Nm with the proposed DTC
Figure 4.13 shows the currents for the primary and secondary windings when the speed changes from 0 rpm to 800 rpm. This is when using the proposed DTC method. For comparison, Figure 4.14 displays the same currents over the 0 to 800 rpm speed change with using the conventional DTC method. These results clearly show that the current ripple is reduced when using the proposed DTC, rather than the conventional DTC, which illustrates the improved performance of the control method.

Based on the computer simulation results put forward here, the proposed DTC algorithm has been shown to have improved performance compared to the conventional DTC method. Across all speed ranges, the machine has better characteristics of speed response and torque ripple.
Chapter 4. Simulation Studies of the Proposed DTC

Figure 4.13: Primary (top) and secondary (bottom) winding currents for a speed change from 0 to 800 rpm with the proposed DTC
Figure 4.14: Primary (top) and secondary (bottom) winding currents for a speed change from 0 to 800 rpm with the conventional DTC.
4.5 Conclusions

Using the proposed theory in the previous chapter, a computer model was developed and simulation studies carried out in the Matlab/Simulink® environment. These simulations validate the performance of the proposed DTC when applied to the BD-FRM machine under study in this work.

All the simulation function blocks have been designed and implemented using the system diagrams given in the previous chapter, for instance, Fig 3.7. The control was based on a d-q model. The BDFRM was also has been implemented in the simulation as a grid-connected machine with shorted control windings. This is asynchronous, or induction machine operation, from 0 to 500 rpm for a 50 Hz power winding connection. A further simulation ran the machine in open-loop mode with it running up with a variable frequency supply applied to the control winding from -50 Hz (0 rpm) to 50 Hz (1000 rpm). These simulations allowed benchmarking of the control and illustrated that the control is necessary for correct operation.

From the all results, it can be concluded the proposed DTC theory for the BD-FRM has been successfully applied in simulation. The proposed DTC is clearly shown to be effective with higher accuracy than the conventional DTC. The real-time experimental system will be discussed in the following chapter; the results will be compared to the simulated results.
Chapter 5

Experimental Results

The proposed DTC method has been successfully demonstrated through the simulation results in Chapter 4. The simulation results from Chapter 4 are validated by experiment in this chapter. This chapter will describe the experimental test system then put forward several experimental test results of the proposed algorithm.

The experimental test system consists of three major components: the prototype BDFRM, the controller, and the 3-phase voltage source inverter. The experimental results have been generated by a 20 kW prototype BDFRM as illustrated in chapter 1. The prototype ducted-rotor BDFRM has 4-pole and 8-pole 3-phase windings and a 6-pole ducted rotor. Both the rotor and stator are laminated. This machine is used for the experiments. More details of the machine are given in Appendix A. The primary winding of the machine’s stator is directly connected to a conventional 3 phase power supply with fixed voltage and frequency. The secondary winding of the machine’s stator is connected to the voltage source inverters.

The control algorithm for the control in this work is implemented in a real time
system using a Texas Instrument DS 28335 board with C++ code. The advantage of using this DSP board is that it provides a flexible control algorithm with common C++ code, which can supervise and control in real time through its interface with different variables. The TI DSP module calculates the torque and flux estimation then generates the IGBT gate signals.

The 3 phase voltage source inverter module is a Semicron SKHI 22A IGBT inverter kit with a three phase diode bridge rectifier, safety isolation, and gate driver circuits. This commercial inverter kit is provides a straightforward and safe experimental environment for controlling the machine. More detailed information of hardware is provided in Appendix B.

5.1 Experimental Test Bench Design

The experimental platform and the actual test bench setup are shown in Figures 5.1 and 5.2, B.4 in Appendix B. The experimental bench consists of a power grid, a circuit breaker and a pre-charge resister (for safety reasons), a 3-phase voltage source inverter and a driver circuit kit, a TI DSP module and its control system, a load DC machine, and the prototype BDFRM.

The prototype machine primary winding is directly connected to the three phase power grid through the circuit breaker and the secondary winding is connected to the 3-phase voltage source inverter which is controlled by the TI DSP module. The
DSP module is connected to a workstation and controlled by a C++ software platform. The prototype BDFRM is controlled by the proposed DTC algorithm under a load torque given by the coupled DC machine. Transient speed changes are used to test the algorithm where the DC machine also provides an additional inertia.

The experimental platform is slightly different from the simulation analysis model. Instead of an estimation module for the machine torque and speed, the torque and speed are directly measured by the sensor. The additional signal for measuring the inverter DC link voltage was only for monitoring and protection purposes.

The microcontroller of the DSP module utilizes the TI TMS320F28335 DSP. This DSP controller provides a 32-bit CPU and the maximum despatch cycle time is 6.67 ns. It has an IEEE-754 single-precision floating point unit for calculating of floating numbers, 16 channels of 12-bit analogue-to-digital converters (ADCs) with
80 $\mu s$ conversion rate, and 6 high resolution pulse width modulation (HRPWM) output channels with 150 $\mu s$ MEP resolution. This DSP controller is sufficient for use as the real time controller. When using this control board, all the measurement signals are calculated and the control signals for the 3-phase inverter are generated by the proposed algorithm. Figure 5.3 shows the basic functional diagram of the TMS320F28335.

The 3-phase inverter and driving circuits was realised using a Semicron SKHI 22A IGBT inverter kit. This kit is made up from a 3-phase rectifier module, an IGBT module, and an IGBT driver module. The IGBT drivers use three SKHI 22A cores, each one controlling an inverter leg. The top and bottom IGBTs are series-connected and form one single IGBT module. The maximum switching frequency of the driver is 50 kHz. It also has an internal 4 $\mu s$ interlock dead-time between the switching of the top and bottom IGBTs in the same leg. The 3-phase diode bridge rectifier module has a line voltage rating of 440 Vrms and a line current rating of 30 Arms.
There are three IGBT modules and each of these is an inverter leg, and formed from two series-connected IGBTs with anti-parallel diodes. Each IGBT module has a rating of 600 Vdc and a current rating of 30 A. The control signals are delivered by an SKHI 22A which is an IGBT driver. This is connected through an additional gate resistor whose value is 30 Ohms. The purpose for this resistor is to limit the gate charging current, and consequently, this limits the collector-emitter over voltages due to parasitic inductance.

All control modules are shown in Figure 5.4 such as the DSP controller board, the
circuit breaker and the pre-charge resistors module and the Semicron SKHI 22A IGBT inverter kit. The controller software has been developed using C++ under Code ComposerTM Studio V6 environment provided by TI.

### 5.2 Experimental Results – Step Changes in Speed

This section puts forward the experimental results from the test system in order to verify the proposed algorithm. The experimental results are presented in the same order as the simulation results for clarity. Firstly, the response to a low speed change will be given. A mid speed step change of the machine will then be presented in order to test the control, and finally a high speed step change from zero speed is given.
The proposed DTC algorithm was applied to the prototype BDFRM system. The performance of the machine in terms of the speed step changes described above can be observed in Figs. 5.5 to 5.7. The successful speed control of the BDFRM is shown in Fig. 5.5 during a reference speed change from 0 to 50 rpm, which is in the low speed range. The machine speed reached the reference speed in less than 0.4 s from standstill. Fig. 5.6 shows the mid speed range response of the machine, just below the synchronous speed. The reference speed changes from 0 to 200 rpm first then it changes again from 200 rpm to 400 rpm after 3 s. The middle range machine speed was successfully changed from 200 rpm to 400 rpm. The high-speed step change is illustrated in Fig. 5.7. The reference speed was changed from 0 to 800 rpm. The machine speed reached the reference speed 800 rpm in less than 1 s and this goes through the synchronous speed point.

As can be seen, the rotor speed changes in a linear fashion in all cases with no overshoot and no oscillation. The target speed is reached in all cases. It can be summarized that all speed waveforms are clear and change smoothly, thus illustrating the good control of the machine. The load for these machines takes the form of the system inertia during the speed change. The results are similar and some calculations are put forward in the next section.
Figure 5.5: Low speed response from 0 to 50 rpm (1 s per horizontal division and 10 rpm per vertical division)

Figure 5.6: Medium speed response from 200 rpm to 400 rpm (with the initial step from 0 to 200 rpm – 1 s per horizontal division and 50 rpm per vertical division)
Figure 5.7: High speed response from 0 to 800 rpm (1 s per horizontal division and 200 rpm per vertical division)
5.3 Comparison of Experimental and Simulations

The inertia for the simulations was set to 3.2 kgm$^2$. For the measurements, it was not possible to use the torque transducer because the transient acceleration is used to assess the performance and the load is the system inertia. Therefore, simple timed speed changes can be used.

If load torque during the speed transient is due to the inertia, then

$$T_L = J \frac{d\omega}{dt} = J \frac{\Delta \omega}{\Delta t} |_{\text{transient run-up}}$$  \hspace{1cm} (5.1)

This ignores the friction and windage. The linear speed changes indicate this is reasonable. In Figure 4.10(a) the rise time from 0 to 800 rpm is about 0.85 seconds in the simulation. Therefore,

$$T_L = J \frac{\Delta \omega}{\Delta t} |_{\text{transient run-up}} = 3.2 \times \frac{800 \times 2\pi}{0.85 \times 60} = 315 \text{Nm}$$  \hspace{1cm} (5.2)

This simple calculation matches well the simulation torque curve. For the experimental machine, if we use the same transient simulation and assume approximately 315 Nm of torque, then, for a 1 s run up time as illustrated in Figure 5.7 the system inertia $J$ can be approximated to

$$J = \frac{T_L}{\frac{\Delta \omega}{\Delta t}} = \frac{315}{\frac{800 \times 2\pi}{1 \times 60}} = 3.76 \text{kgm}^2$$  \hspace{1cm} (5.3)

This is for the BDFRM and the DC load machine which are mechanically coupled. The machine appears to be able to deliver over 300 Nm of torque in a controlled manner. The maximum control winding current is 30 A which the machine can
sustain in steady state. This is an approximate calculation and further calculation can be done using an approximation of the inertia and this is given below. However, what can be seen is that the torque is almost linear and goes through the synchronous speed point (500 rpm), where the control winding current is DC, which illustrates a good control. The slope above 500 rpm is steeper than below 500 rpm showing there is more torque. This could be due to the fact that below 500 rpm both windings have power flowing into them (the slip is negative as discussed in Chapter 2), above 500 rpm the power flows out of the control winding (slip is positive). This reversal will increase the back-emf in the control winding for a given terminal voltage, which is being controlled, since the voltage across the winding resistance and leakage inductance is reversed, thus increasing the torque production. This is one possible explanation.

Another explanation is that the inverter is not bidirectional because the grid is rectified through a diode bridge rectifier. This will lead to the DC voltage capacitor being pumped up between 500 and 800 rpm. Once 800 rpm is reached the power being generated will be dissipated by winding and inverter losses since there is no mechanical load. For full DFIG systems it is normal to have back-to-back inverters. In this prototype system, the DC link for the control winding inverter is supplied by a diode bridge rectifier.

An alternative method can assess the inertia of the system shown in Fig. 5.7. From the dimensions of the rotor and assuming 8000 kg/m³ for the steel, then the inertia is about 1 kgm² for the BDFRM. The DC machine was attached to give an inertia weight and this is estimated to be a little higher at 1.2 kgm² since the machine is about the same size but it will have the commutator. The shaft is long and there are large couplings and a torque transducer. This is estimated to be about 0.5 kgm² so
that the total inertia is estimated to be about 2.7 kgm$^2$ so that torque is estimated to be 226 Nm. This is an estimation but again the system shows good performance.

Therefore, the torque density can be calculated: the diameter of the stator is 420 mm and the axial length is 251 mm. The volume is therefore 0.0348 m$^3$ which gives a torque density of 6.50 to 8.62 kNm/m$^3$, using the estimates of the torque from above.

### 5.4 Conclusions

A suitable system was constructed to investigate the control strategies developed in this thesis. The machine was a bespoke machine specially built for a large project related to the BDFRM. This study was part of a larger project to investigate its operating characteristics. This chapter highlights the system developed and further details of the machine and system are given in the appendix.

The Proposed Direct Torque Control was applied to the prototype BDFRM. The experimental results shown in this chapter demonstrated and verify the proposed control method for a set of step changes in speed. These underpin the simulation results in Chapter 4.

The torque response is one of the important characteristics that verifies the proposed DTC algorithm. However, the step change in speed used the inertia if the system to load the machine and this also allowed the control to be tested with transient response. To estimate the torque, indirect methods were used to assess the torque and these calculations illustrate good torque production and shows that the
modified DTC can deliver the demanded torque.

The machine speed was successfully changed for 0 to 800 rpm within 1 s and this passes through the synchronous speed point, where the control winding frequency is zero. There was a change in slope of the speed change at about this point. This was not shown in the simulation result in Fig. 4.8(a). Reasons for this are put forward and these are related to the regeneration action in the inverter.
Chapter 6

Conclusions

An improved Direct Torque Control algorithm for the BDFRM has been proposed in this thesis. The basic objective of this research was to improve the conventional DTC method for the BDFRM and this is the main claim to originality of this work. To achieve this objective, a comprehensive review of the conventional DTC algorithm was carried out and a literature of the BDFRM was provided in Chapter 2. The main concepts of the proposed theory, which is developed in order to overcome the disadvantages of the conventional DTC, was discussed in Chapter 3 and verified through both computer simulation and experimental verification in Chapters 4 and 5. However, further works are still remained to be carried out. This chapter will conclude the proposed theory and discuss possible future research.
6.1 Conclusions

The conventional DTC method has some well-known advantages such as simple implementation, fast torque and speed response, and independence from machine parameters. However, despite these advantages, the conventional DTC method also has some drawbacks including high torque and speed ripple, variable switching frequency, and poor performance during the transient periods.

Thus, in this thesis, an improved DTC algorithm has been proposed in order to overcome the disadvantages of the conventional DTC method. The main focus of the work is to adapt the Space Vector Modulation method to the control problem instead of the vector selection method. SVM is a well-known method which can reduce the torque and speed ripple, and provide a fix to the problem to the variation of switching frequency and possible over-frequency. The proposed theory is implemented with simple modification, so that all the advantages of the conventional DTC, such as easy implementation and fast response, still remain, but the issues concerned with torque and current ripple are reduced.

The proposed theory has been verified as giving improved performance. This was done by comparing the simulation results of the conventional DTC algorithm and the proposed algorithm. It has been shown to significantly reduce the torque and speed ripple, and improve the speed response during the transient period. For verification of the simulation results, the proposed theory and the real time control system have been implemented using the TI DSP module and the prototype ducted laminates 4-8-6 pole BDFRM. The experimental results show good performance, thus validating the proposed DTC theory.
6.2 Suggestions for further work

The control method was verified by investigation of speed step changes. However, due to limitations of the hardware system, some future study still remains with regards to the proposed theory. The most important aspect is to measure the torque response dynamically with the speed when there is a speed step change, and compare to the simulations. The second issue is the study of the steady-state speed response characteristic due to changing load conditions. This will demonstrate a robust control even when the load varies. Both of studies will support the proposed theory.

There is no doubt that the BDFRM has several advantages, with a possibility of being used in the next generation of electrical machine. Compared with the induction machine, the BDFRM has could have the possibility of being smaller in size with higher power density, higher efficiency and lower maintenance. The doubly-fed induction machine has been widely used in the wind power generation. Thus, the one of the most important application areas for the BDFRM should be wind power generation area. This work has focused on the control aspects as a motor, but the generating mode of the BDFRM will be an important direction for further research.
Bibliography


Appendix A

Prototype BDFRM Design

A.1 Winding Specification for BDFRM

Above details are the winding specification for prototype BDFRM which is used in this paper.

- slot stator = 12 slot per pole for 4 pole and 6 slots per pole for 8 pole

- Slots to be lined and liner between windings

- Two separate single-layer lap-wound 3-phase windings to be marked A41, A42, B41, B42, C41, C42, A81, A82, B81, B82, C81, C82 to indicate phase, winding set, and polarity.

- 24 coils per winding set = 8 coils per phase to be connected in series.

- For 4 pole winding: coil pitch = 11 slots; coils per pole per phase = 2; phase offset = 8 slots
• For 8 pole winding: coil pitch = 5 slots; coils per pole per phase = 1; Phase offset = 4 slots

• Maximum clearance: 100 mm (can be taken up to 120 mm if necessary)

• Wire diameter: 1.65 mm (AW15) with 18 turns per coil. Slot fill = 0.4.

• Insulation class F.

• Resin impregnation not required.

A.2 Stator and Rotor Specification

Figure A.1: Stator Dimensioning
Figure A.2: Rotor Dimensioning
Appendix A.
Prototype BDFRM Design

Figure A.3: Stator Drawing

Stator Retaining Rings, and Compression and Alignments Rods for Brushless Doubly-fed Machine

Lam will be located within a frame formed from one ring and rods. Four rods may be insufficient so six or eight may be needed.

Diameter to slot bottoms = 345.8 mm
Stator core ID = 272.8 mm
Stator core OD = 420 mm
Diameter of centre of M10 rods = 430 mm

Retaining Ring Quarter Section
4 x 18 mm retaining rods
45 mm
25 mm
15 mm
recess
12.5 mm

Thickness of retaining rings (set here are 25 mm) can vary depending on material available (formed from steel plate)

Stack length should be closest to 250.3 mm when compressed

Compression and alignment rods

Retaining ring holes to be tight fit for compression rods and rods to be perfectly perpendicular. Ends turned down with M16 nuts to give compression.

The air gap is only 0.65 mm and there is no keyway in outer radius of stator so careful alignment is important.

Stator Lamination
Stator core back = 210 - 172.8 = 37. mm
Figure A.4: Rotor Mounting
Figure A.5: Bed Plate and Stator Mountings

End supports welded and brazed

Bottom of stator retaining ring sits on 100 mm square, 10 mm thick pads bolted on to ring and into plate

End supports are 40 mm wide and 20 mm thick steel

Mounting plate 800 mm × 520 mm, 25 mm thick with machined surface

Plate welded on 45 mm box section (or something similar) frame

Plate width = 520 mm
Plate length = 880 mm
Figure A.6: Actual Machine
Appendix B

Hardware System

The actual hardware system is presented by photos in this appendix. It will be shown in detail what the main components look like, such as TI DSP module, SEMI-CRON’s SKHI 22A IGBT inverter kit and all connection of the experimental system.

Figure B.1: TI TMS320F28335 DSP Board
Figure B.2: DSP I/O Interface Module
Figure B.3: BDFRM with torque transducer mounted on the shaft

Figure B.4: The overall photos of the hardware system
Appendix C

Publications

   Artificial Intelligence, Modelling and Simulation (AIMS), 2015 IEEE International Conference on.


   Industrial Electronics (ISIE), 2013 IEEE International Symposium on, pp. 1-5.